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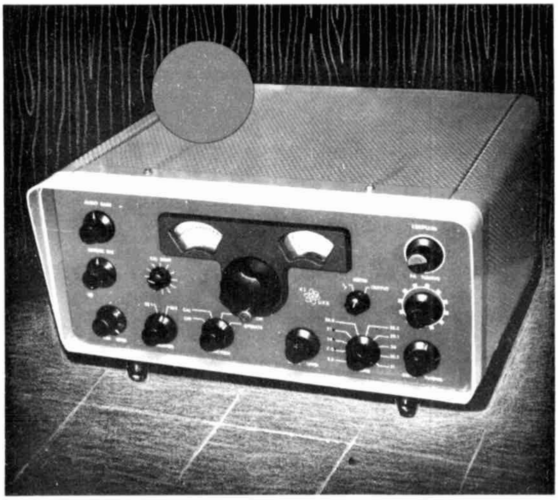
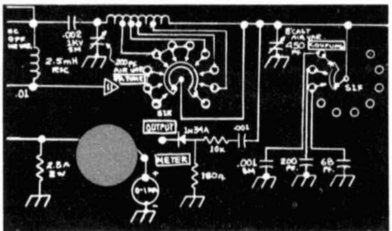
FEBRUARY 1968





this month

special section

- **homebrew 5-band ssb exciter**

complete construction details

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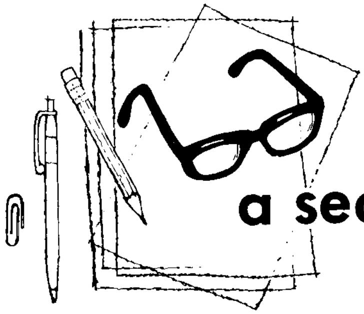


a second look

by jim

The birth of a magazine is a complicated process. It starts as an idea and develops through the coordinated efforts of authors, artists, typographers and pressmen. The ingredients are varied: articles, schematics, long-distance telephone calls, photography, advertising, subscriptions, public relations, paper and ink, all lovingly stirred together with prodigious amounts of midnight oil. Name: **ham radio,** born January, 1968.

Why **ham radio?** Very simple. The electronics and communications industry is moving forward at a tremendous clip, and so is amateur radio. Single sideband has largely replaced a-m, transistors are taking the place of vacuum tubes, and integrated circuits are finding their way into the ham workshop. The problem today, as it has always been, is to keep the amateur well informed. This doesn't mean that you have to impress your readers by printing every piece of state-of-the-art news that appears on the horizon.

Quite to the contrary. If you limit yourself to the state of the art, you'll get lost in a hurry. Advances are being made so rapidly, you just get tuned in and there's something new to worry about. And, since each new advance is built on what has been done in the past, if you don't get a clear idea of present techniques, you'll be hopelessly lost as time goes on.

Transistors have been with us for fifteen years, but it has just been within the last year or so that hams have started really thinking in terms of solid state. True, you usually wait for that old vacuum-tube equipment to wear out before replacing it with solid-state gear, but many of the little gizmos around the shack that you've put together in the past few years should be transistorized. Are they? Probably not.

When vacuum tubes became practical devices, amateurs were among the first to use them in home construction. Likewise with semiconductor diodes. Not so with Transistors. Why? Because hams didn't have enough good **practical** information to go by. Unfortunately, transistors blow out a lot easier than tubes, and many amateurs who experimented with them in the early days were disillusioned by a row of "dead soldiers" on their bench.

What it amounts to is this. There is an awful lot of state-of-the-art practice which you've got to understand before you get into state-of-the-art design. If you're an engineer, fine, but not all hams are engineers. Most hams want practical circuits which they can adapt to their special jobs.

As an example of where this state-of-the-art phobia has taken us, consider single sideband for a moment. I would guess that at least 80% of the hams operating on our lower bands are using single-sideband equipment. But how many of them actually understand what goes on inside that box? Not nearly enough!

The problem here is that ssb appears to be a lot more complex than a-m. Certainly the gear required to generate a ssb signal is more complex than that old a-m rig, but understanding what makes it tick is not. The two modes are very closely related. Not particularly compatible, but related. The problem is that a simple, concise explanation of sideband has been lacking. Several excellent articles along this line were written in the early fifties, but many of the hams who have sideband equipment now have never seen them.

Ham radio is designed to fill this gap. It is designed to inform. It will be geared to the state of the art—the state of the art in practice. It will be a magazine which shows you how to use new devices and old. Although we will encourage the use of solid state, we will not discriminate against vacuum tubes for the sake of being modern. There are a lot of places where vacuum tubes are still very practical and desirable gadgets.

When new techniques and devices become available, look for a complete description of how to use them in **ham radio.** Our articles will run the gamut from the simple to the complex, but they will all be oriented to the practical approach—the amateur approach. Not all of you are interested in the same things, nor do you have similar electronic backgrounds, so the fare served up in **ham radio** will be varied.

We will have simple projects for the novice and the one-night-a-week experimenter, involved projects for the experienced ham who likes to work in his shop, practical design and theory articles for the fellow who wants to start from scratch, and the last word for the VHF'er, RTTY bug and ATV enthusiast.

Amateur radio, just by its nature, is a very diversified hobby. Each ham follows his own special interests, whether home construction, public service, DX, traffic or a multitude of others. If you don't see an article that covers your particular plane of interest, it's because no one has taken the time to write it. If you have a little gimcrack that you've just put together, and think others would be interested in what it can do, draw a schematic, take some pictures and write it up. You don't have to be a professional writer to get your name in print; most of ours are not.

In addition to full-length feature articles, we are in the market for shorties for the **ham notebook.** If you have found a new and better way of doing something in the shack, have a new construction wrinkle, or have some small gem of technical information to convey, send it in to **ham notebook.** This monthly feature will cover everything from Antennas to Zener diodes, construction and design—technical tips that are useful around the shack and shop. You'll get paid for your efforts, and the rest of our readers will benefit from your ingenuity.

If you're interested in contributing something to **ham radio,** write for our "Author's Guide." This handy little pamphlet shows how to put your story together, the essentials of clear writing, the abbreviations we use, and what we need in terms of schematics and photographs. It also outlines our rates and payment policy.

Payments for manuscripts are generous and immediate. If we like your article, you'll get a check with our letter of acceptance, usually within a week of when you put it in the mail. After we accept it, it won't sit in our files for months or years waiting for publication. It will be put into print just as soon as possible. Articles that sit in the file are no good to the reader, to the author or to us. Nobody wants to read about (or build) a VHF converter today that was the hottest thing on two wheels when it was originally designed three years ago!

Ham radio will not stand still. We will always be looking for ways to improve because amateur radio is a dynamic hobby, always on the move. As the equipment, techniques and challenges of amateur radio change, so will we. We'll constantly try to make **ham radio** more useful to you as well as more interesting and stimulating. We can promise you now, we'll never become complacent, we'll always try to make **ham radio** better.

Jim Fisk W1DTY
Editor

here we are

a word from the publisher

Here we are! It's been a long trip since early October when Jim and I decided that there was definitely a place for **ham radio.** Where will the money come from? What will we call it? What will it look like? How will we get subscribers? How will we obtain advertisers? All of these questions, and many more, had to be answered one by one as **ham radio** began to take the final shape that you see here.

It has not been an easy road. We did not expect it to be. However, there have been many unexpected rewards along the way as many folks, both individual amateurs and those in industry, have given us valuable support and encouragement. We quickly found that there were many others who felt just as strongly as we that our magazine would be quite useful to the amateur community.

Can amateur radio really support a new magazine? We think it can, and here are some reasons why: Much has been said in recent years about what is wrong with our hobby, but altogether too little about what is healthy and right about it. We feel that amateur radio is a healthy patient and that it is going to be with us for a long time to come. Let's face it, we wouldn't be investing our time and money in this project if this were not the case.

In spite of what others may say, you need only look at the Call Book to see that the total number of licensees is continuing to grow. Admittedly, this growth may not be as fast as CB or other parts of the electronics industry, but still, we are growing. This growth could be, and should be, faster, but it is up to us in amateur radio to provide growth. No one else is going to do it for us.

There was, perhaps, a certain period of indecision during the incentive-licensing controversy. This issue has been decided, and the amateur knows exactly where he stands in the future. Home construction ideas and equipment purchases can now be planned with the full assurance that you'll be able to use this gear as intended. The potential newcomer need have no apprehension over what his future operating privileges will be.

One of our major enemies is considered to be Citizens Band, and yet, it has been a failure. Even Chairman Hyde of the FCC has now suggested that major changes, including an increase in the license fee and a technical examination, should be considered as a means of improving this service. This could well have a very positive effect on amateur radio. The many youngsters who are introduced to radio each year via CB might well be channelled to our novice bands. Both radio services would benefit from this change in policy, as would the individuals involved.

A new outlook is necessary. If we continue to work with old ideas and concepts, we can hardly expect to maintain our traditional spot in the electronics world. We are a branch of one of the fastest moving areas of technology. If you have any doubts, look at the developments of the past few years in solid-state techniques or satellite communications. Amateur radio will have to look and act the part if we are to keep up.

Ham radio intends to take a very positive step in the advancement of amateur radio. As a start, this magazine looks years ahead of others because we have taken advantage of new techniques and ideas in technical publishing to bring you something very timely in appearance. Equally as much effort is going into the preparation of editorial material. A lot of work has been done to make the articles easier to read. Ideas have been more clearly organized, words and sentences have been carefully put together with professional techniques designed to make **ham radio** easier and more enjoyable to read!

We still have a lot to do. Our business is communications technology. This goes just as much for the printed word as for the latest integrated circuit. There will be many more new ideas in **ham radio** as the months go by. We think you'll like them.

Skip Tenney W1NLB
Publisher

homebrew 5-band ssb exciter

Here's an all-band sideband exciter that's simple, straight-forward, inexpensive and easy to build.

Fred Randall K1UKX, RFD 2, Blackstone Street, Uxbridge, Massachusetts 01569

Many opinions, both pro and con, have been expressed in regards to homebrew equipment. It is not my intent either to promote or discourage building equipment. I believe the only valid reason for a ham to build his station equipment is his own sincere desire.

Many of those who profess a desire to build their own ssb gear claim it is difficult or impossible because:

1. It is so complex that it takes an engineer to build it.
2. It is difficult to get parts; nobody stocks them, and they are too expensive.
3. A well equipped machine shop is necessary for the mechanical work.
4. Homebrew equipment looks shoddy or makeshift and is, usually, much larger and more bulky than commercial equipment.
5. No time.

The exciter shown on these pages is my answer to those who use any of the standard excuses. It was built in its entirety on a hamshack workbench, using common hand tools and readily available, inexpensive parts. I can appreciate the time problem since I get home at 1:30 AM—this is when I can steal an hour or so to work on my little projects.

Building equipment is no bed of roses, and the subsequent debugging would try the patience of a saint. However, when the work is done and the bugs are exterminated, the satisfaction more than compensates for the barked knuckles, frayed nerves and gallons of coffee and midnight oil.

circuit description

A block diagram of the homebrew 5-band ssb exciter is shown in fig. 1. In this circuit a 6CW4 crystal oscillator generates a signal at either 453.5 or 456.5 kHz, depending upon which sideband is desired. This signal is applied to a germanium-diode ring modulator. One half of a 12AX7A is used as the speech amplifier. The other half is connected as a cathode follower to present the proper low-impedance audio signal to the ring modulator. The output of the modulator is a dsb suppressed carrier signal, which is transformer-coupled to the mechanical filter. The filter passes the chosen sideband and, for all practical purposes, eliminates the unwanted one.

A 455-kHz i-f amplifier follows the filter; the output of this amplifier is fed to a 12AT7 balanced mixer. The signal from the VFO, a Colpitts oscillator using a 6BA6, is also applied to this mixer. The VFO tunes from 3955 kHz to 4455 kHz—a 500-kHz tuning range. This signal is nulled out in the balanced mixer.

The output of the 12AT7 mixer is applied to another i-f amplifier. The input and output of this amplifier are gang-tuned by additional sections of the VFO tuning capacitor to the difference frequency (3500 to 4000 kHz). This tracking amplifier further eliminates the possibility of any VFO signal appearing at the output.

A 6AU8A is used as a converter to heterodyne the 80-meter signal up to other bands. For 80-meter operation, the plate circuit of the 6AU8A pentode section is provided with a load resistor and operates as a low-gain amplifier. For 40- through 10-meter operation a signal from a 6CB6 crystal oscillator is applied to the triode section, and the plate circuit of the pentode section is tuned to the

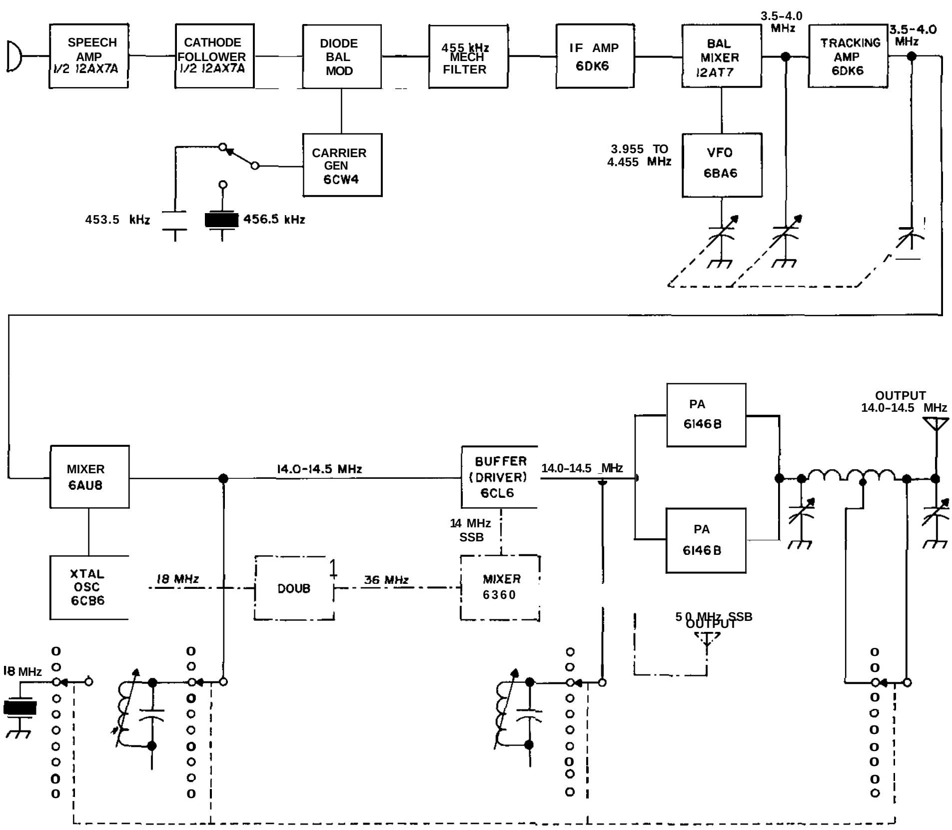


fig. 1. Block diagram of the homebrew five-band ssb exciter with the bandswitch set to 20 meters. The components shown by the dotted lines may be added to provide six-meter capability if so desired, although the circuitry is not described here.

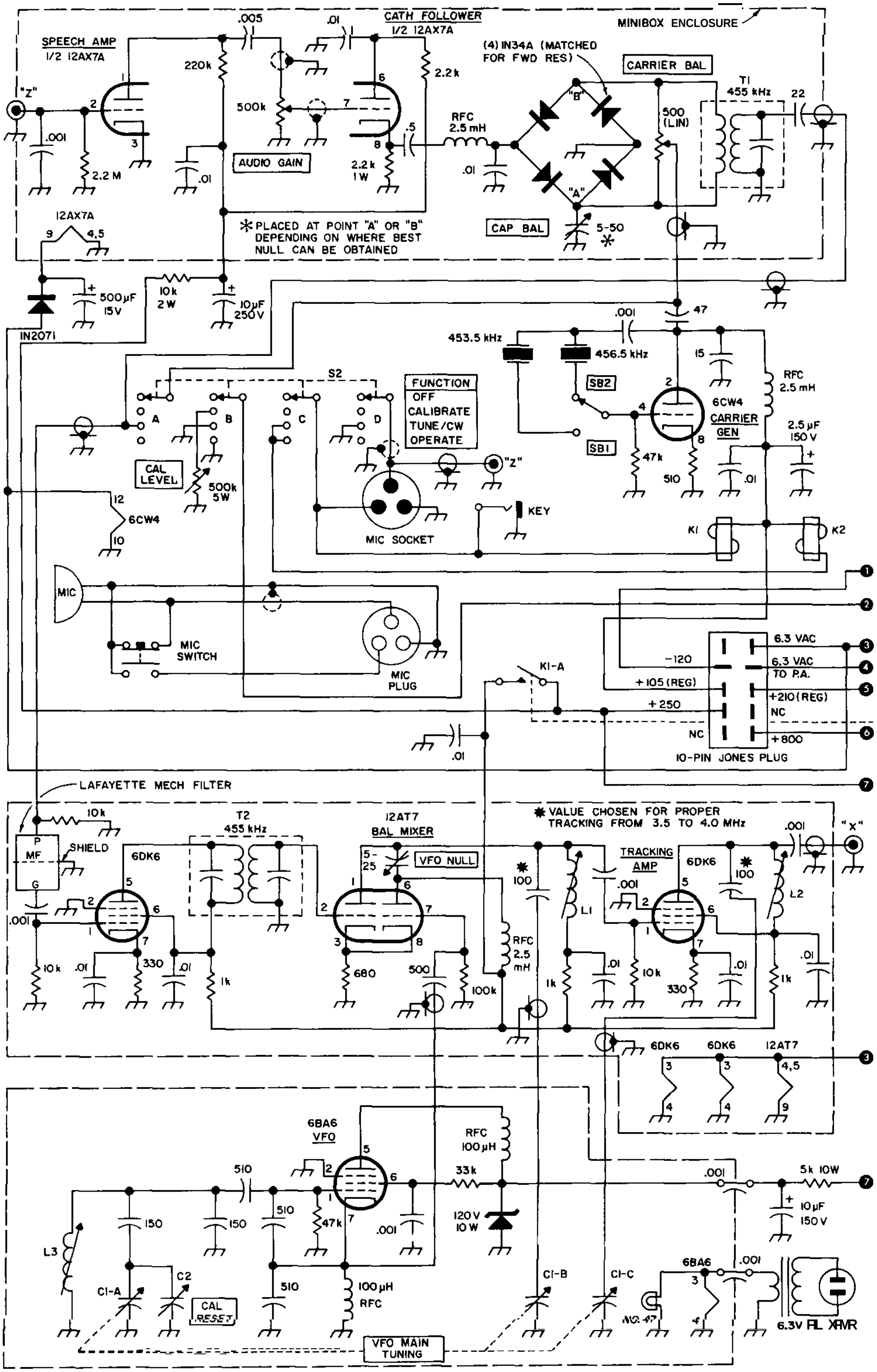
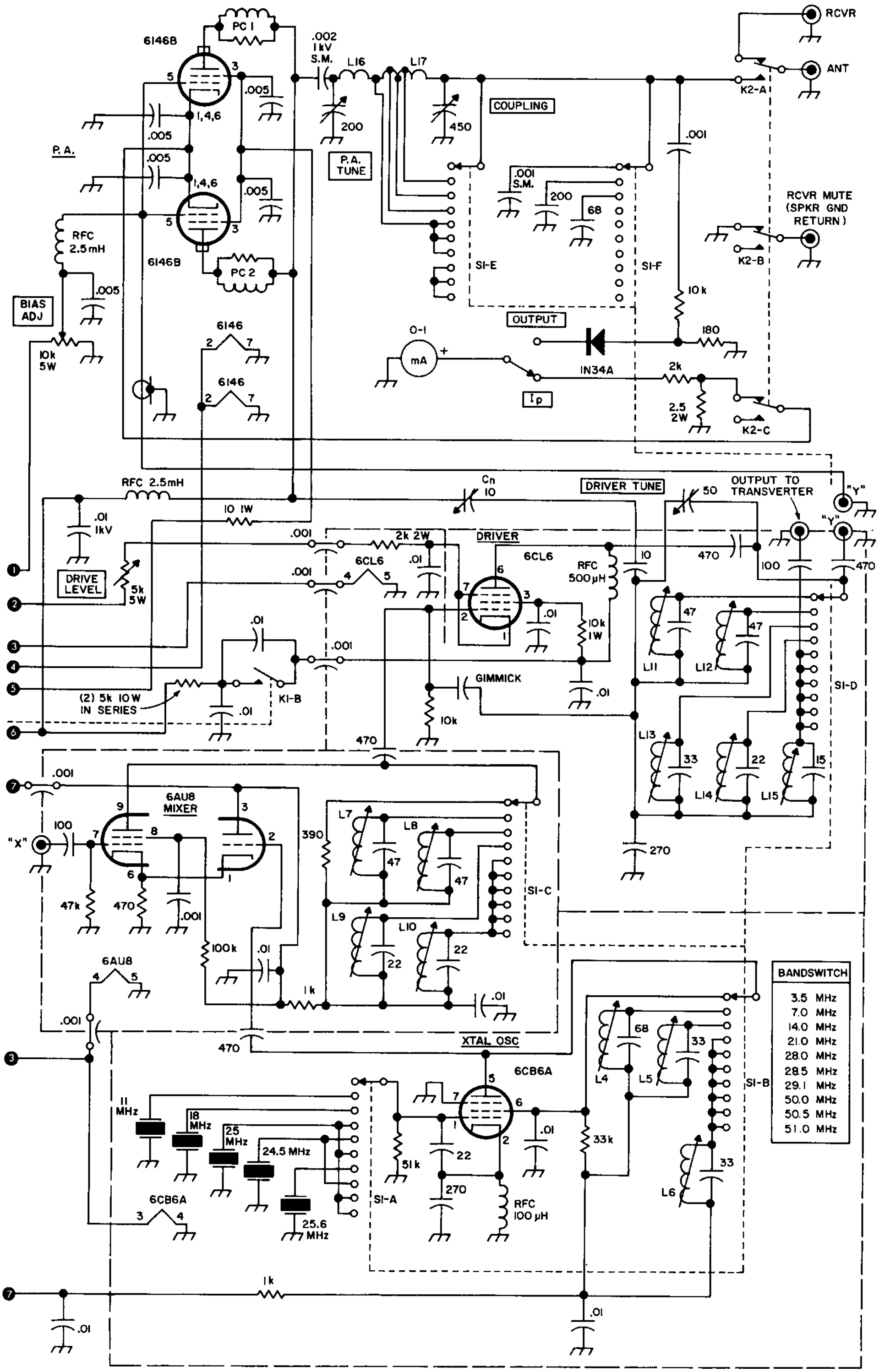


fig. 2. Schematic diagram of the low-cost five-band ssb exciter. Switch S2 (function) is in the tune-CW position,



switch S1 **(band)** at 3.5 MHz. Relays K1 and K2 are deenergized. The front-panel labeling is shown in boxes desired band. No direct coupling is used between the tube sections; sufficient mixing action is provided by inter-electrode capacity and by having the cathodes strapped together. The 6AU8A mixed is followed by a conventional class-A buffer amplifier and a bandswitched final using class-AB_1 6146B's.

The rule I use in anything I build is to use what I have available, rather than to try and locate and purchase what I don't have on hand.

This general rule was followed in building this exciter. The crystals and mechanical filter had to be purchased of course, but they were available "off the shelf" at Lafayette.* The 455-kHz i-f cans are available at any radio store or they may be removed from an old broadcast radio.

All of the circuits used in this exciter were obtained from various sources and adapted for use with components I had. It may be possible to further optimize the circuits, but the values I used work very well.

The carrier crystals and heterodyning crystals as well as the VFO frequency match the Lafayette HA-350 Receiver in use at K1UKX. For this reason, transceive operation is quite practical, with only slight modifications to the receiver.

* Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L. I., New York 11791. Order replacement part for HA-350 recetver, ***$12.95.***

Because both summing and differencing heterodyning is used to obtain the various bands, the tuning direction, as well as the selected sideband, changes with the band in use. This is a small inconvenience which you soon become accustomed to, and allows simpler and less expensive construction. The chart in **table 1** shows the tuning direction and normal sideband in use on the 80- through 10-meter amateur bands.

chassis and cabinet

A good starting point is the cabinet. I used an LMB* CO-1; price $20.00. One point which bears mentioning at this time is that when you use a decent appearing foundation for construction projects, it goes a long way toward encouraging the builder to complete them.

The dial bezel is cut from 1/4-inch thick stock. I used plexiglass, but wood would be a perfectly acceptable substitute. After cutting and filing it to shape, I sprayed it with machine-gray Krylon. While still tacky, the bezel was given a "dusting" coat of the same paint from about three feet away. This gives an attractive sandblast matte finish which blends well with the cabinet. The dial windows are cut from 1/8-inch plexiglass and

* Available from Newark Electronics Corporation, 500 North Pulaski, Chicago, Illinois ***60624.*** Catalog number 91F1192, $19.95 plus postage.

fig. 3. Construction of the chassis and tront panel The chassis is cut down from a Bud AC419 to 12 1/4" deep. When the 3/8" lip is formed around the front, the overall depth is twelve inches. A four-inch slot, 1/4" wide, is cut out along the front of the chassis for dial clearance and lead routing. The completed chassis is mounted to the 1/8-thick stiffener plate (sub-panel with countersunk screws.

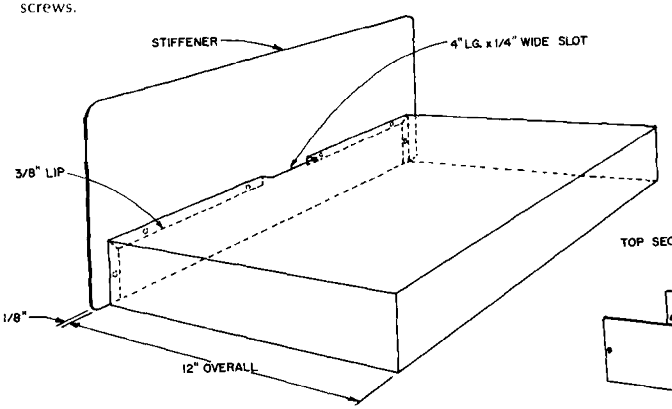
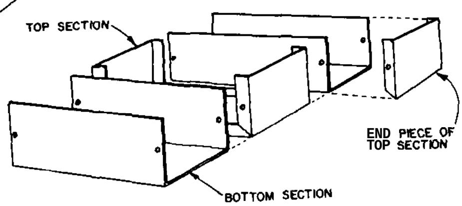


fig. 4. Method used for making shielded compartments for the bandswitch using two 1 1/2" x ***2" x 4"*** miniboxes. The circuits are built into each compartment, tested, and then assembled into a complete unit.



cemented to the rear of the bezel. A hairline is scribed in the center of the left-hand window.

The chassis consists of a front sub-panel cut from rack panel material and a commercial chassis, Bud AC419. The chassis is cut to size with a hacksaw or nibbling tool, bent to shape and bolted to the front panel using recessed-head screws. The use of a double panel permits a face plate completely free of unsightly screw heads. It also provides the rigidity necessary for mechanical stability of the VFO. The VFO compartment side plates are also cut from rack-panel material. These plates are bolted to the chassis and front panel after construction of the VFO. The rear panei of the cabinet is cut away to provide access to the chassis apron.

The front panel lettering is done with Walsco decals. After they are in place, the panel is given a coat of clear Krylon. A light dusting coat is then applied to return the panel to a semi-gloss finish. This seals the decals and completely hides their edges, giving the appearance of silk-screen lettering.

speech amplifier and balanced modulator

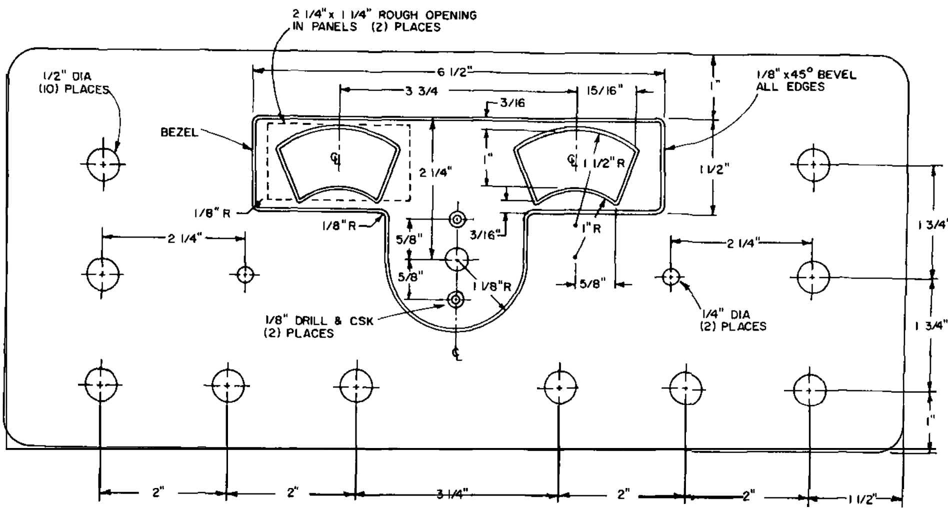
The speech-amplifier and balanced-modulator circuitry is built as a unit in a small mini-

fig. 5. Front finish panel for the ftve-band SSB transmitter. A full-scale template is available from K1UKX for 25c to cover postage and handling. Lips arc formed at the top and bottom of panel as shown in fig **6.**

table **1.** Parts list **for** exciter

C1 — ARC-5 receiver tuning capacitor
C2 — **ARC-5** receiver antenna-trim capacitor
L1, L2, L3—40 turns **#26** enameled on 1/2" slug-tuned form.
L4—12 turns #22 enameled on 3/8" slug-tuned form.
L5 — 10 turns #22 enameled on 3/8" slug-tuned form.
L6, L9 — 8 turns #22 enameled on 3/8" slug-tuned form.
L7 — 22 turns #22 enameled on 3/8" slug-tuned form.
L8, L13 — 15 turns #22 enameled on 3/8" slug-tuned form.
L10 — **6** turns #22 enameled on 3/8" slug-tuned form.
L11 — 65 turns #30 enameled on 3/8" slug-tuned form.
L12 — 18 turns #22 enameled on 3/8" slug-tuned form.
L14 — 9 turns #22 enameled on 3/8" slug-tuned form.
L15 — 4 turns #22 enameled on 3/8" slug-tuned form.
L16, L17 — Pi-Dux 820-D-10 with 4 turns removed from the 10 turns-per-inch end; replaced with **5** turns B&W **3006** mounted at right angles as shown in photograph. Tapped at the junction of the Pi-Dux unit and B&W 3006 and at **34,** 41, and **44** turns from the coupling-capacitor end.
K1, K2 — 4PDT, 15 kilohms, 110 Vdc. (Allied Control T163X-25)
PC1, PC2 — **6** turns 1-16" thick, 1/8" wide copper strip wound around 39-ohm, 2-watt resistors.
S1, S2 — Built up from Centralab index section and wafers.
T1 — 455 **kHz** i-f transformer (Workman TF11 although others will work) with primary winding and tuning capacitor removed and replaced with **50** turns #32 enameled, scramble wound next to secondary.
T2 — **455** kHz i-f transformer (Workman TF11).

box. The balance potentiometer is located in front of the minibox with an extension shaft through to the front panel. The audio-gain control is located on a bracket attached to the minibox by the nut that retains the



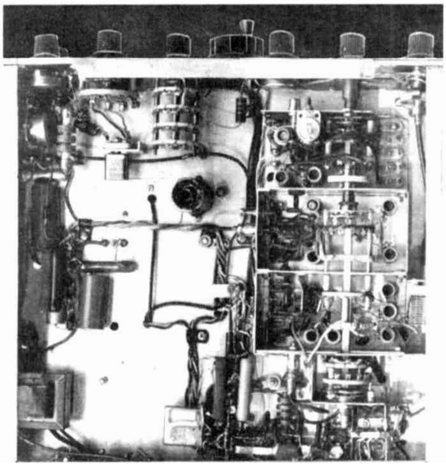
balance pot. Wiring to the gain control is accomplished with small shielded cable; power and signal leads come out through a hole in the bottom. It might be well to mention here that this method of construction not only provides excellent shielding, it makes for more pleasant building — you don't have to horse a big chassis all over the bench during construction. It also allows individual testing of sub-assemblies, a procedure which I highly recommend because it's much easier than aligning the whole exciter at one whack.

carrier generator

The 6CW4 carrier generator tube is located in the small space between the minibox and the front panel on the left side of the chassis as shown in **fig. 6**.

filter-mixer i-f amplifier

These stages are built into another minibox. Small sheet-metal protrusions provide room for the use of full-size 1/2-inch tuned coils on the input and output circuits of the tracking amplifier. All the leads from the box, as well as the coil leads to the ARC-5 tuning capacitor, are shielded. Access to the balanced-mixer nulling capacitor is available from the bottom of the main chassis through a small access hole. Proper tracking is established by juggling capacitor values in series with the ARC-5 variable capacitor sections. I trimmed some plates from my unit, but by proper choice of series capacitors, this should be unnecessary.

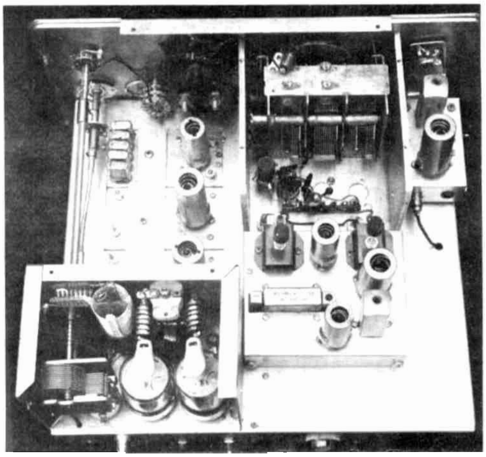


Below-chassis view of the five-band ssb exciter. The VFO tube and carrier generator crystals are in the upper left center; the bandswitch is to the right.

photos by Ted Woolner WA1ABP

variable frequency oscillator

The VFO is a standard Colpitts oscillator using a 6BA6 tube. The tube is located under the chassis and all circuitry associated with the VFO is contained in a shielded compartment. The 6BA6 is provided with a separate filament transformer and the heater runs as long as the exciter is plugged in. A number 47 bulb, which is used to keep the VFO compartment warm, is also connected. These provisions make warm-up drift negligible and no temperature-compensation circuitry is required.



Inside the five-band ssb exciter. Power amplifier compartment to the lower left.

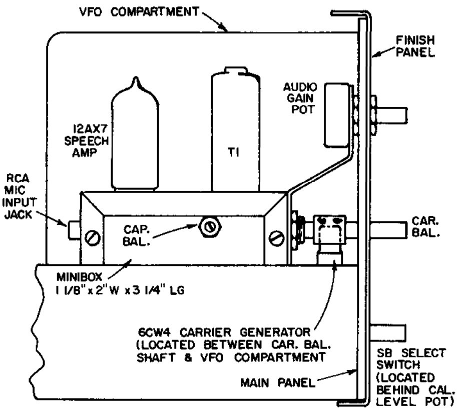
Plate and filament voltage to the VFO is run through feed-through capacitors from under the chassis. The VFO plate voltage is zener regulated at 120 volts. A small 10-pF variable capacitor is connected in parallel with the main tuning capacitor and provides about ±10 kHz variation for tuning-dial calibration. This capacitor was removed from the same ARC-5 receiver that yielded the main-tuning capacitor. It is placed between the tuning capacitor and the face plate; this is a tight squeeze but it fits after careful positioning.

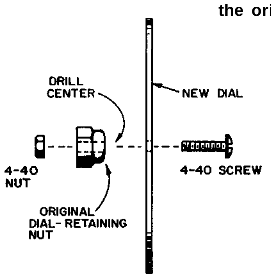
The center section of the main-tuning capacitor is used for the VFO. The remaining sections are used to tune the tracking amplifier. Note that a capacitor is placed in series with the main-tuning capacitor—this limits the tuning range to 500 kHz, and makes the dial calibration extremely linear.

final amplifier

The final amplifier is a conventional neutralized circuit using a pi-network output. The two 6146B tubes are connected in parallel and are provided with regulated grid bias and screen voltage. The cathode pins on the tube sockets are bent inward toward the center and strapped together with a tinned-copper strip. Flat copper strips are also used for the plate parasitic suppressors, and many of the connections in the bandswitch assembly make use of the same material.

fig. 6. Side view of the chassis showing the position of the mechanical filter compartment on the left-hand side of the chassis. Note the lips at the top and bottom of the front finish panel.





control functions

Most of the control functions are either self-explanatory or discussed elsewhere in the text. A few functions, perhaps, require further explanation.

The calibrate position of the function switch permits the operator to "talk" himself on frequency, or, if desired, the carrier may be inserted and a conventional zero beat can be accomplished. In either case, the push-to-talk switch (or key) must be depressed to energize the transmitter. The calibrate level control sets the spotting signal to a comfortable listening level.

In the tune-CW position of the function switch, the balanced modulator is bypassed to provide a CW carrier when the key or push-to-talk switch is depressed. Full break-in CW permits maximum operating convenience. This is accomplished through the use of high-speed relays.

The meter is switched to read either final cathode current or output. Antenna change-over and receiver muting are built in, as is push to talk. Any power supply that will provide the indicated voltages at the power-plug terminals may be used.

No tuning or alignment instructions are given here since it is assumed that anyone with the necessary skills to build an exciter such as this one will be thoroughly familiar with the proper techniques.

It is not expected that anyone will build an exact copy of this transmitter. However, some of the construction methods and circuits used here should be of interest to other appliance operators who prefer, as I do, to build their own appliances.

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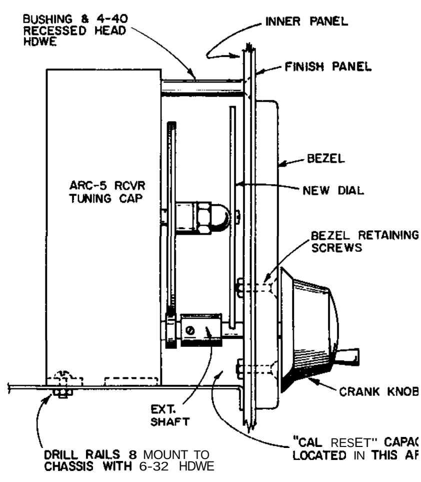
fig. 7. Construction of the driving mechanism for the ARC-5 tuning capacitor. The dial scale is mounted to the original dial retaining nut with a 4-40 screw.



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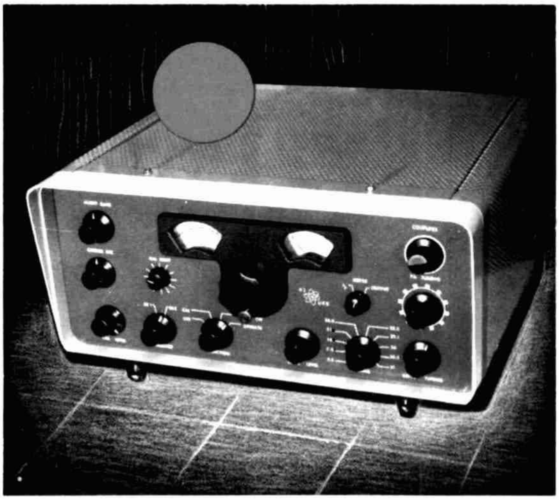

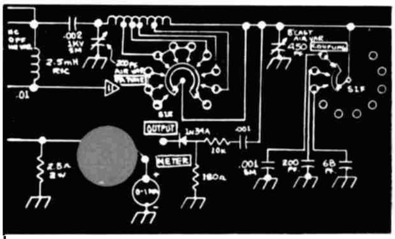

this month

special section

- **homebrew 5-band SSB exciter**

complete construction details

first issue

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Receiver Incremental Tuning	Yes	No	No
Built-in notch Filter	Yes	No	No
Sharp CW Filter	Yes 200 cycles	No	[illegible]
Sensitivity	.3 uv for 10 db S/N	.5 uv for 10 db S/N	.5 uv for 10 db S/N
1 kHz dial readout	Yes	Yes	
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ham radio

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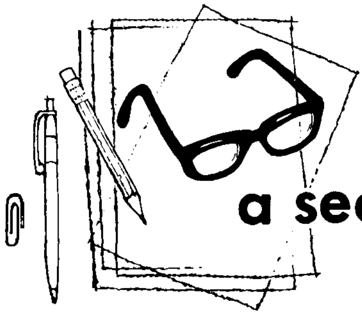

a second look

by jim fisk

The birth of a magazine is a complicated process. It starts as an idea and develops through the coordinated efforts of authors, artists, typographers and pressmen. The ingredients are varied: articles, schematics, long-distance telephone calls, photography, advertising, subscriptions, public relations, paper and ink, all lovingly stirred together with prodigious amounts of midnight oil. Name: **ham radio,** born January, 1968.

Why **ham radio?** Very simple. The electronics and communications industry is moving forward at a tremendous clip, and so is amateur radio. Single sideband has largely replaced a-m, transistors are taking the place of vacuum tubes, and integrated circuits are finding their way into the ham workshop. The problem today, as it has always been, is to keep the amateur well informed. This doesn't mean that you have to impress your readers by printing every piece of state-of-the-art news that appears on the horizon.

Quite to the contrary. If you limit yourself to the state of the art, you'll get lost in a hurry. Advances are being made so rapidly, you just get tuned in and there's something new to worry about. And, since each new advance is built on what has been done in the past, if you don't get a clear idea of present techniques, you'll be hopelessly lost as time goes on.

Transistors have been with us for fifteen years, but it has just been within the last year or so that hams have started really thinking in terms of solid state. True, you usually wait for that old vacuum-tube equipment to wear out before replacing it with solid-state gear, but many of the little gizmos around the shack that you've put together in the past few years should be transistorized. Are they? Probably not.

When vacuum tubes became practical devices, amateurs were among the first to use them in home construction. Likewise with semiconductor diodes. Not so with Transistors. Why? Because hams didn't have enough good **practical** information to go by. Unfortunately, transistors blow out a lot easier than tubes, and many amateurs who experimented with them in the early days were disillusioned by a row of "dead soldiers" on their bench.

What it amounts to is this. There is an awful lot of state-of-the-art practice which you've got to understand before you get into state-of-the-art design. If you're an engineer, fine, but not all hams are engineers. Most hams want practical circuits which they can adapt to their special jobs.

As an example of where this state-of-the-art phobia has taken us, consider single sideband for a moment. I would guess that at least 80% of the hams operating on our lower bands are using single-sideband equipment. But how many of them actually understand what goes on inside that box? Not nearly enough!

The problem here is that ssb appears to be a lot more complex than a-m. Certainly the gear required to generate a ssb signal is more complex than that old a-m rig, but understanding what makes it tick is not. The two modes are very closely related. Not particularly compatible, but related. The problem is that a simple, concise explanation of sideband has been lacking. Several excellent articles along this line were written in the early fifties, but many of the hams who have sideband equipment now have never seen them.

Ham radio is designed to fill this gap. It is designed to inform. It will be geared to the state of the art—the state of the art in practice. It will be a magazine which shows you how to use new devices and old. Although we will encourage the use of solid state, we will not discriminate against vacuum tubes for the sake of being modern. There are a lot of places where vacuum tubes are still very practical and desirable gadgets.

When new techniques and devices become available, look for a complete description of how to use them in **ham radio.** Our articles will run the gamut from the simple to the complex, but they will all be oriented to the practical approach—the amateur approach. Not all of you are interested in the same things, nor do you have similar electronic backgrounds, so the fare served up in **ham radio** will be varied.

We will have simple projects for the novice and the one-night-a-week experimenter, involved projects for the experienced ham who likes to work in his shop, practical design and theory articles for the fellow who wants to start from scratch, and the last word for the VHF'er, RTTY bug and ATV enthusiast.

Amateur radio, just by its nature, is a very diversified hobby. Each ham follows his own special interests, whether home construction, public service, DX, traffic or a multitude of others. If you don't see an article that covers your particular plane of interest, it's because no one has taken the time to write it. If you have a little gimcrack that you've just put together, and think others would be interested in what it can do, draw a schematic, take some pictures and write it up. You don't have to be a professional writer to get your name in print; most of ours are not.

In addition to full-length feature articles, we are in the market for shorties for the **ham notebook.** If you have found a new and better way of doing something in the shack, have a new construction wrinkle, or have some small gem of technical information to convey, send it in to **ham notebook.** This monthly feature will cover everything from Antennas to Zener diodes, construction and design—technical tips that are useful around the shack and shop. You'll get paid for your efforts, and the rest of our readers will benefit from your ingenuity.

If you're interested in contributing something to **ham radio,** write for our "Author's Guide." This handy little pamphlet shows how to put your story together, the essentials of clear writing, the abbreviations we use, and what we need in terms of schematics and photographs. It also outlines our rates and payment policy.

Payments for manuscripts are generous and immediate. If we like your article, you'll get a check with our letter of acceptance, usually within a week of when you put it in the mail. After we accept it, it won't sit in our files for months or years waiting for publication. It will be put into print just as soon as possible. Articles that sit in the file are no good to the reader, to the author or to us. Nobody wants to read about (or build) a VHF converter today that was the hottest thing on two wheels when it was originally designed three years ago!

Ham radio will not stand still. We will always be looking for ways to improve because amateur radio is a dynamic hobby, always on the move. As the equipment, techniques and challenges of amateur radio change, so will we. We'll constantly try to make **ham radio** more useful to you as well as more interesting and stimulating. We can promise you now, we'll never become complacent, we'll always try to make **ham radio** better.

Jim Fisk W1DTY
Editor

Note: If you received more than one copy of this first issue of **ham radio,** don't worry about it–it simply means that you're an active amateur and your name has appeared on more than one of our mailing lists. Just give the extra copy to a friend who hasn't seen it.

here we are

a word from the publisher

Here we are! It's been a long trip since early October when Jim and I decided that there was definitely a place for **ham radio.** Where will the money come from? What will we call it? What will it look like? How will we get subscribers? How will we obtain advertisers? All of these questions, and many more, had to be answered one by one as **ham radio** began to take the final shape that you see here.

It has not been an easy road. We did not expect it to be. However, there have been many unexpected rewards along the way as many folks, both individual amateurs and those in industry, have given us valuable support and encouragement. We quickly found that there were many others who felt just as strongly as we that our magazine would be quite useful to the amateur community.

Can amateur radio really support a new magazine? We think it can, and here are some reasons why: Much has been said in recent years about what is wrong with our hobby, but altogether too little about what is healthy and right about it. We feel that amateur radio is a healthy patient and that it is going to be with us for a long time to come. Let's face it, we wouldn't be investing our time and money in this project if this were not the case.

In spite of what others may say, you need only look at the Call Book to see that the total number of licensees is continuing to grow. Admittedly, this growth may not be as fast as CB or other parts of the electronics industry, but still, we are growing. This growth could be, and should be, faster, but it is up to us in amateur radio to provide growth. No one else is going to do it for us.

There was, perhaps, a certain period of indecision during the incentive-licensing controversy. This issue has been decided, and the amateur knows exactly where he stands in the future. Home construction ideas and equipment purchases can now be planned with the full assurance that you'll be able to use this gear as intended. The potential newcomer need have no apprehension over what his future operating privileges will be.

One of our major enemies is considered to be Citizens Band, and yet, it has been a failure. Even Chairman Hyde of the FCC has now suggested that major changes, including an increase in the license fee and a technical examination, should be considered as a means of improving this service. This could well have a very positive effect on amateur radio. The many youngsters who are introduced to radio each year via CB might well be channelled to our novice bands. Both radio services would benefit from this change in policy, as would the individuals involved.

A new outlook is necessary. If we continue to work with old ideas and concepts, we can hardly expect to maintain our traditional spot in the electronics world. We are a branch of one of the fastest moving areas of technology. If you have any doubts, look at the developments of the past few years in solid-state techniques or satellite communications. Amateur radio will have to look and act the part if we are to keep up.

Ham radio intends to take a very positive step in the advancement of amateur radio. As a start, this magazine looks years ahead of others because we have taken advantage of new techniques and ideas in technical publishing to bring you something very timely in appearance. Equally as much effort is going into the preparation of editorial material. A lot of work has been done to make the articles easier to read. Ideas have been more clearly organized, words and sentences have been carefully put together with professional techniques designed to make **ham radio** easier and more enjoyable to readl

We still have a lot to do. Our business is communications technology. This goes just as much for the printed word as for the latest integrated circuit. There will be many more new ideas in **ham radio** as the months go by. We think you'll like them.

Skip Tenney W1NLB
Publisher

A question only serious hams should answer...

by Jack Quinn, W6MJG

How come you are still asking for our obsolete book? The one called "The Care and Feeding of Power Tetrodes." Look. we've already mailed out over 100.000 copies of the thing. It's just got to be in the hands of every amateur who ever went on the air. Don't get mc wrong. I'm happy you find it useful. But now you should be asking for our NEW book. "The Care and Feeding of Power Grid Tubes."

It so happens that right now on my desk is a pile of these new books. They're really pretty interesting. You see, one of the fellows on our staff—Bob Sutherland. W6UOV—took it upon himself to incorporate the answers to over 400 questions asked of us in a year's time. In fact. he has spent just about every spare moment away from his shack. preparing this new pocket-size book. I couldn't believe that it has almost 200 pages. Bob said he just got carried away. He has expanded the original book, which we published back in '46, so that in its new form it covers all types of power grid tubes in RF and AF service. Even has graphs and things like that.

Now you're probably wondering, where can I get it? Thought you'd never ask. Right this minute there is another pile of these books at your nearest Eimac/Varian distributor. or your favorite technical bookstore. Figuring all the time we've spent in getting them ready for you. they're really a bargain at 93.95 each. If it's inconvenient to get to the distributor or the bookstore, write me. and I'll send your request along to the book retailer.

Jack Quinn
Division Marketing Manager

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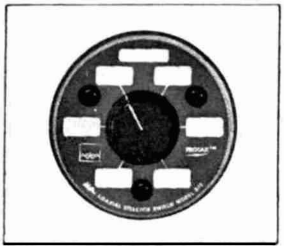

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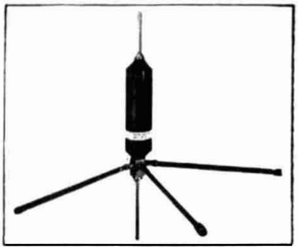

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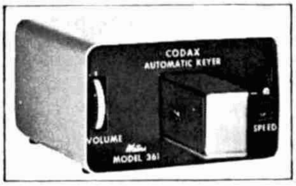

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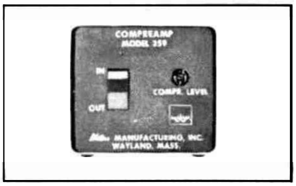

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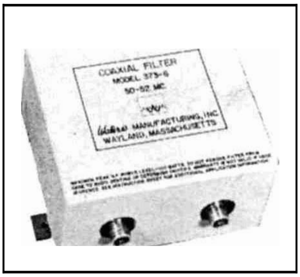

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homebrew 5-band ssb exciter

Here's an all-band sideband exciter that's simple, straight-forward, inexpensive and easy to build.

Randall K1UKX, FD 2, Blackstone St., Uxbridge, Massachusetts 569

Many **opinions,** both pro and con, have been expressed in regards to homebrew equipment. It is not my intent either to promote or discourage building equipment. I believe the only valid reason for a ham to build his station equipment is his own sincere desire.

Many of those who profess a desire to build their own ssb gear claim it is difficult or impossible because:

1. It is so complex that it takes an engineer to build it.

2. It is difficult to get parts; nobody stocks them, and they are too expensive.

3. A well equipped machine shop is necessary for the mechanical work.

4. Homebrew equipment looks shoddy or makeshift and is, usually, much larger and more bulky than commercial equipment.

5. No time.

The exciter shown on these pages is my answer to those who use any of the standard excuses. It was built in its entirety on a hamshack workbench, using common hand tools and readily available, inexpensive parts. I can appreciate the time problem since I get home at 1:30 AM—this is when I can steal an hour or so to work on my little projects.

Building equipment is no bed of roses, and the subsequent debugging would try the patience of a saint. However, when the work is done and the bugs are exterminated, the satisfaction more than compensates for the barked knuckles, frayed nerves and gallons of coffee and midnight oil.

circuit description

A block diagram of the homebrew 5-band ssb exciter is shown in **fig. 1**. In this circuit a 6CW4 crystal oscillator generates a signal at either 453.5 or 456.5 kHz, depending upon which sideband is desired. This signal is applied to a germanium-diode ring modulator. One half of a 12AX7A is used as the speech amplifier. The other half is connected as a cathode follower to present the proper low-impedance audio signal to the ring modulator. The output of the modulator is a dsb suppressed carrier signal, which is transformer-coupled to the mechanical filter. The filter passes the chosen sideband and, for all practical purposes, eliminates the unwanted one.

A 455-kHz i-f amplifier follows the filter; the output of this amplifier is fed to a 12AT7 balanced mixer. The signal from the VFO, a Colpitts oscillator using a 6BA6, is also applied to this mixer. The VFO tunes from 3955 kHz to 4455 kHz—a 500-kHz tuning range. This signal is nulled out in the balanced mixer.

The output of the 12AT7 mixer is applied to another i-f amplifier. The input and output of this amplifier are gang-tuned by additional sections of the VFO tuning capacitor to the **difference** frequency (3500 to 4000 kHz). This tracking amplifier further eliminates the possibility of any VFO signal appearing at the output.

A 6AU8A is used as a converter to heterodyne the 80-meter signal up to other bands. For 80-meter operation, the plate circuit of the 6AU8A pentode section is provided with a load resistor and operates as a low-gain amplifier. For 40- through 10-meter operation a signal from a 6CB6 crystal oscillator is applied to the triode section, and the plate circuit of the pentode section is tuned to the

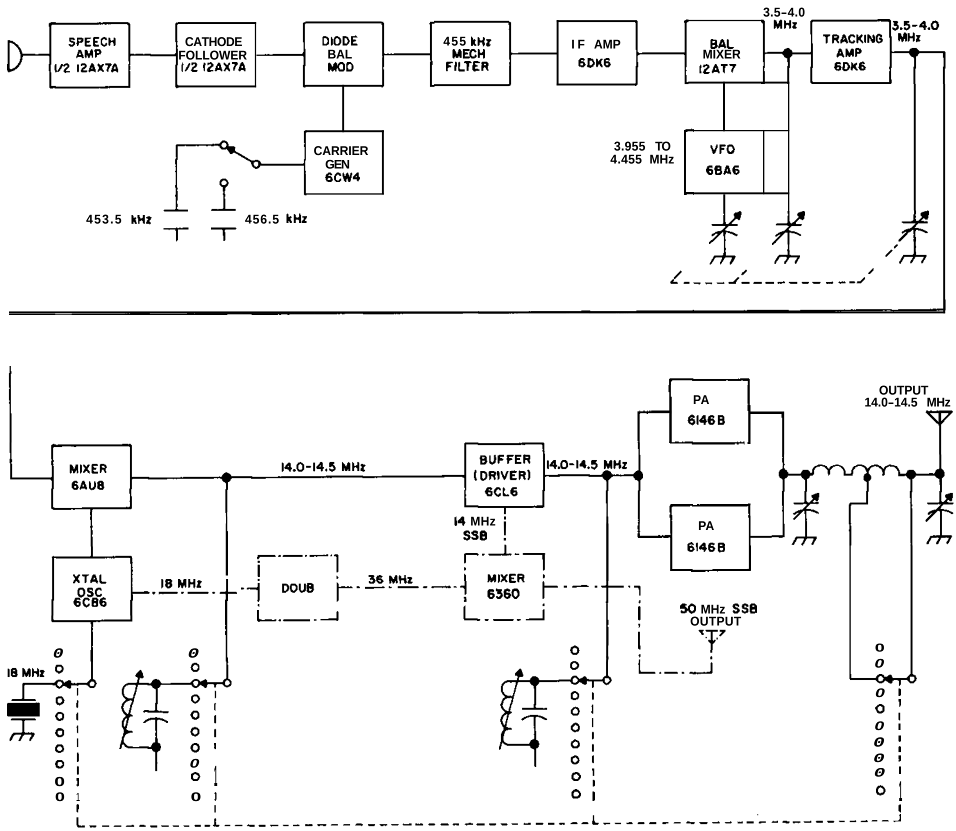


fig. 1. Block diagram of the homebrew five-band ssb exciter with the bandswitch set to 20 meters. The components shown by the dotted lines may be added to provide six-meter capability if so desired, although the circuitry is not described here.

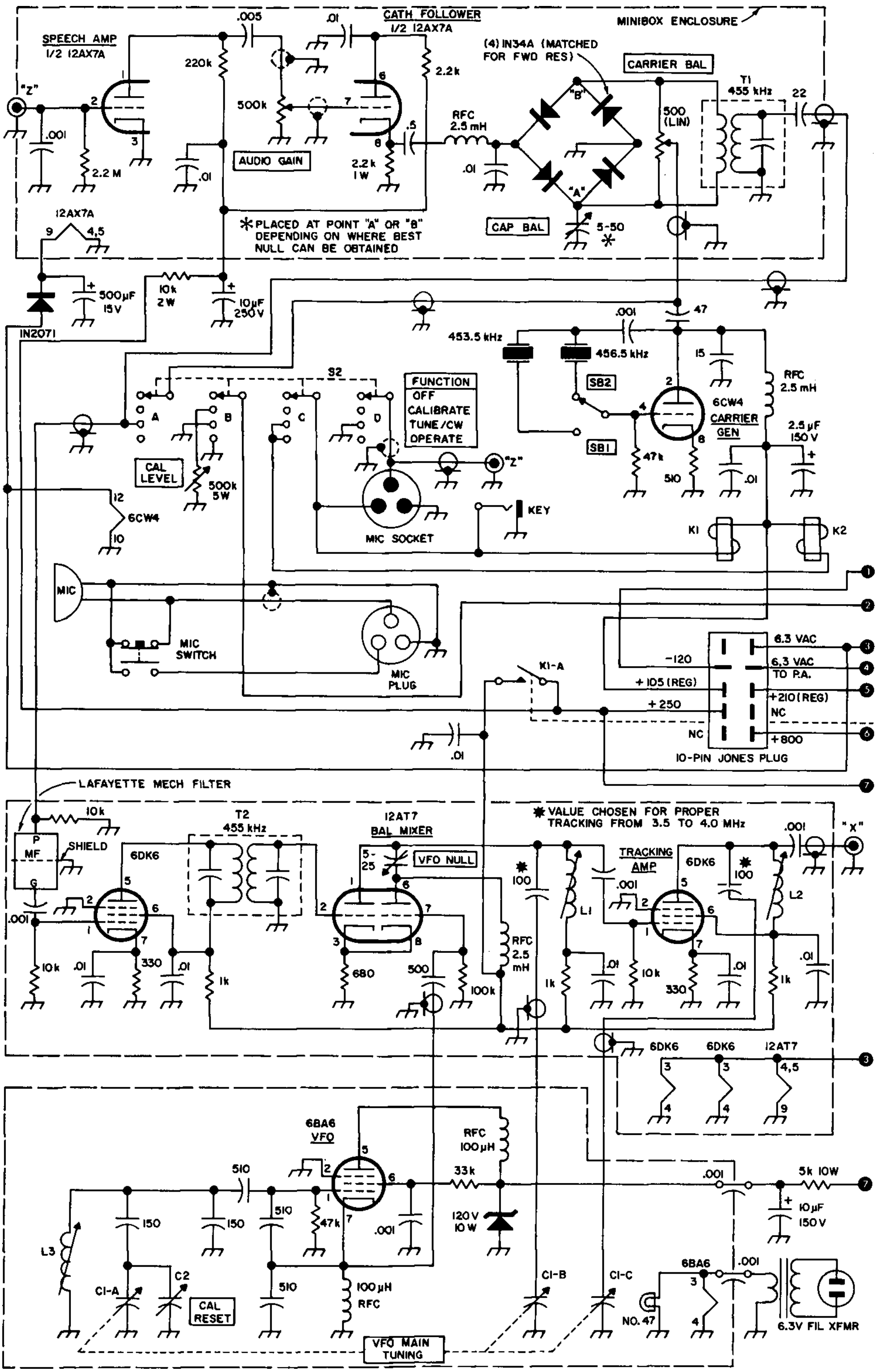
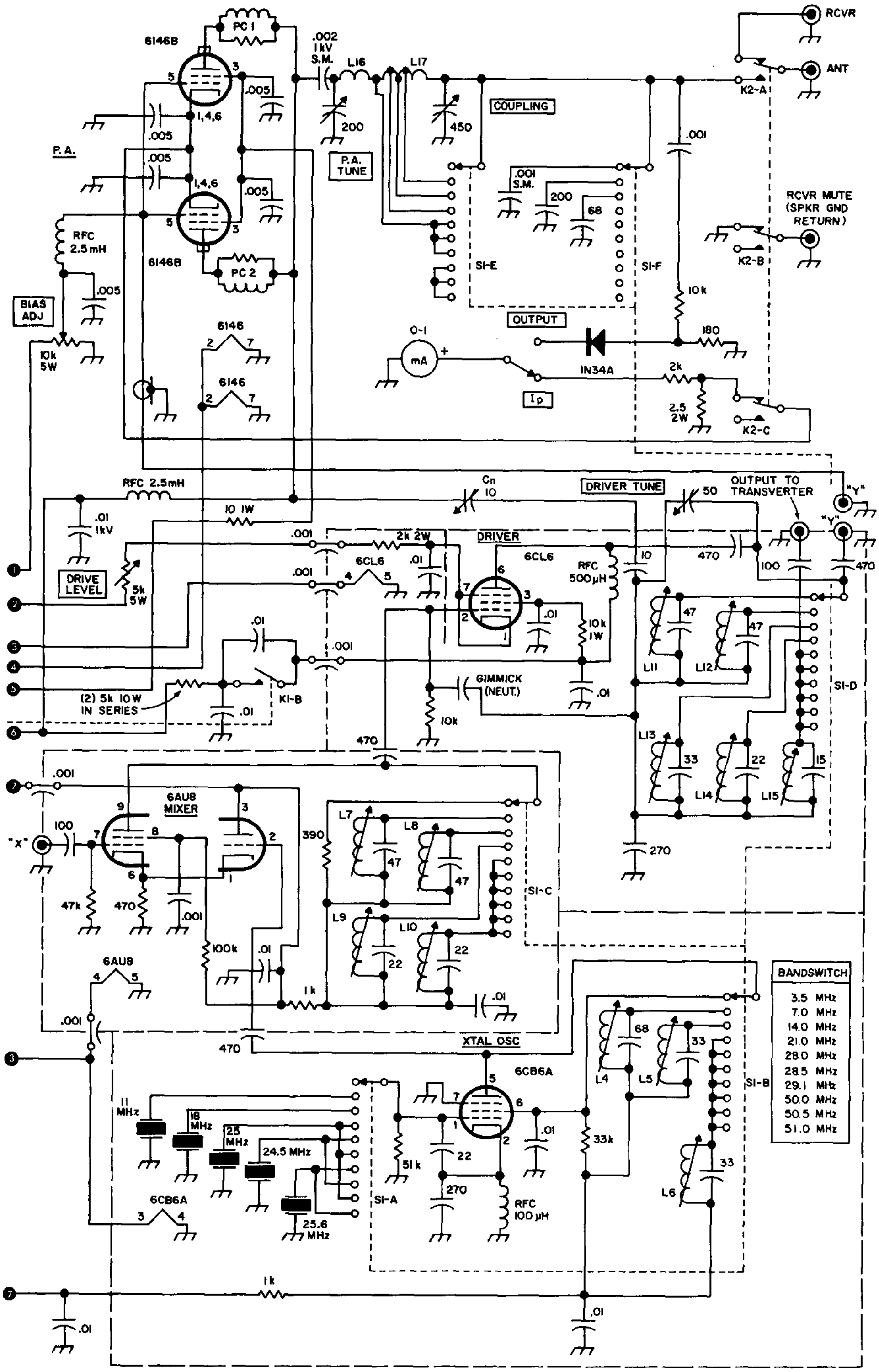


fig. 2. Schematic diagram of the low-cost five-band ssb exciter. Switch S2 **(function)** is in the tune-CW position,



switch S1 (band) at 3.5 MHz. Relays K1 and K2 are deenergized. The front-panel labeling is shown in boxes desired band. No direct coupling is used between the tube sections; sufficient mixing action is provided by inter-electrode capacity and by having the cathodes strapped together. The 6AU8A mixer is followed by a conventional class-A buffer amplifier and a bandswitched final using class-AB_1 6146B's.

The rule I use in anything I build is to use what I have available, rather than to try and locate and purchase what I don't have on hand.

This general rule was followed in building this exciter. The crystals and mechanical filter had to be purchased of course, but they were available "off the shelf" at Lafayette.* The 455-kHz i-f cans are available at any radio store or they may be removed from an old broadcast radio.

All of the circuits used in this exciter were obtained from various sources and adapted for use with components I had. It may be possible to further optimize the circuits, but the values I used work very well.

The carrier crystals and heterodyning crystals as well as the VFO frequency match the Lafayette HA-350 Receiver in use at K1UKX. For this reason, transceive operation is quite practical, with only slight modifications to the receiver.

Because both summing and differencing heterodyning is used to obtain the various bands, the tuning direction, as well as the selected sideband, changes with the band in use. This is a small inconvenience which you soon become accustomed to, and allows simpler and less expensive construction. The chart in **table 1** shows the tuning direction and normal sideband in use on the 80- through 10-meter amateur bands.

table 1

band	tuning direction	sideband selection SB-1	SB-2
3.5	normal	upper	lower
7.0	reversed	lower	upper
14.0	reversed	lower	upper
21.0	reversed	lower	upper
28.0	normal	upper	lower
28.5	normal	upper	lower
29.1	normal	upper	lower

chassis and cabinet

A good starting point is the cabinet. I used an LMB* CO-1; price $20.00. One point which bears mentioning at this time is that when you use a decent appearing foundation for construction projects, it goes a long way toward encouraging the builder to complete them.

The dial bezel is cut from '1s-inch thick stock. I used plexiglass, but wood would be a perfectly acceptable substitute. After cutting

fig. 3. Construction of the chassis and front panel. The chassis is cut down from a Bud AC419 to 12¼" deep. When the 3/8" lip is formed around the front, the overall depth is twelve inches. A four-inch slot, ¼" wide, is cut out along the front of the chassis for dial clearance and lead routing. The completed chassis is mounted to the 1/8-thick stiffener plate (sub-panel with countersunk screws.

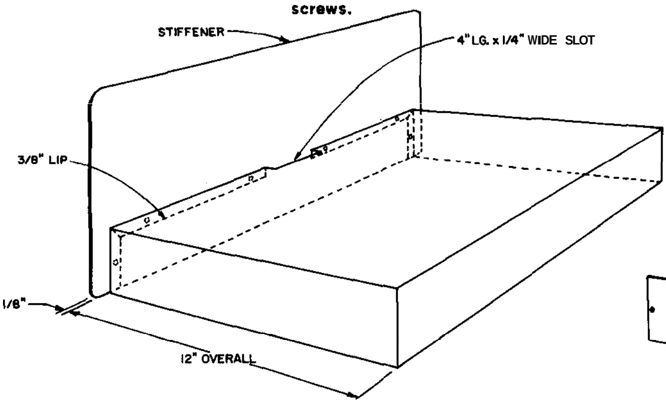
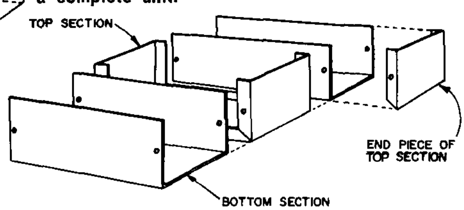


fig. 4. Method used for making shielded compartments for the bandswitch using two 1½" x 2" x 4" miniboxes. The circuits are built into each compartment, tested, and then assembled into a complete unit.



* Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L. I., New York 11791. Order replacement part for HA-350 receiver, $12.95.

* Available from Newark Electronics Corporation, 500 North Pulaski, Chicago, Illinois 60624. Catalog number 91F1192, $19.95 plus postage.

and filing it to shape, I sprayed it with machine-gray Krylon. While still tacky, the bezel was given a "dusting" coat of the same paint from about three feet away. This gives an attractive sandblast matte finish which blends well with the cabinet. The dial windows are cut from 'la-inch plexiglass and cemented to the rear of the bezel. A hairline is scribed in the center of the left-hand window.

The chassis consists of a front sub-panel cut from rack panel material and a commercial chassis, Bud AC419. The chassis is cut to size with a hacksaw or nibbling tool, bent to shape and bolted to the front panel using recessed-head screws. The use of a double panel permits a face plate completely free of unsightly screw heads. It also provides the rigidity necessary for mechanical stability of the VFO. The VFO compartment side plates are also cut from rack-panel material. These plates are bolted to the chassis and front panel after construction of the VFO. The rear panel of the cabinet is cut away to provide access to the chassis apron.

The front panel lettering is done with Walsco decals. After they are in place, the panel is given a coat of clear Krylon. A light dusting coat is then applied to return the panel to a semi-gloss finish. This seals the decals and completely hides their edges, giving the appearance of silk-screen lettering.

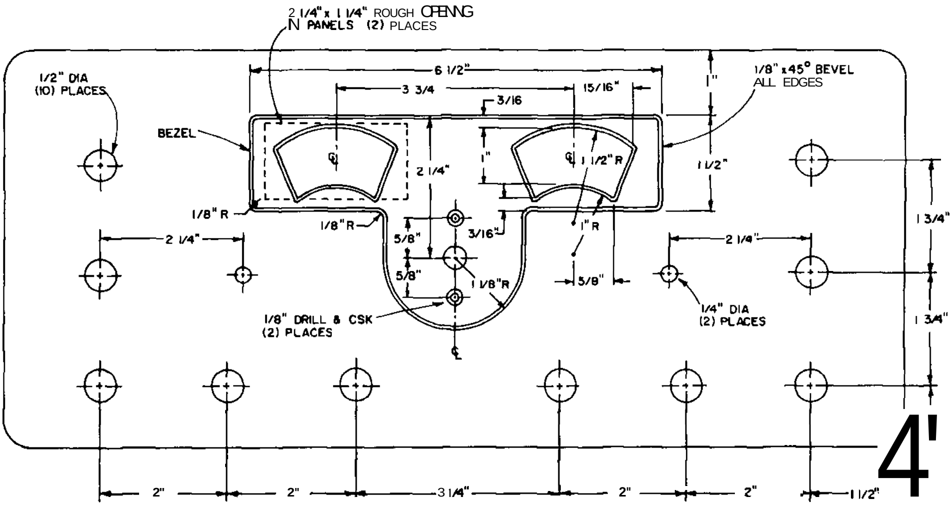
fig. **5.** Front finish panel for the **five-band** SSB transmitter. A full-scale template is available from **K1UKX** for **25c** to cover postage and handling. Lips are formed at the top and bottom of panel as shown in fig. 6.

table 2. Parts list for exciter

- **C1** — ARC-5 receiver tuning capacitor
- **C2** — ARC-5 receiver antenna-trim capacitor
- **L1, L2, L3**—40 turns #26 enameled on ½" slug-tuned form.
- **L4**—12 turns #22 enameled on ⅜" slug-tuned form.
- **L5** — 10 turns #2? enameled on ⅜" slug-tuned form.
- **L6, L9** — 8 turns #22 enameled on ⅜" slug-tuned form.
- L7 — 22 turns #22 enameled on ⅜" slug-tuned form.
- **L8,** L13 — 15 turns #22 enameled on ⅜" slug-tuned form.
- **L10** — 6 turns #22 enameled on ⅜" slug-tuned form.
- **L11** — 65 turns #30 enameled on ⅜" slug-tuned form.
- **L12** — 18 turns #22 enameled on ⅜" slug-tuned form.
- L14 — 9 turns #22 enameled on ⅜" slug-tuned form.
- **L15** — 4 turns #22 enameled on ⅜" slug-tuned form.
- L16, L17 — Pi-Dux 820-D-10 with 4 turns removed from the 10 turns-per-inch end; replaced with 5 turns B&W 3006 mounted at right angles as shown in photograph. Tapped at the junction of the Pi-Dux unit and B&W 3006 and at 34, 41, and 44 turns from the coupling-capacitor end.
- **K1, K2** — 4PDT, 15 kilohrns, 110 Vdc, (Allied Control T163X-25)
- **PC1,** PC2 — 6 turns 1-16" thick, ⅛" wide copper strip wound around 39-ohm, 2-watt resistors.
- **S1, S2** — Built up from Centralab index section and wafers.
- **T1** — 455 kHz i-f transformer (Workman TF11 although others will work) with primary winding and tuning capacitor removed and replaced with 50 turns #32 enameled, scramble wound next to secondary.
- **T2** — 455 kHz i-f transformer (Workman TF11).

speech amplifier and balanced modulator

The speech-amplifier and balanced-modulator circuitry is built as a unit in a small mini-



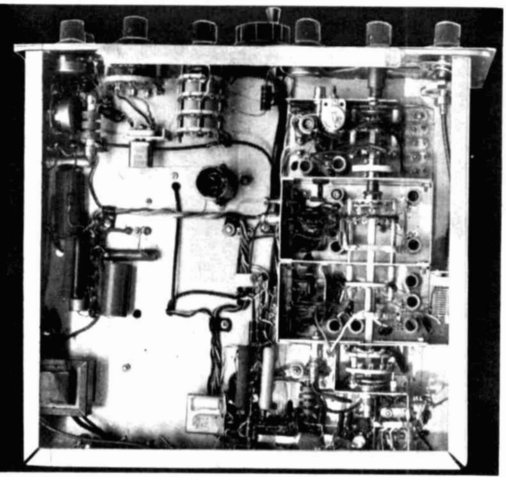
box. The balance potentiometer is located in front of the minibox with an extension shaft through to the front panel. The audio-gain control is located on a bracket attached to the minibox by the nut that retains the balance pot. Wiring to the gain control is accomplished with small shielded cable; power and signal leads come out through a hole in the bottom. It might be well to mention here that this method of construction not only provides excellent shielding, it makes for more pleasant building — you don't have to horse a big chassis all over the bench during construction. It also allows individual testing of sub-assemblies, a procedure which I highly recommend because it's much easier than aligning the whole exciter at one whack.

carrier generator

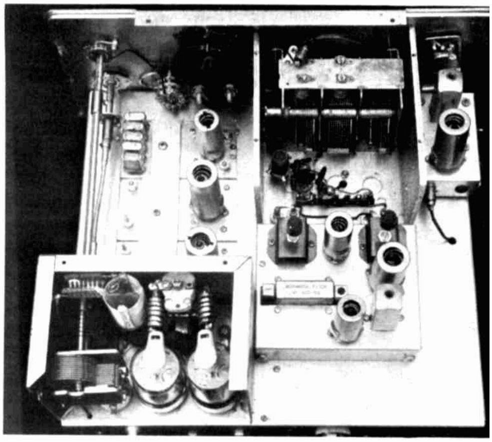
The 6CW4 carrier generator tube is located in the small space between the minibox and the front panel on the left side of the chassis as shown in **fig.** 6.

filter-mixer i-f amplifier

These stages are built into another minibox. Small sheet-metal protrusions provide room for the use of full-size '12-inch tuned coils on the input and output circuits of the tracking amplifier. All the leads from the box, as well as the coil leads to the ARC-5 tuning capacitor, are shielded. Access to the balanced-mixer nulling capacitor is available from the bottom of the main chassis through a small access hole. Proper tracking is established by juggling capacitor values in series with the ARC-5 variable capacitor sections. I trimmed some plates from my unit, but by proper choice of series capacitors, this should be unnecessarv.



Below-chassis view of the five-bend ssb exciter. The VFO tube and carrier generator crystals are in the upper left center; the bandswitch is to the right.



Inside the five-bend ssb exciter. Power amplifier compartment to the lower left.

variable frequency oscillator

The VFO is a standard Colpitts oscillator using a 6BA6 tube. The tube is located under the chassis and all circuitry associated with the VFO is contained in a shielded compartment. The 6BA6 is provided with a separate filament transformer and the heater runs as long as the exciter is plugged in. A number 47 bulb, which is used to keep the VFO compartment warm, is also connected. These provisions make warm-up drift negligible and no temperature-compensation circuitry is required.

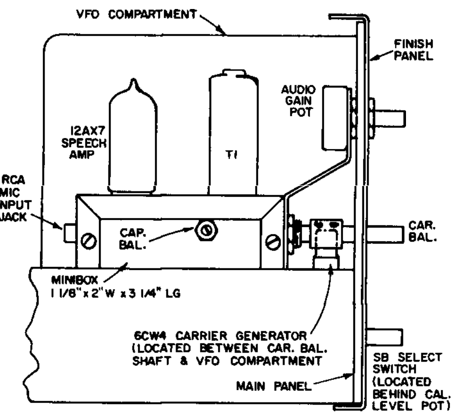
Plate and filament voltage to the VFO is run through feed-through capacitors from under the chassis. The VFO plate voltage is zener regulated at 120 volts. A small 10-pF variable capacitor is connected in parallel with the main tuning capacitor and provides about ±10 kHz variation for tuning-dial calibration. This capacitor was removed from the same ARC-5 receiver that yielded the main-tuning capacitor. It is placed between the tuning capacitor and the face plate; this is a tight squeeze but it fits after careful positioning.

The center section of the main-tuning capacitor is used for the VFO. The remaining sections are used to tune the tracking amplifier. Note that a capacitor is placed in series with the main-tuning capacitor—this limits the tuning range to 500 kHz, and makes the dial calibration extremely linear.

bandswitching details

The bandswitch is built up from Centralab wafer and index sections. These are coupled together with fixed and flexible couplings. The crystal oscillator, mixer and buffer are constructed separately in minibox sections and, after testing, assembled into a three-section unit and installed in the chassis. The

fig. 6. Side view of the chassis showing the position of the mechanical filter compartment on the left-hand side of the chassis. Note the lips at the top and bottom of the front finish panel.



main chassis is cut out slightly to provide clearance for tubes, slugs, and crystals as shown in the below-chassis photo.

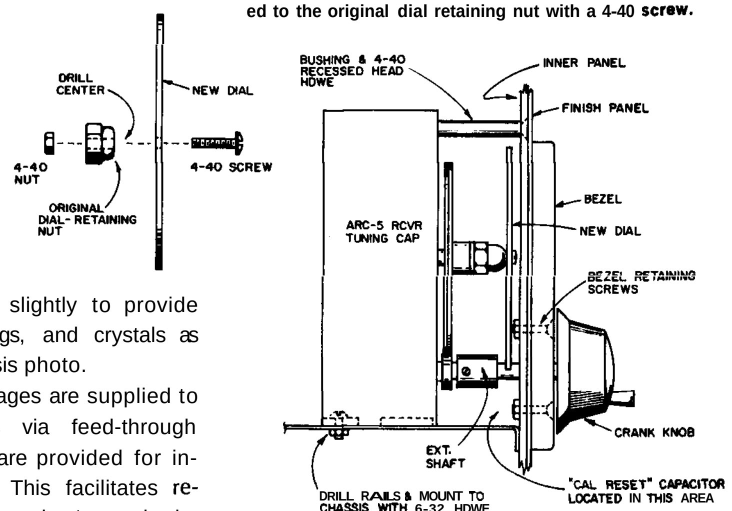
Plate and filament voltages are supplied to the individual sections via feed-through capacitors; phono jacks are provided for input and output signals. This facilitates removal of the assembly if service is required.

The small size of the final amplifier compartment prevents the use of a three-section output capacitor, so fixed capacitors are switched in on the 80-, 40-, and 20-meter bands.

In the 50.0, 50.5, and 51.1 position of the bandswitch, all ten-meter stages at 28.0, 28.5, and 29.1 MHz are turned on except the final amplifier. The output of the 6CL6 driver stage is taken from an RCA jack on top of the chassis and applied to an external six-meter transverter. The transverter is built in an identical cabinet which it shares with the power supply that runs the exciter.

It would be feasible, in the 14-MHz position, to run a doubler stage after the 18-MHz oscillator used to heterodyne the ssb signal to 14 MHz. The resultant 36-MHz signal could be mixed with the 14 MHz ssb signal already being generated to provide an output on 50 MHz. With a 6360 mixer, the entire set-up could be contained within the exciter itself. This would give 80- through 6-meter coverage with a tuning range of 50.0 to 50.5 MHz on six. It would be asking too much to run the 61468 finals on six meters, but the output from the 6360 could be used to drive an external amplifier or used barefoot at about 5 watts PEP. This circuitry is not included in the schematic, but is shown in the block diagram.

fig. 7. Construction of the driving mechanism for the ARC-5 tuning capacitor. The dial scale is mounted to the original dial retaining nut with a 4-40 screw.



final amplifier

The final amplifier is a conventional neutralized circuit using a pi-network output. The two 6146B tubes are connected in parallel and are provided with regulated grid bias and

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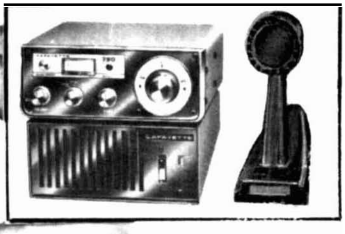

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screen voltage. The cathode pins on the tube sockets are bent inward toward the center and strapped together with a tinned-copper strip. Flat copper strips are also used for the plate parasitic suppressors, and many of the connections in the bandswitch assembly make use of the same material.

control functions

Most of the control functions are either self-explanatory or discussed elsewhere in the text. A few functions, perhaps, require further explanation.

The **calibrate** position of the function switch permits the operator to "talk" himself on frequency, or, if desired, the carrier may be inserted and a conventional zero beat can be accomplished. In either case, the push-to-talk switch (or key) must be depressed to energize the transmitter. The **calibrate level** control sets the spotting signal to a comfortable listening level.

In the **tune-CW** position of the function switch, the balanced modulator is bypassed to provide a CW carrier when the key or push-to-talk switch is depressed. Full break-in CW permits maximum operating convenience. This is accomplished through the use of high-speed relays.

The meter is switched to read either final cathode current or output. Antenna change-over and receiver muting are built in, as is push to talk. Any powrr supply that will provide the indicated voltages at the power-plug terminals may be used.

No tuning or alignment instructions are given here since it is assumed that anyone with the necessary skills to build an exciter such as this one will be thoroughly familiar with the proper techniques.

It is not expectrd that anyone will build an exact copy of this transmitter. However, some of the construction methods and circuits used here should be of interest to other appliance operators who prefer, as I do, to build their own appliances.

References

1. ARRL, **Single Sideband for the Radio Amateur,** American Radio Relay League, Newington, Connecticut. Third Edition.
2. D. Stoner, **New Sideband Handbook,** Cowan Publishing, New York, 1962.

Construction of the FET converter for 50 MHz. Input circuits a n to the left.

fet converters

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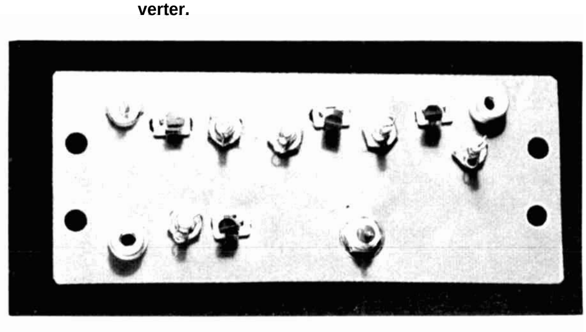
Several new field effect transistors are available to radio amateurs at very attractive prices from 90 cents to $1.10 apiece. These FET's are similar to vacuum tubes, with characteristic curves similar to pentodes, but with input-to-output feedback capacitance more like a VHF triode. When FET's are used in neutralized circuits, they provide fairly good gain values at radio frequencies and compare favorably or surpass tubes at VHF. These devices have less cross-talk or intermodulation problems than ordinary bipolar transistors and usually less than vacuum tubes. In addition, noise figures are normally lower than with tubes costing several times as much. Lower noise figures mean better weak-signal reception. At lower radio frequencies, FET's can be used without neutralization by mismatching load impedances; this is frequently done with bipolar transistor circuits. However, at VHF and UHF, neutralization of the feed-through capacitance is required for best noise figure and stage gain.

50-MHz converter

The use of grounded-gate circuits eliminates the need for neutralization, but with some deterioration of noise figure. The stage gain is roughly half as much as with gate input, grounded-source circuits with neutralization. Grounded-gate FET circuits are comparable to grounded-grid tube circuits but have lower noise-figures when the proper types of FET's are chosen. A good example is the 50-MHz converter shown in **fig. 1**. Here two stages of grounded-gate FET's provide enough gain at 50 MHz to permit the use of an FET mixer stage with its marvelous low cross-modulation characteristics.

FET mixers require several times as much oscillator injection power as a bipolar transistor mixer and only provide a fraction as much conversion gain as a good bipolar mixer stage. However, the low cross-modulation characteristic is extremely desirable in the 50-MHz band where signals are extremely strong during band openings, and an FET mixer is a must in a good converter. The problem is to provide enough rf gain ahead of the mixer stage to give a low overall noise figure, but not enough to overload the mixer when strong signals are present. The source bias resistors in **fig. 1** are values which resulted in the best noise figure with MPF-102 FET's when used with a regulated 10-volt power supply. A supply of 12 or 15 volts would require an increase in bias resistance to keep the drain-to-source current at 4 to 5 mA per stage.

The 50-MHz FET converter.



At my location there are no active 50-MHz stations nearby, so quite a bit of rf amplification can be used ahead of the mixer. In other locations, less gain might be desirable, and larger source bias resistors may be used in the two rf stages. In such cases, the bias resistor and bypass capacitor in each stage should be in the ground-return lead of the secondary winding of the tuned circuits. The bias resistor of each rf stage is connected to a single 500- or 100-ohm variable resistor so that the overall rf gain for a minimum mixer cross-modulation can be readily adjusted for optimum results.

The two 1N100 diodes connected back-to-back across the input jack of all of these converters is a standard procedure at W6AJF to protect the input transistor when operated in conjunction with high-powered transmitters. These diodes can produce cross-modulation if a neighborhood transmitter is being operated on a frequency near the desired signal reception spot. When this

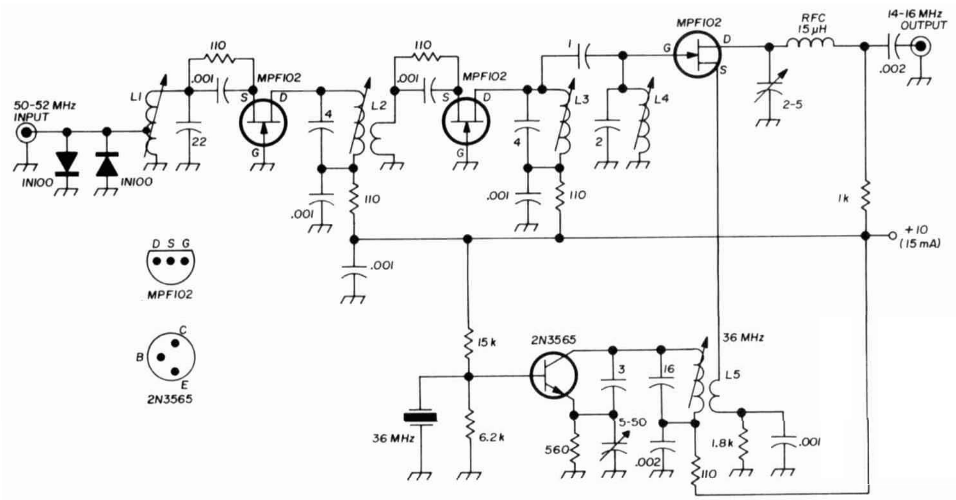


fig. 1. Six-meter FET converter with a noise figure of 1.5 dB. L1 is seven turns #20 on a 3/8" slug-tuned form, spaced 3/8" long, tapped at 2½ turns. L2, L3, L4, and L5 consist of 15 turns #24 enameled on n 3/8" slug-tuned form; secondary of L2 and L5 is two turns.

happens, a 6- to 20-dB pad in the antenna lead will cure the problem but will result in pretty poor reception of anything but fairly strong local or skip signals.

Any nearby transmitter which can produce cross-modulation in the front-end diodes, will overload the mixer stage, even with little or no rf stage gain, so a pad in the coaxial line or a null position of the beam antenna is about the only solution.

In **fig. 1,** four 50-MHz slug-tuned circuits provide a fairly flat frequency response over a 2- to 3-MHz bandwidth. The output impedance-matching circuit will cover as much as 4-MHz bandwidth. This circuit is an L-matching circuit resonant at 15- or 16-MHz to transform the mixer impedance of several thousand ohms down to the 50- or 75-ohm coaxial line to the communications receiver. Mixer injection from the oscillator is introduced at the source electrode. The coupling coil to the oscillator should be adjusted in number of turns and coil coupling to provide about 1/2 volt of oscillator energy at the mixer source terminal. A diode rf voltmeter is useful at this point. If an rf voltmeter is not available, oscillator injection can he adjusted to a level which gives a little less than maximum mixer gain from a signal source. This is because the modulation characteristics are not obtained at maximum mixer conversion gain.

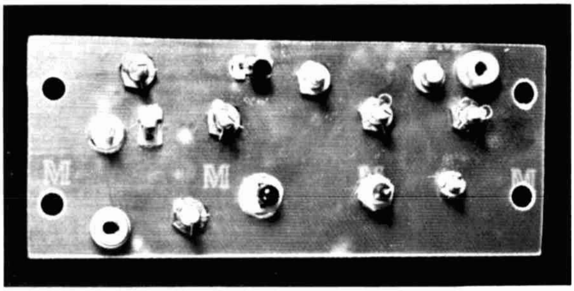
The crystal oscillator circuit can be used with any overtone crystals in the range of 20- to 40-MHz by changing the collector circuit L-C values so the tank circuit tunes to the desired overtone frequency. The adjustable 50-pF emitter bypass capacitor is useful in obtaining maximum oscillator output. The base resistors should be values which will

The 144-MHz FET converter; noise figure below 2.5 dB.

hold the collector current of the 2N3565 to about 2 or 3 mA. These Fairchild 2N3565 plastic-cased bipolar transistors have high-gain values and give excellent outputs in overtone-oscillator and frequency-doubler or -tripler circuits. Other types of bipolar transistors having an h_{fe} over 100 and an f_t of at least 200 MHz could be used as an oscillator.

The TIS34 FETs from Texas Instruments can be used in place of Motorola MPF-102's (or -104's) by changing the source bias resistor in the two rf stages. Values of 330 to 510 ohms are required to keep the drain current to 5 mA or less with TIS34's. Any of the FETs

Inside the two-meter converter. The input coil and small neutralizing coil are to the left; the oscillator. lower right.



used in these converters will function as mixers up to about 200 MHz with 1000- to 1800-ohm source bias resistors and drain currents of 1 to 2 mA with oscillator injection.

The 50-MHz converter shown here has a noise figure of about 1.5 dB in the 50- to 52-MHz range. If the unit is stagger-tuned to obtain 4-MHz bandwidth, the gain drops, and the average noise-figure is around 2.5 dB. This is low enough for nearly all "average-noise" locations, but the 1.5 dB noise figure is better for forward-scatter signal reception and helps on weak voice signals even in average locations. Man-made noise in some locations can wipe out the performance of any low-noise front end and cause much operator frustration.

144-MHz converter

At 144-MHz, the MPF-102 doesn't give as low a noise figure as the slightly more expensive TIS34 or Union Carbide UC734. The latter seems to be a little better than the TIS34 and was used in the first stage of the 144-MHz converter with the feedback capacitance neutralized out with a tuned coil connected from gate to drain through a dc blocking capacitor. The neutralizing coil should be mounted on the gate side of the interstage shield and enough copper plating scraped away from the coil-mounting area to eliminate grounding the tuning slug and mounting hardware. This slug-tuned coil has to have enough inductance to resonate at 145 MHz with the gate-to-drain capacitance of the transistor plus the miscellaneous shunt capacitances across the coil. A good starting point is 20 turns of number 30 enameled wire on a 3116-inch diameter slug-tuned form. More or less turns may be needed with the layout and FET used in order to put its tuning

meant that an FET mixer, with its attendant low-conversion gain, was highly desirable to reduce cross-modulation from mountain-top relay stations and other strong signal sources. This means that two rf stages are generally desirable in order to give good weak-signal reception. A minimum of four tuned circuits plus a large cavity-tuned antenna filter is needed to put image signals down -80 dB or more. The converter shown here has -60 dB image suppression by itself for the 14- to 18-MHz i-f output range. The noise figure measured from 1.8 to 2.5 dB over the 144- to 148-MHz range. This results in very good weak-signal reception with a large antenna and fairly low-noise location.

The second stage of the 144-MHz converter is a grounded-base circuit using either a TIS34 or UC734. Even a MPF-102 does a

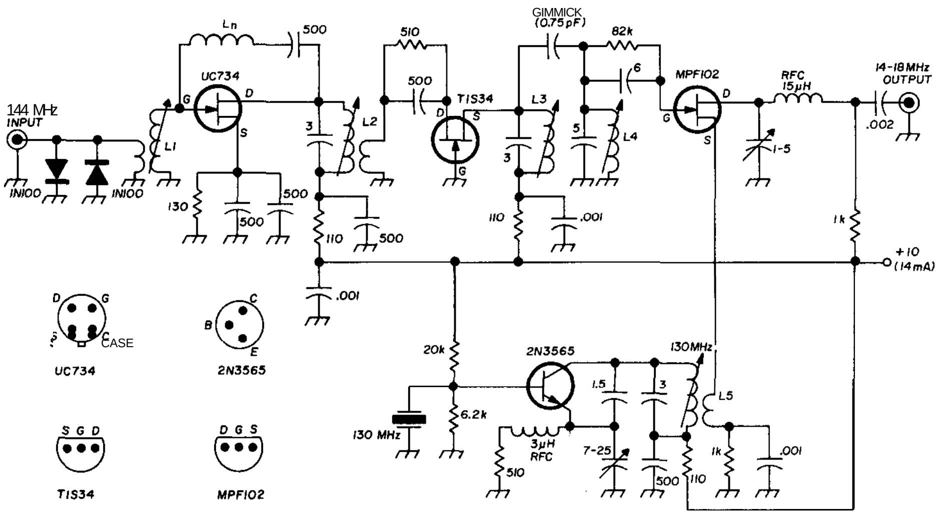


fig. 2. Low-noise FET converter for 144 MHz. L1, L2, L3, L4, and L5 consist of five turns #20 spaced 5/16" on 5/16" diameter slug-tuned form; primary of L1 and secondary of L5 are one turn; secondary of L2 is two turns. The gimmick capacitor is a short length of hookup wire twisted together as shown in the photo. The neutralizing inductor L_n consists of 30 turns #30 enameled closewound on a 3/16" diameter slug-tuned form.

range within the 144 MHz band. Careful adjustment of coils L_1, L_2, and L_n is needed to arrive at good neutralization and lowest noise figure in the two-meter band.

The popularity of the two-meter band, with its great number of stations in nearly all areas,

good job at 144 MHz, but each of these FET's requires a different socket connection, so it's difficult to make comparisons. When this converter was first built, a few TIS34's were occasionally available, and MPF-102's and selected MPF-104's were readily available. The UC734 units weren't out yet, so the Union Carbide 2N4416 was used. The 2N4416 costs several times as much, but is quite similar to the UC734 FET's.* Later tests with several UC734's indicated the same gain

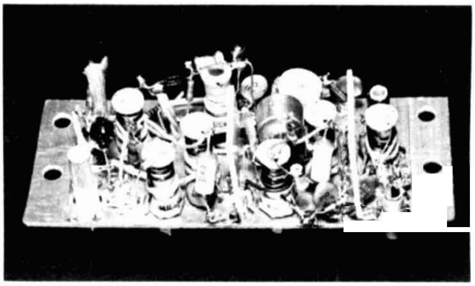
The UC734 is essentially the same as the 2N4416 except for a slight relaxation in g_m, I_{GSS} and I_{DSS} specifications.

and noise figure results as with the 2N4416's. The socket connections are identical, and in testing either type, some adjustment of the neutralizing coil was needed to eliminate rf oscillation in the first stage.

The overtone crystal oscillator in fig. 2 functions quite well with 130-MHz seventh-overtone crystals at 1 to 2 mA collector current. If 43.33-MHz crystals are used, the base-bias resistors must be changed to get about 3 mA collector current to drive the FET mixer. These mixers require from 3 to 5 times as much oscillator injection voltage as a bipolar mixer, and if there isn't enough, the conversion gain rapidly drops off.

220-MHz converter

Except for the mixer and oscillator stages, this converter is somewhat similar to the one built for the 144-MHz band. The mounting



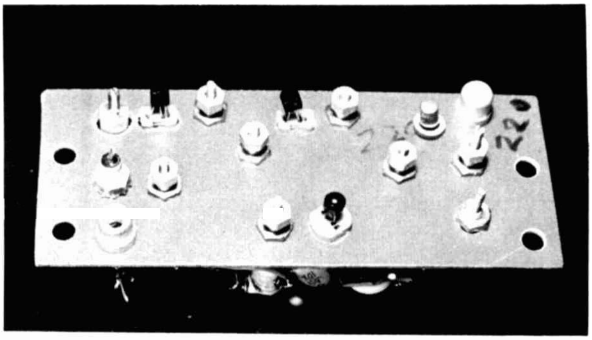
Layout of the 220-MHz converter. **Input coil and neutralizing coil to the right; oscillator at top center.**

plate is five inches long by two inches wide, copper plated on the bottom side; this is the same as the 50- and 144-MHz converters. As can be seen in the photograph, to get all the circuitry into a 5 x 2-inch space results in a pretty cluttered layout. A surplus low-frequency "rolling-pin" overtone crystal was available; the oscillator triples to 69 MHz and a low-capacitance, fast-computer diode, a 1N914, is used to triple again to reach 206 MHz for oscillator injection into the mixer stage.

The output from the 1N914 was too low for an FET mixer, and since this band isn't ordinarily bothered with strong signals at my location, a bipolar 2N3478 mixer was connected as shown in fig. 3. The resultant gain of the whole converter is very high, and in some locations the rf stage or stages may need a gain-control potentiometer in series with the source-bias resistors. TIS34's, if available, or UC734's can be used in both rf stages.

The TIS34 has higher feedback capacitance and would require less turns on the neutralizing coil (L_n). It would also require different socket connections and, perhaps, a 200-ohm bias resistor. Again, the bias resistors depend upon the power-supply voltage. The first FET should be run at approximately 5 mA drain current, and the second stage at a little less.

The 220 MHz converter using MPF-104 FET'S.



The noise figure of this converter seems to be very good, measuring 2.0 dB at 220 and 222 MHz and 2.5 dB at 224 MHz. The 2N4416 and UC734 FET's are a little better than the TIS34 units I tested. The UC734 could have been used to advantage in the second stage with a change in bias resistor and socket connections, but its higher gain wasn't necessary.

The oscillator injection voltage into the 2N3478 mixer was measured at from 0.1 to 0.2 volts with a diode rf voltmeter as compared to the usual 0.5 volts or more for the FET mixer. If the station location requires an FET mixer because of strong signals from radar or radio stations, the oscillator chain would require a much more expensive diode tripler using a varactor or other highly efficient frequency multiplier. The 69-MHz output of the oscillator would probably have to be set up for maximum rf power output. An

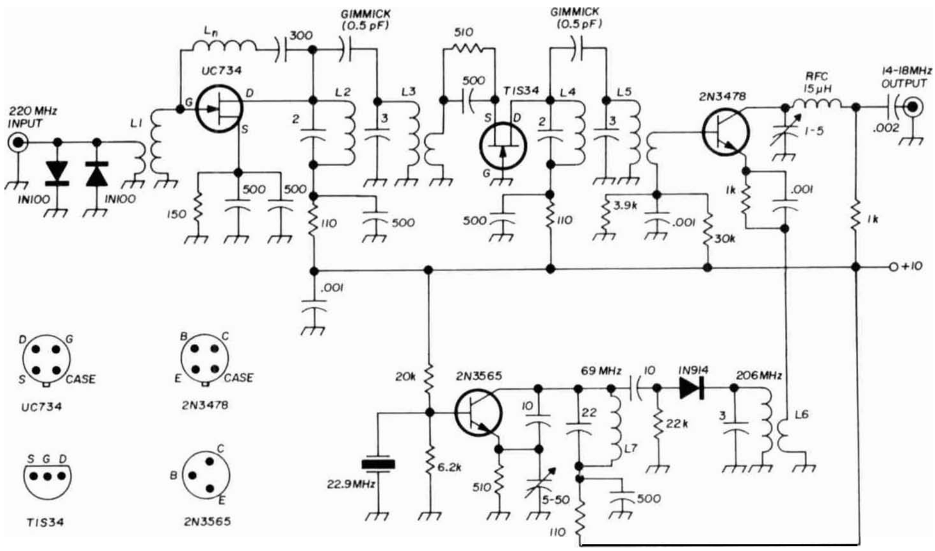


fig. 3. A FET converter for 220 MHz using low-noise field effect transistors. L1, L2, L3, L4, L5, and L6 consist of four turns #20 spaced 5/16" on a 5/16" diameter slug-tuned form, primary of L1 and secondary of L5 and L6 are one turn, secondary of L3 is 1½ turns. L7 is five turns #20 spaced 5/16" on a 5/16" diameter slug-tuned form. The neutralizing indicator L_n is ten turns #26 enameled closewound on a 3/16" diameter slug-tuned form.

alternate solution would be the use of a selected 2N3465 or 2N3463 as the second tripler stage in order to get enough 206-MHz rf ouput from the tripler.

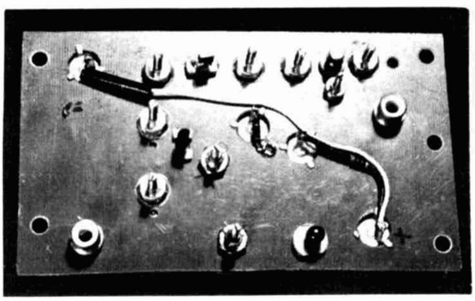
432-MHz converter

Since I wanted to use copper-strap tuning circuits, the 432 MHz converter required a little more space—it was built on a 5 x 2-3/4-inch copper plated board. The photographs show the layout of the copper-strap tank circuits. These resonant circuits, fig. 4, were made by using pieces of thin sheet copper about 1-1/4 inches long and 1/4 inch wide, soldered to the chassis or button feed-through capacitators at one end, and to the sleeve of the plastic piston-type tuning capacitator at the other. Since these piston capacitators won't stand much heat, the soldering has to be done with the sleeve removed from the plastic insulation. Glass or ceramic piston capacitators would be more desirable, but at the time this converter was built, I was too "Scotch" to buy the glass types, and the economical, ceramic, German-made units had not arrived on the West Coast.

In addition, when I built this converter only 2N4416 and TIS34 field effect transistors were available. The 2N4416 is expensive, but much better at 432 MHz than the TIS34, so I used one in each of the two rf stages. When the Union Carbide UC734 became available, one was used in the first rf stage. If I hadn't been so lazy, I would have modified the second stage to use the newer transistors. It is interesting to note that one-third of the UC734's tested in the first stage were as good as the best of four 2N4416's I tried.

The copper plating on the mounting board should be scraped away from the mounting hardware of the neutralizing coil (L_n) to minimize its effects on the input- and output-tuning adjustments. This effect can be very frustrating to the operator when first set-

Low-noise 432-MHz converter with 2N4416's.



ling up one of these hot 332-MHz neutralized stages.

The final result of all adjustments in the first stage is to prevent rf oscillation, obtain best noise figure and good, but not maximum, gain. With only one rf stage these adjustments would probably also coincide at maximum gain. With two rf stages, oscillations took place several MHz up the band where the antenna-lead impedance became something other than 50-ohms resistive. With this in mind, the input and neutralizing circuits must be adjusted to knock out this spurious oscillation for smooth operation and best noise figure. The drain circuit was always tuned for maximum gain at 432 MHz.

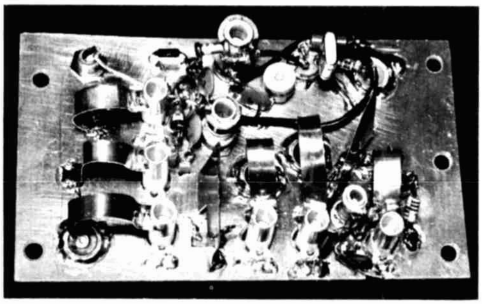
The following source circuit was detuned slightly and closely coupled to the first stage to help get rid of oscillations when the antenna impedance changes with frequency or weather. The following tuned circuits (three in a row at the end of the converter strip) were tuned for maximum signal response.

The 418-MHz oscillator injection voltage to the mixer was obtained in the simplest way possible—a 139.333-MHz overtone-crystal oscillator and 1N914 tripler. No attempt was made to use an FET mixer at 432-MHz.

One item that I have not mentioned was the use of two disc bypass capacitors from source to ground in the first stage of the neutralized rf amplifiers. This capacitor will normally tend to cool off the first stage for rf oscillation when using socket-mounted transistors. I am not enough of an optimist to believe that a converter can maintain peak performance without changing front-end transistors—hence the sockets!

ham radio

FET's on 432 MHz. Two **rf** stages with copper-strap tank circuits; **rf** input to the right. The mixer and oscillator circuits are on the left.



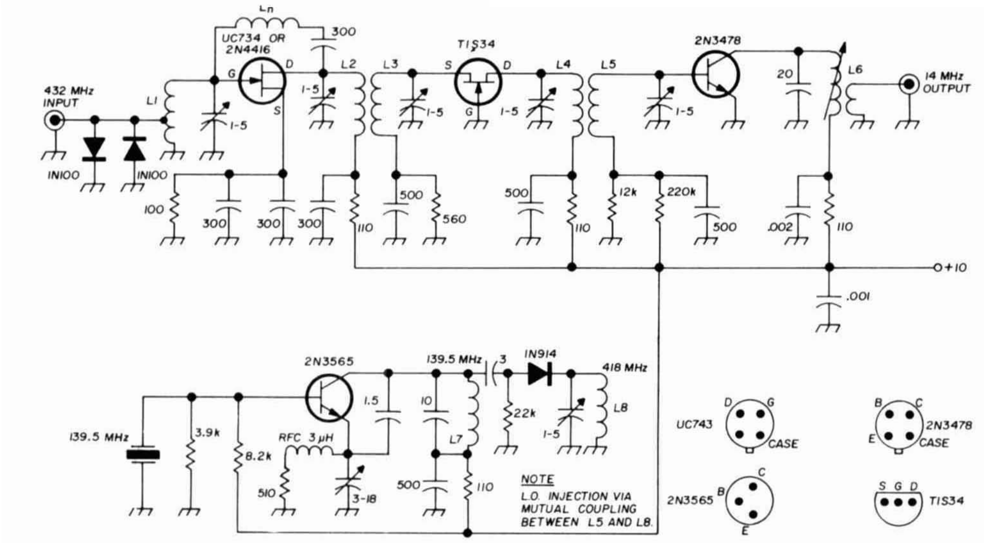
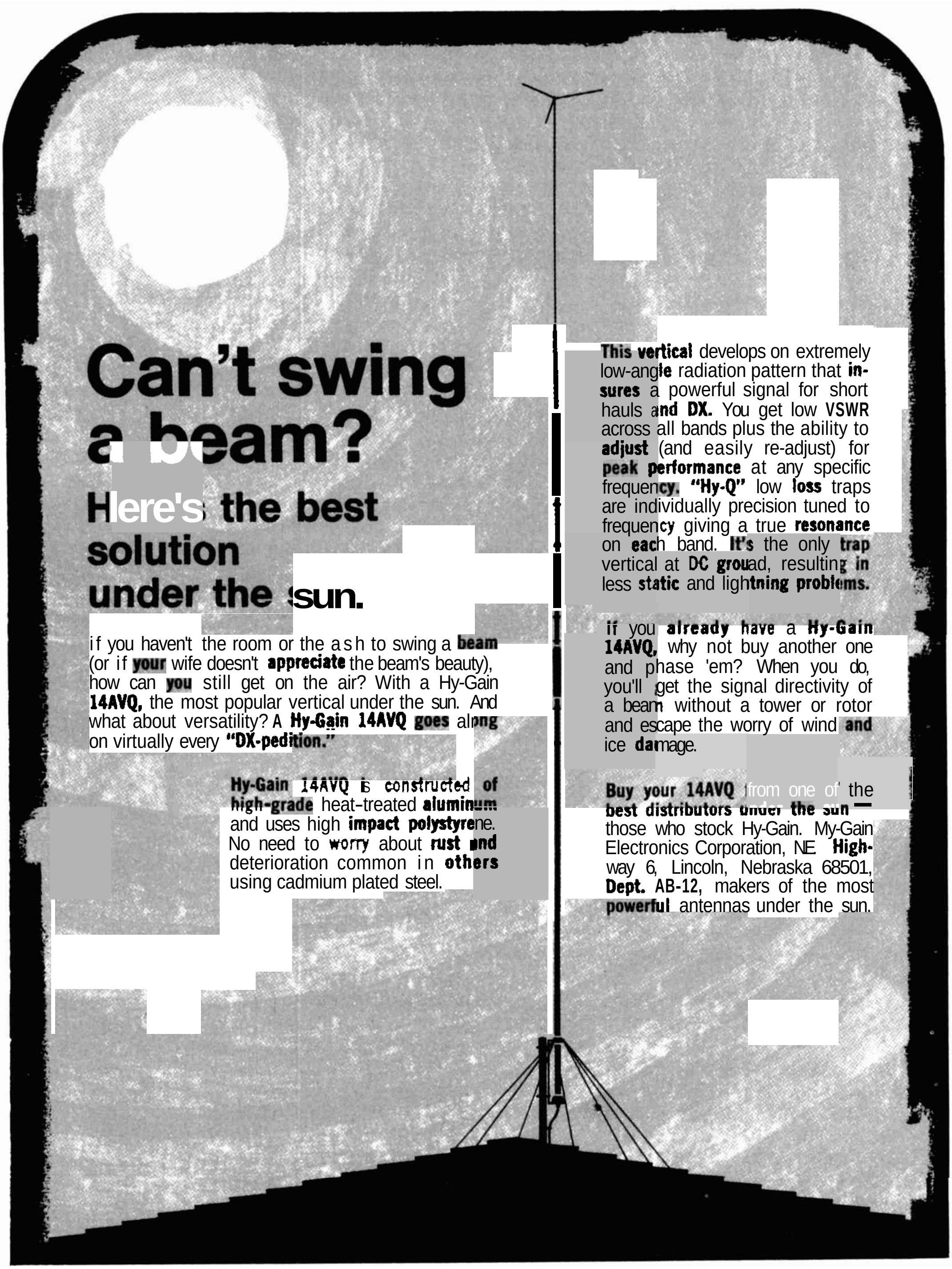


fig. 4. This low-noise converter for 432 MHz exhibits a noise **figure** of 2.5 dB. **L1, L2**, L3, **L4**, L5, and **L8** are copper straps approximately **¼"** wide by **1¼"** long. Local-oscillator injection into the mixer is accomplished by **mutual** coupling between **L5** and **L8. L6** consists of 28 turns #30 enameled closewound on a **¼"** slug-tuned form with a **2½** turn **secondary.** L7 is 3 turns #20 spaced **¼"** long on a **5/16"** slug-tuned form. **L** is 4 turns #20 spaced **¼"** on a **3/16"** slug-tuned form.



Hy-Gain 14AVQ

HY-GAIN ANTENNAS, FOR THE MOST POWERFUL SIGNAL UNDER THE SUN

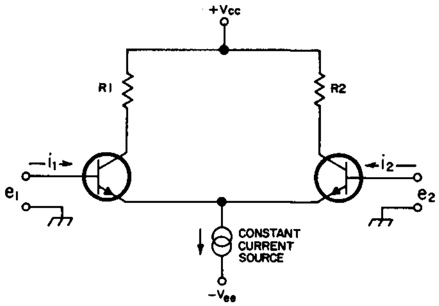


fig. 1. The basic differential-amplifier circuit which is used in most linear integrated circuits. This circuit requires both positive and negative supply voltages.

IC-regulated power supplies for integrated circuits

These plus and minus power supplies for IC work use a new circuit element— the integrated-circuit voltage regulator

Hank Olson W6GXN, 3780 Starr King Circle, Palo Alto, California 94306

The big push in integrated circuits in the last year or so has been in the linear IC area. With internal fighting between RTL, DTL, ECL and TTL digital types still going on, many IC manufacturers seem content to make all these logic families available, and concentrate their new-product efforts on linears. As a result of the "linear-push," prices have dropped to the point where amateurs can become linear IC users.

The linear IC configuration which is the most practical to fabricate is the differential amplifier. A basic amplifier of this type is shown in **fig. 1**. It has become popular because it is versatile and is readily adapted to single-chip construction. Since the resistors and transistors are on the same chip, nearly perfect temperature-tracking is obtained— i.e., things stay balanced! A typical linear IC (operational amplifier) is shown in **fig.** 2; note the differential circuitry. Note also that this amplifier is entirely dc coupled.

It is this dc-coupled, differential-amplifier configuration which dictates that most linear IC's require equal positive and negative power supplies. Two standards seem to be emerging in the field: those requiring plus and minus 6 volts, and those requiring plus and minus 15 volts. The following dual-regulated supplies are presented to simplify testing of larger circuits utilizing these new linear IC's.

dual-regulated power supplies

Both of the dual-regulated supplies described here use a new linear IC made especially for this service: the National Semiconductor Corporation LM300.* It sells for $6.40 in small quantities and is an economy version of an IC regulator (the LM100 costing nearly ten times as much.) As such, it is a real bargain for ham use. The various application notes which describe the use of the LM100 apply to the LM300 within specification limits.[1,2] The supplies are basically the same as those described in reference 1, but use less expensive semiconductors for amateur use.

A duel IC-regulated power supply which provides plus end minus fifteen volts.

The first supply is designed for linear IC's which require plus and minus 15 volts. It uses a slightly modified Triad F40X power-transformer to provide two separate 13.4 Vac windings, instead of the 26.8 Vac center-tapped configuration. This modification is very simple—the insulating paper is simply cut to expose the point where the centertap lead is soldered to the two formvar windings. The separated formvar leads are then soldered to new flexible insulated leads and the

* The name of your local National Semiconductor distributor may be obtained from National Semiconductor Corporation, 2950 San Ysidro Way, Santa Clara, California 95051.

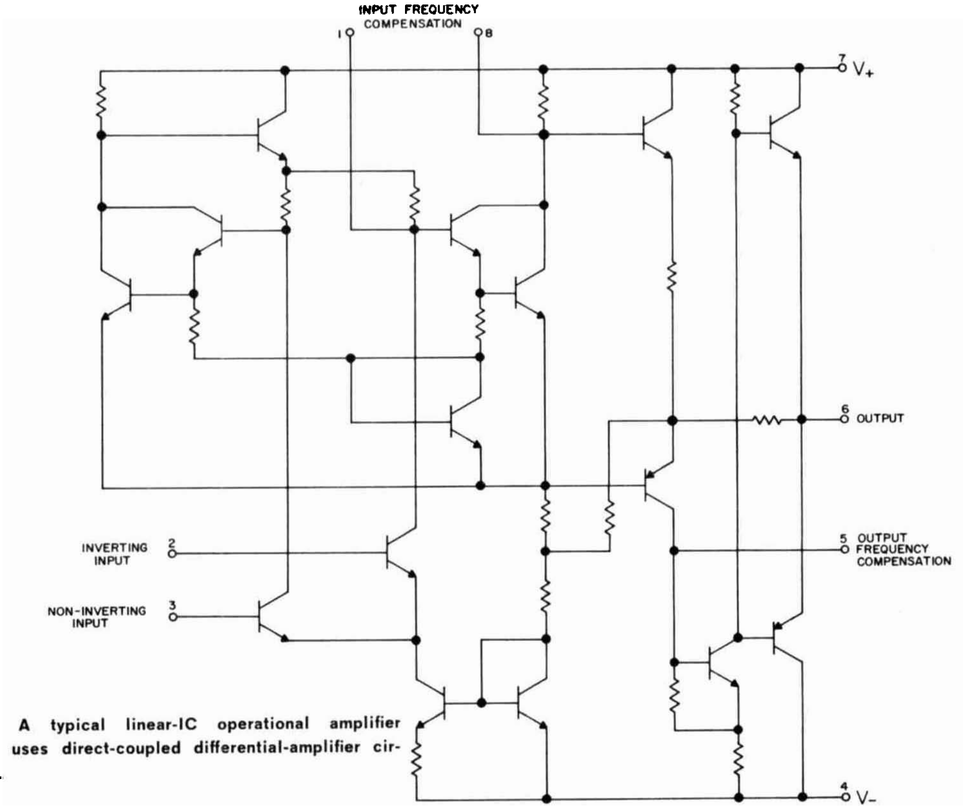


fig. 2. A typical linear-IC operational amplifier which uses direct-coupled differential-amplifier circuity.

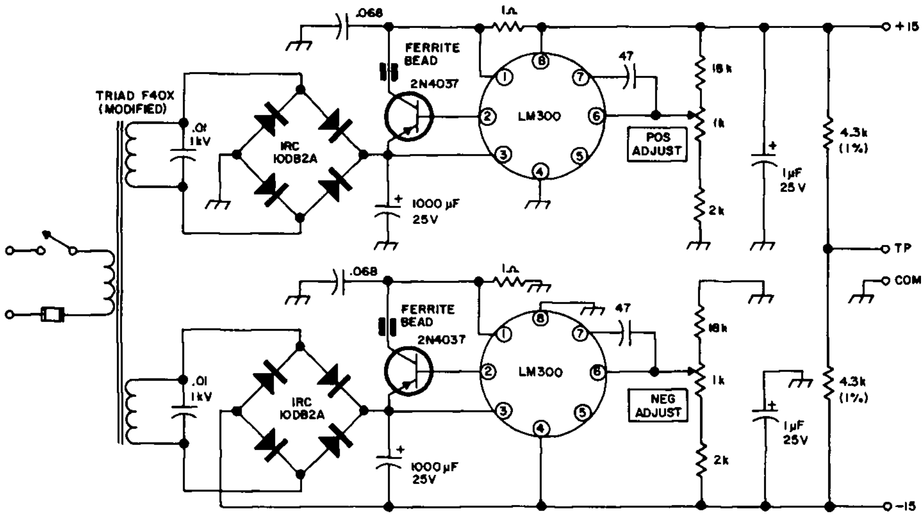


fig. 3. A dual-regulated **IC** power supply using **LM300** voltage regulators which puts out ± 15 volts. Tha **2N4037** power transistors **are** mounted on **Wakefield 254S1** insulated **heat** sinks. Tha ferrite beads may be purchased from Ami-tron **Associates.***

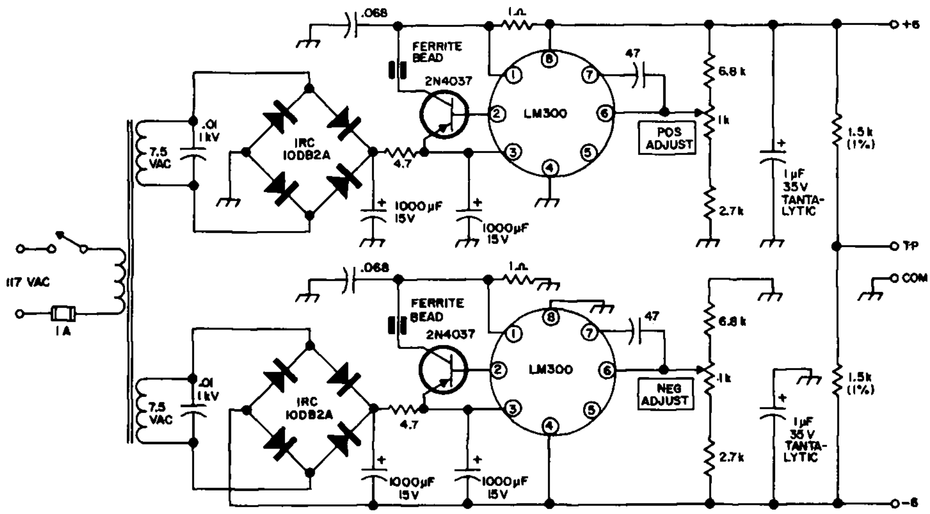
gouged paper is covered with RTV Silastic for insulation and support. The IRC 10DB2A encapsulated-bridge rectifier could be replaced with a Motorola MDA920-4 or HEP176, or by a Mallory FW200.

The second supply is designed to accommodate those linear IC's that require plus and minus 6 volts. The transformer for this unit is a dual 7.5 Vac filament type which costs between three and four dollars.

A **balance** test point (TP in **fig.** 3 and **4)** is provided in both circuits for a quick check of symmetry between the plus and minus output voltages. Two matched resistors (1%) are series-connected to draw a few milliamperes of current between supplies. If the two supplies are providing equal voltages, the TP should be at zero voltage, as measured to ground with a VTVM. The TP is normally connected in operation of the supply.

fig. 4. Dual regulated supply which provides ± 6 volts. The **2N4037's** ara mounted on Wakefiald **254S1** heat sinks; **the** ferrite beads are available from Ami-tron Associates.

* Ami-tron Associates, 12033 Otsego Street, North Hollywood, California 91607.



For those of you who are distrustful of IC's in power-supplies, fig. 5 represents a plus and minus 15-volt supply using discrete components. Of course, a person who distrusts IC's in the power supply would not likely want a ± 15-volt supply to test other IC's, but this discrete circuit dramatically shows how much extra circuitry the LM300's eliminate. A careful cost comparison, using Allied Radio's catalog, will show that there is only a dollar or two difference between the discrete and integrated-circuit supplies.

construction

Several precautions are necessary in building the supplies shown in fig. 3 and 4. The 47-pF capacitor between pins 6 and 7, and the 1-μF capacitor between pins 4 and 8 of the LM300 should be placed as close to the IC as possible. One of the regulators developed some parasitic oscillations at higher currents; the 0.068μF capacitor from pin 1 to chassis-ground and the ferrite bead on the 2N4037 collector lead corrected the problem. Even though this trouble only showed up in the unit when it was spread out on the bench, these measures were included in both circuits.

operation

Both IC-type supplies are capable of at least 100 mA output with less than 10 mV drop in voltage. The regulator sections are not the limitations in these circuits; they will operate to 200 mA with 0.1% regulation. The rectifier-filter system is the limiting factor. The circuit shown in fig. 5 can also be used to 100 mA; the output voltage drops about 100 mV at 100 mA and 2.5 volts at 200 mA.

References 3 and 4 are recommended for general information on linear IC's. These inexpensive paper-bound books show applications for, and precautions to be taken in, using linear integrated circuits.

ham radio

bibliography

1. Widlar, R., "A Versatile, Monolithic Voltage Regulator," National Semiconductor Corporation Application Note AN-1, February 1967.

2. Widlar, R., "Designing Switching Regulators," National Semiconductor Corporation Application Note AN-2, April 1967.

3. R.C.A., "Integrated Circuit Fundamentals," R.C.A. Technical Series, IC-40, July 1966.

4. Giles, J., "Fairchild Linear Integrated Circuits Applications Handbook," Fairchild Semiconductor Corporation, 1967.

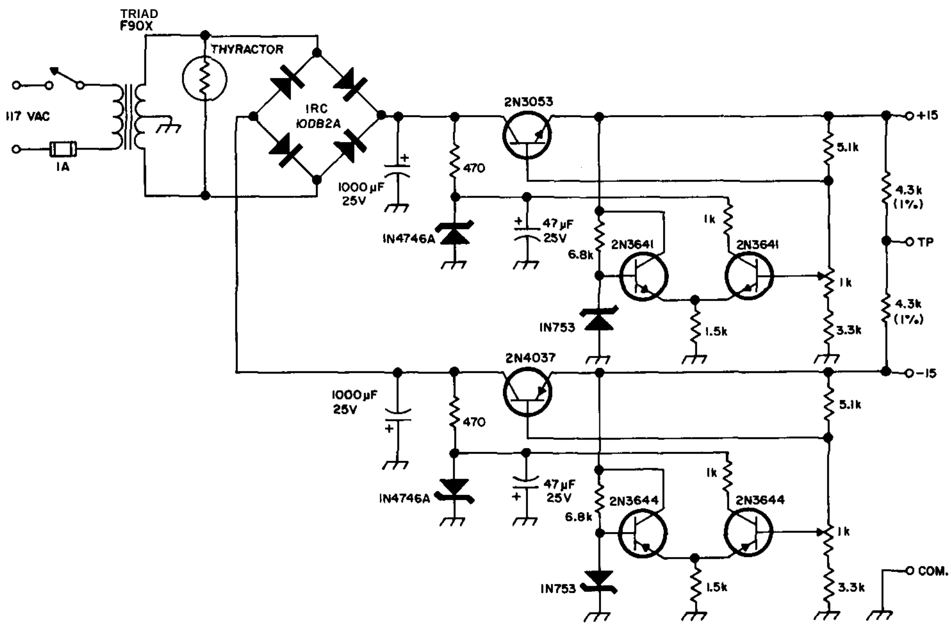
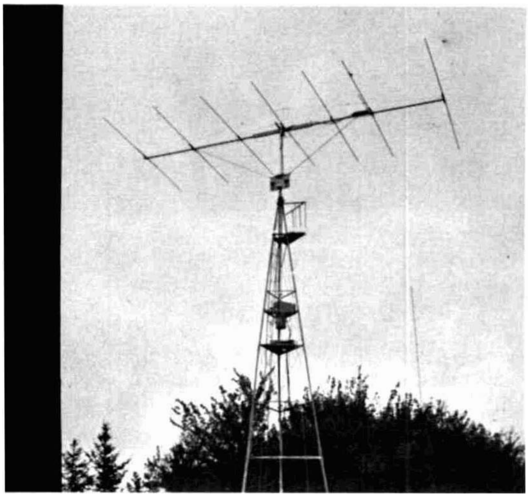


fig. 5. Dual-regulated 15-volt supply which uses all discrete components. There is little price difference between this model and the one which uses an integrated-circuit voltage regulator. The 2N3053 transistors are equipped with clip-on heat dissipators.



The completed seven-element. remotely - tuned, ten - meter beam mounted on top of the 50-foot tower. The two support struts below the beam are made from dural tubing end give the long boom additional support.

a big beam for ten meters

George Cousins VE1TG, RR 2, Box 18, Lower Sackville, Nova Scotia

Here's a beam that's designed for performance–seven elements on a thirty-foot boom

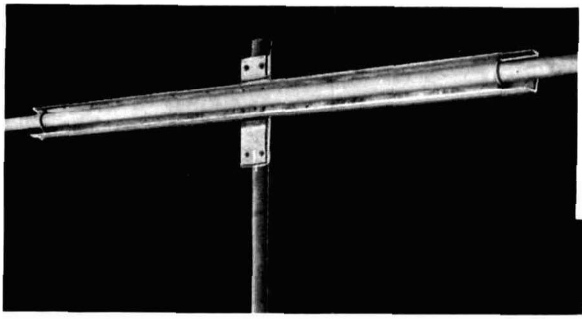
With the 10-meter band returning to life, and the rapidly improving DX conditions, thousands of hams are turning to thoughts of antennas. The usual controversies are developing over which type is best and which to erect. At my station, both beams and quads have been used with excellent results on ten. The last effort, a two-element quad, proved to be especially good.

However, the building urge came along just after the last series of DX contests, and I felt that the next project should be a new yagi. This beam would be somewhat larger than previous efforts, with emphasis on front-to-back ratio and ease of tuning. The last effort along these lines was a four-element affair which performed well enough to indicate that a few more elements would be even better. The antenna which eventually took shape is described here and is an excellent project, even for a beginner. Construction is straightforward, and although the antenna is large, it may be scaled down to a size you consider appropriate for your station.

design

This beam is a seven-element yagi with a 30-foot boom. This configuration requires a lot of attention to mechanical as well as electrical details—particularly if you live in a bad-weather location. However, the results obtained with this beam are well worth the effort put into its construction. An interesting feature, and one which I have never seen described before, is the remotely-controlled gamma match which can be precisely operated from the hamshack. This permits accurate gamma-match tuning over the entire ten-meter band. Anyone who has ever attempted to tune a beam from the top of a tower can appreciate the convenience of this device.

Center support for the boom shown mounted on the mast.



You should start this project with pencil and paper rather than with drill and metal. First, decide upon the number of elements. You should consider cost, availability of materials, time, experience, and probably most important of all, the mechanical characteristics of the tower and rotator. Even with light-weight materials, this antenna presents a lot of stress on its supporting structure. By using standard formulas, you can work out the dimensions (in feet) of the beam elements for any chosen operating frequency:

$$\text{Driven element} = \frac{473}{\text{Frequency (MHz)}}$$

$$\text{Directors} = \frac{450}{\text{Frequency (MHz)}}$$

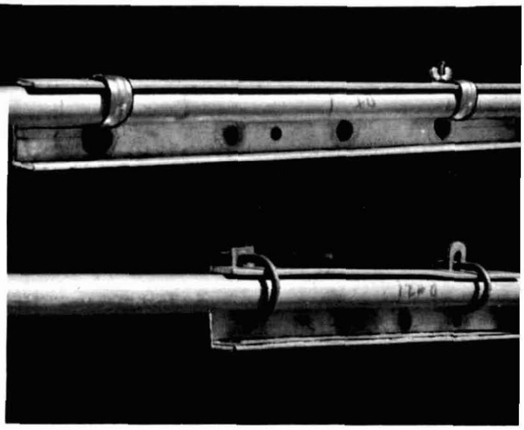
$$\text{Reflector} = \frac{501}{\text{Frequency (MHz)}}$$

Although the number of elements makes the bandwidth narrower than a smaller beam, this is not a disadvantage, since few operators actually cover the entire ten-meter band. In the case of the phone DX'er, for example, the beam would be optimized around 28.4 to 28.6 MHz, while the CW man would prefer a design centered on 28.1 MHz or so.

construction

A look at fig. 1 will show the general arrangement of elements. The 30-foot boom permits spacing to the operator's preference. For wider bandwidth and less critical adjustment, the elements should be spaced out to about seven feet. However, this will only allow room for five elements. In my case, I was interested in maximum front-to-back ratio rather than gain, so the element spacings I used were selected to achieve this purpose. This is indicated by the wide spacing from driven element to reflector, and narrow spacing from driven element to first director.

Two methods of fastening the elements to the boom mounts—standard pipe clamps and muffler clamps.



With this spacing, I was able to add two more elements, sharpening the beam pattern and increasing the front-to-back ratio.

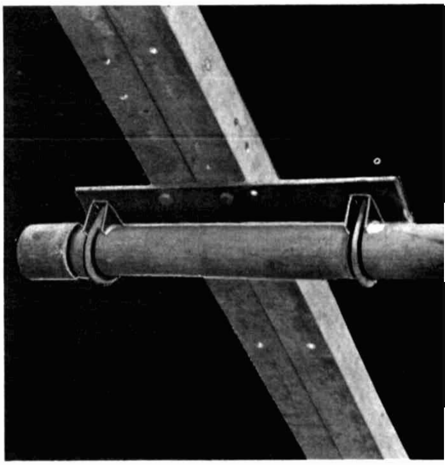
Despite the obvious advantages of so many elements, there is a matching problem because the radiation resistance falls to a low value. This not only makes it more difficult to match the feedline to the driven element, it means that the operating bandwidth for a given SWR is quite narrow. The remote gamma-match tuning was incorporated to overcome this problem.

the elements

In selecting materials for the beam, keep the important factors of strength and weight in mind. Aluminum tubing is the logical choice—6061-T6 alloy. This tubing comes in standard 12- to 13-foot lengths. A full length should be used for the center section of each element. The extension pieces on each end of the element are made from smaller diameter tubing which telescopes into the larger. I used 'Is-inch diameter, 0.058 wall, for the larger tubing, and 3/4-inch diameter for the smaller. The 3/4-inch tubing can have very thin walls because little strength is required in such short lengths. The 0.058-inch wall of the larger tubing results in a snug fit when the sections are telescoped together.

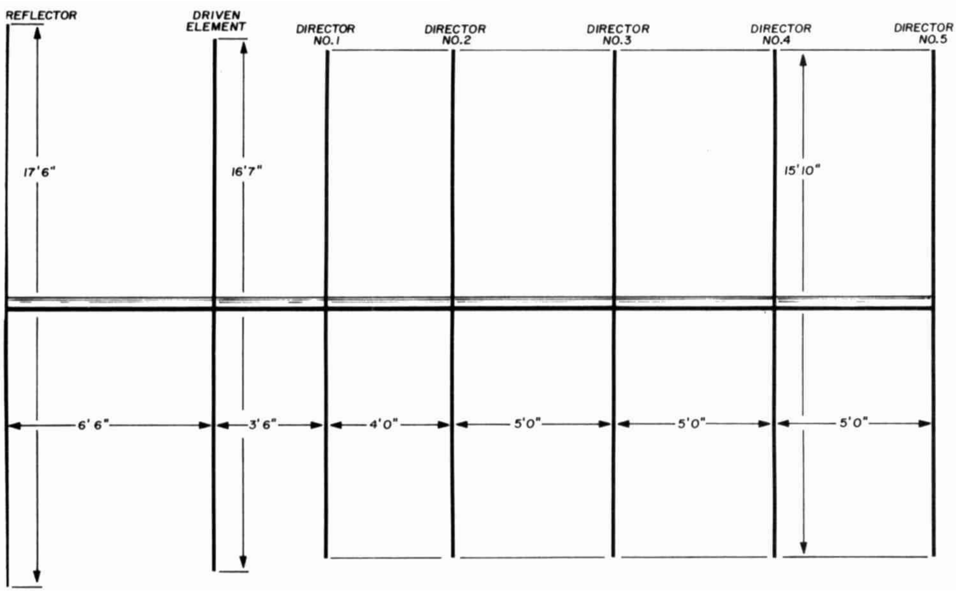
The ends of the large tubing are split with a hacksaw for about one inch. Then the short extension pieces are inserted, and the overall length of each element is adjusted to the proper dimensions. A stainless-steel hose clamp is used to clamp the large center tubing over the smaller end sections. This prevents them from slipping in or out. It's also a good idea to paint the joints with aluminum paint and wrap them with waterproof tape. Each element can be made up in the basement or garage and set aside until they're put on the boom.

Several methods can be used to mount the elements on the boom. I prefer small pieces of aluminum channel which are fastened to the boom with standard automobile muffler clamps. The element is laid in the trough of the channel and fastened securely in place with pipe clamps or small U- or J-bolts. Since aluminum channel is expensive unless you can find some in a junkyard, small sections of angle iron or aluminum will do just as well. Just remember to give special attention to rust-proofing anything that isn't aluminum.



Detailed view of the boom support and boom-to-mast mounting plate.

fig. 1. Overall dimensions of the seven-element beam designed for a center frequency of 28.5 MHz.

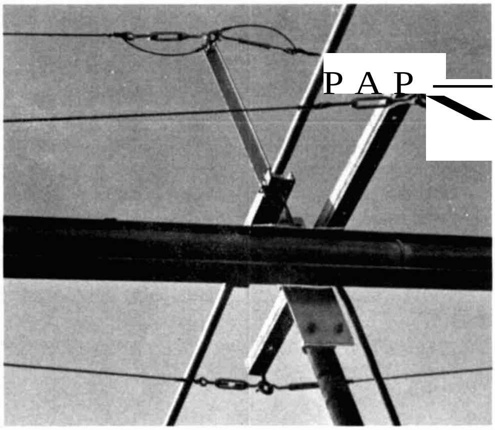


Muffler clamps may still be used to fasten the angle to the boom, and the element can be hung on the side of the angle.

the boom

The boom is made from a 30-foot length of aluminum irrigation pipe. This pipe should have a diameter of two inches or more. I found the two-inch material fine for a four-element beam on a 20-foot boom (an earlier project), but too flexible for this long beam. However, since it had the advantage of light weight, I strengthened it and found it completely satisfactory.

First of all, I used a four-foot support section made from two pieces of aluminum angle at the center of the boom. The boom is fastened to this mount with three clamps, and the boom-to-mast mount is made from '14-inch thick iron plate fastened to the angle with bolts. Two more muffler clamps are used to fasten the whole affair to the mast. Careful examination of the photos will show the details.



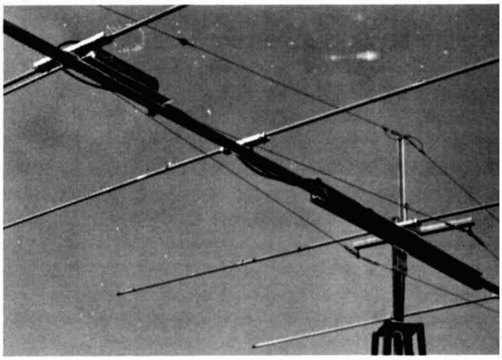
Closeup of the center of the boom showing the bracing wires. struts and turnbuckles.

The elements are fastened to the boom after carefuly checking the spacing. In addition, make sure you have all the elements lined up horizontally with each other. If the array is mounted on a couple of boxes or saw-horses, an inexpensive level can be used to adjust the elements and line them up properly. When all the elements are mounted, the ends of the boom will droop. An upright made from angle iron is mounted at the center of the boom mount. A large eye-bolt is mounted at the top of this upright and two wire braces are run out from it to the ends of the boom. To prevent rusting, the wires should be bronze or aluminum. In addition, they are broken up with small strain insulators to eliminate any resonance difficulties. A small turnbuckle is installed on each wire and adjusted until the boom is straight and level.

There is also a tendency for the boom to whip sideways. This is overcome by another angle brace at the center. This one is bolted at right angles to the boom mount. Four bracing wires are run from it part way out the boom. Four turnbuckles are used to adjust wire tension until the boom is straight and rigid.

When the wires are all taut, the boom is very firm and there is no noticeable flexing. Since I used bronze wire, 1 wrapped heavy vinyl insulation around the wire where it is looped around the ends of the aluminum boom. These two metals set up an electrolytic action, especially in a salt-air atmosphere, which will eventually weaken the beam.

The assembled beam. The driven element is to the left with the gamma-match tuner. The wire bracing provides rigidity.



the remotely-tuned gamma match

The mount for the gamma-match tuner is made from a thin aluminum plate. This plate is mounted on the boom between the driven

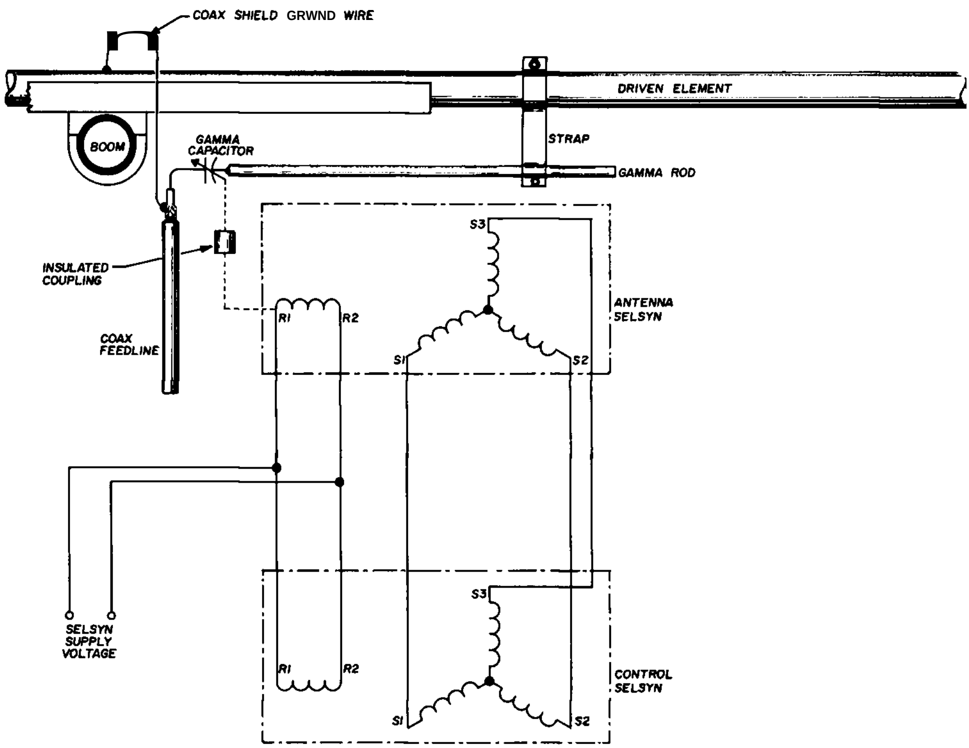


fig. 2. Wiring diagram of the remotely-tuned gamma match and its control system.

element and the first director. The gamma tuner is actually the heart of the beam, but it is very simple to build. It consists of the usual gamma capacitor, driven by a selsyn motor. If you are not familiar with selsyn motors, it is a device which can be remotely controlled through a five-wire cable which is connected to a similar unit. When the system is energized, the shaft of one unit turns until it is in electrical synchronism with the other. Moving one shaft will cause the other shaft to move exactly in step.

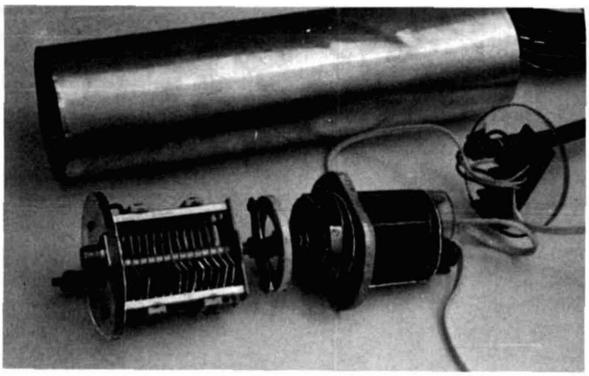
In the remotely-tuned gamma match, the capacitor which is coupled to the shaft of the antenna selsyn will exactly follow adjustments of the controlling selsyn in the shack. By watching an SWR meter in the comfort of your house, you can turn the selsyn shaft and automatically adjust the gamma capacitor on top of the tower. Although it might be simpler to use a small reversible motor for this application, the correct point for the gamma capacitor is very critical. It would be difficult to control a motor accurately enough to adjust the gamma match for an SWR of 1:1. The precise control afforded by the selsyn makes this method much more acceptable.

Ordinary receiver spacing is satisfactory for the gamma capacitor, but the larger spacing will guard against corrosion which may short out the plates after a long period of time. Let the climate be your guide!

To construct the gamma match, I used a pair of surplus selsyns along with a fairly wide-spaced 50-pF capacitor. Suitable selsyns are available from almost any surplus supply house. The antenna selsyn and gamma capacitor were coupled together with an insulated coupling and enclosed in a piece of aluminum vent pipe which was intended for electric clothes dryers. This pipe can be opened for easy installation of the gamma-match components and resealed with the locking lip built into the pipe. Two discs made from 1/4-inch plexiglass are cut to fit snugly into the ends of the pipe. The capacitor must be insulated from the vent pipe, so it's mounted directly to one plexiglass disc. The coaxial input connector is also mounted on the disc. An aluminum strap is used to connect the capacitor to a large bolt which passes through the disc to the gamma rod. Fig. 2 shows the wiring diagram of the gamma match; the photos show its construction.

The second disc is mounted in the opposite end of the vent pipe, and the control cable is passed through it to the selsyn. The selsyn body will usually have some sort of mounting lip which prevents it from fitting snugly into the vent pipe, so small pieces of wood or plexiglass are used to wedge it firmly in place. The selsyn body must be firmly clamped in place so it will rotate the capacitor.

Disassembled remotely-controlled gamma-match tunar.



The antenna and control selsyns are then connected together temporarily and checked. As soon as 110 Vac (or whatever voltage is required by the selsyns used) is applied to their rotors, the two shafts should synchronize immediately. Turning one shaft will cause the other shaft to follow exactly. If the two shafts rotate in opposite directions, reverse the connections to any two stator terminals (marked with the letter "S") of either selsyn. This isn't too important in this application since it doesn't really matter if they turn in opposite directions or not.

When the gamma-match assembly is operating properly, the whole unit should be sealed against moisture by applying a suitable compound to small cracks and openings. I used ordinary rubber-tb-metal cement.

The gamma-match was mounted on my beam with two large surplus Marmon clamps around the vent-pipe enclosure. The bolt ends of the clamps acted as mounting bolts through the plate on the boom. My particular gamma match was fairly heavy because of the large selsyn and capacitor, but any small units may be substituted. Selsyns may be rated for 110, 24, or 36 volts ac, but 110-Vac units eliminate step-down transformers. Use 60-Hertz units to avoid any heating and torque problems. Power is only applied to the units for a short time when tuning up the beam; after that, you only have to adjust the gamma match after a wide shift in operating frequency.

tuning it *up*

Tune-up must be done with the beam at least ten feet above the ground. The gamma match is adjusted by setting the length of the gamma rod to about 20 inches and adjusting the capacitor for minimum reading on the SWR meter. See fig. 3 for the set-up required. The setting of the rod will depend a lot upon the number of elements and their spacing, so it may be necessary to try several taps along the driven element before getting the SWR down below 1:1.5. It should be possible to get it down to 1:1 at a point close to the design frequency, but it will rise as the transmitter is tuned up or down from this point.

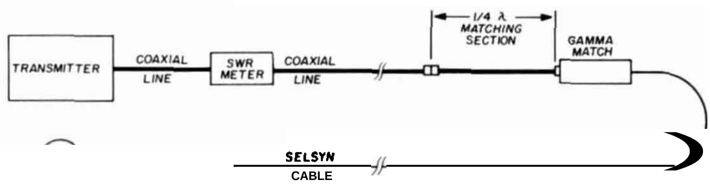


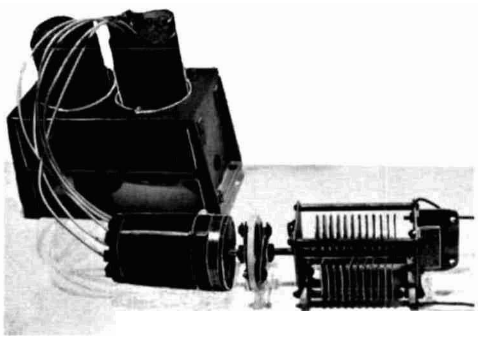
fig. 3. Gamma match adjustment. Tho use of a quarter-wave section of 52-ohm coax is discussed in the text.

You can get a good idea of the bandwidth by plotting the SWR over a band of frequencies without readjusting the gamma match. With the remote match of course, it is no problem to readjust for minimum SWR. Remote tuning is not necessary, but it does add greatly to operating convenience. If it is left out, all adjustments must be made before the beam is in its final position on the tower; otherwise, the capacitor cannot be reached.

Adjustment of element lengths can be done in order to get the best possible gain or front-to-back ratio, but the process is long and tedious and simply not worthwhile. If the elements have been cut and measured properly, they will be very close to optimum dimensions.

The beam may be fed with either 52- or 75-ohm coaxial cable. However, the more elements you use, the lower the radiation resistance at the center of the driven element.

Workbench setup for testing the remotely-controlled gamma-match tuner.



To match the lower resistance, the gamma rod must he tapped farther and farther out on the driven element to obtain a proper impedance match. To overcome this problem, a quarter-wavelength of lower impedance coaxial cable may be used to produce a terminal impedance which is lower than that of either the feedline or the quarter-wave section.

I used 75-ohm feedline in my installation, connected to an 8-foot section of 52-ohm line. This provides an output impedance of about 27 ohms, a much closer match to the beam impedance than afforded by either the 52-ohm or 75-ohm line by itself. The gamma match can easily compensate for any remaining mismatch, and the gamma rod becomes fairly short (and easier to adjust). Although the quarter-wave matching line would normally restrict the beam to a rather small bandwidth, the remote gamma tuning overcomes this problem very nicely. I can adjust my seven-element version for a maximum SWR of 1.7:1 over the range from 28.0 to 29.0 MHz. Over 75% of this range the SWR can be held below 1.5:1.

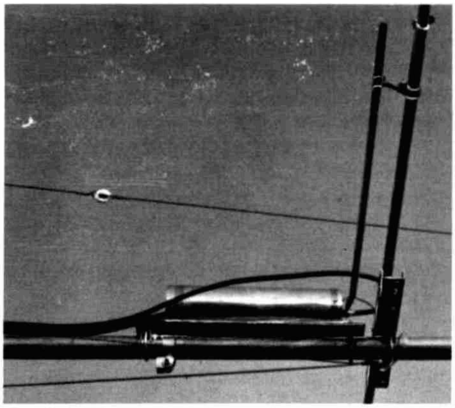
results

The gain and front-to-hack ratio of the antenna are excellent. Calibrated tests indicate a front-to-side ratio of a little over 40 dB, a front-to-back ratio of 31 dB, and two very deep nulls of 60 dB about 10 degrees off the axis of the beam to the rear. I can't explain this latter characteristic, but it makes me very happy in any case! Repeated tests show this peculiar pattern is definitely there, but it may not necessarily be the same on a duplicate beam.

Forward gain figures are not available yet because I have not erected a reference dipole, but based on the other known figures, it should he on the order of 12 dB. On-the-air results have been most pleasing!

ham radio

The gamma-match assembly.



editor's note

Although VE1TG used the remotely-tuned gamma match on a ten-meter beam, the same technique could be used just as well on other antennas with built-in tuning sections.

Normally ham radio will not publish antenna gain figures because of the many vagaries in making accurate measurements. The figures presented here represent an average of many on-the-air measurements and are given only as a guideline.

WIDTY

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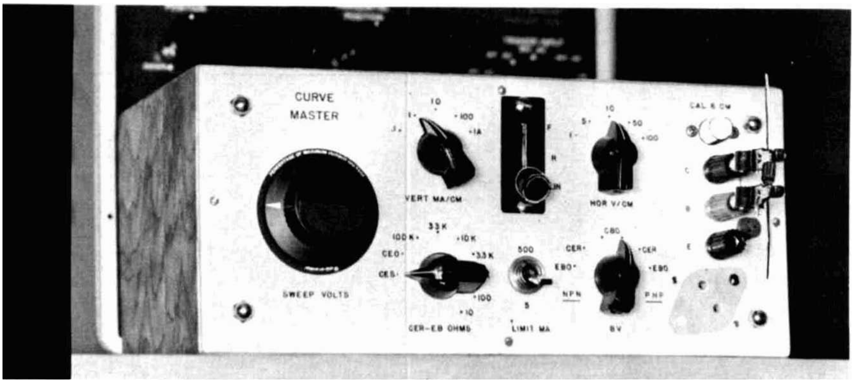


the curve master

An oscilloscope attachment for tracing the characteristic curves of active devices

Tom Lamb K8ERV, 1066 Larchwood Road, Mansfield, Ohio 44907

The curve master is an attachment for an oscilloscope which will display the volt-ampere characteristics of almost any two-terminal device as well as most transistors. As examples, here are some of the devices that can be electrically displayed:

1. Common signal and power diodes (silicon, germanium, selenium, gallium arsenide, etc.)
2. Zener diodes.
3. Tunnel diodes and back diodes.
4. Neon bulbs and V-R tubes.
5. Photocells (photo-resistors).
6. SCR's and light-activated-switches (LAS).
7. Thyrites and thyrectors.
8. Transistors; NPN and PNP, germanium or silicon.

The range of the **curve master** is **1000** volts reverse, and several amperes forward. With a little experience, you can identify a diode or transistor as silicon or germanium by their unique forward-conduction and reverse-leakage characteristics. Since they vary quite widely, semiconductor ratings should always be checked. With a curve tracer you can easily pick out especially good units from your stock for special uses. The **curve master** will check and grade surplus and bargain devices—and may even convince you not to buy any more!

how it works

The basic circuits of a two-terminal curve tracer are shown in **fig. 1.** An ac voltage is applied to the device under test, and at the same time, to the horizontal axis of an oscilloscope. Any current drawn by the device is displayed as a voltage drop on the vertical axis. The horizontal voltage can be increased beyond the device's breakdown point without harm, since the voltage sweeps this region very quickly and the breakdown current is limited by circuit resistance. Since the test voltage goes both positive and negative, both the forward and reverse regions of the device are displayed as shown in **fig. 1A.**

A series diode can be added **(fig. 1B** and **1C)** to eliminate either part of the trace. This permits expansion and detailed examination of the remaining part, and also establishes the start of the trace for voltage measurement reference.

A typical transistor set up is shown in **fig. 1D.** Each transistor junction is checked individually. The **curve master** is made up of these simple circuits with shunts and multipliers to allow a wide range of readings on any scope. It will display voltages from less than one volt to 1000 volts, and current from less than 100 micro-amps to several amperes.

the circuit

The complete schematic of the **curve master** is shown in **fig. 2.** The 750-Vac transformer supplies the forward and reverse test voltages. Diodes D1 and D3 are the blocking diodes described above. S3 selects the region to be displayed—forward, reverse or both. S2 selects the value of the shunt resistance to give vertical ranges of 0.1, 1.0, 10, 100, and 1000 mA per division. S4 has two functions: sections C and D are the horizontal voltage range multipliers, providing ranges of 1, 5, 10, 50, and 100 volts per division; sections A and B select the current-limiting resistances, in two ranges, 5 and 500 mA maximum, as selected by the limit switch S1. S6 selects the transistor junction to be displayed and reverses the test polarities for NPN or PNP devices. S7 adjusts the base-emitter resistance for the BVcer tests. An internal calibrator (S5 and D4) is provided for calibrating the scope.

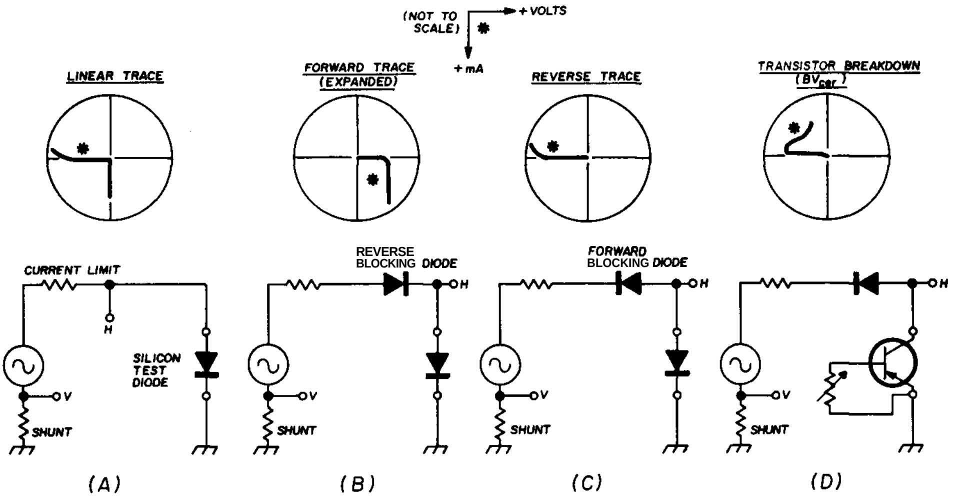
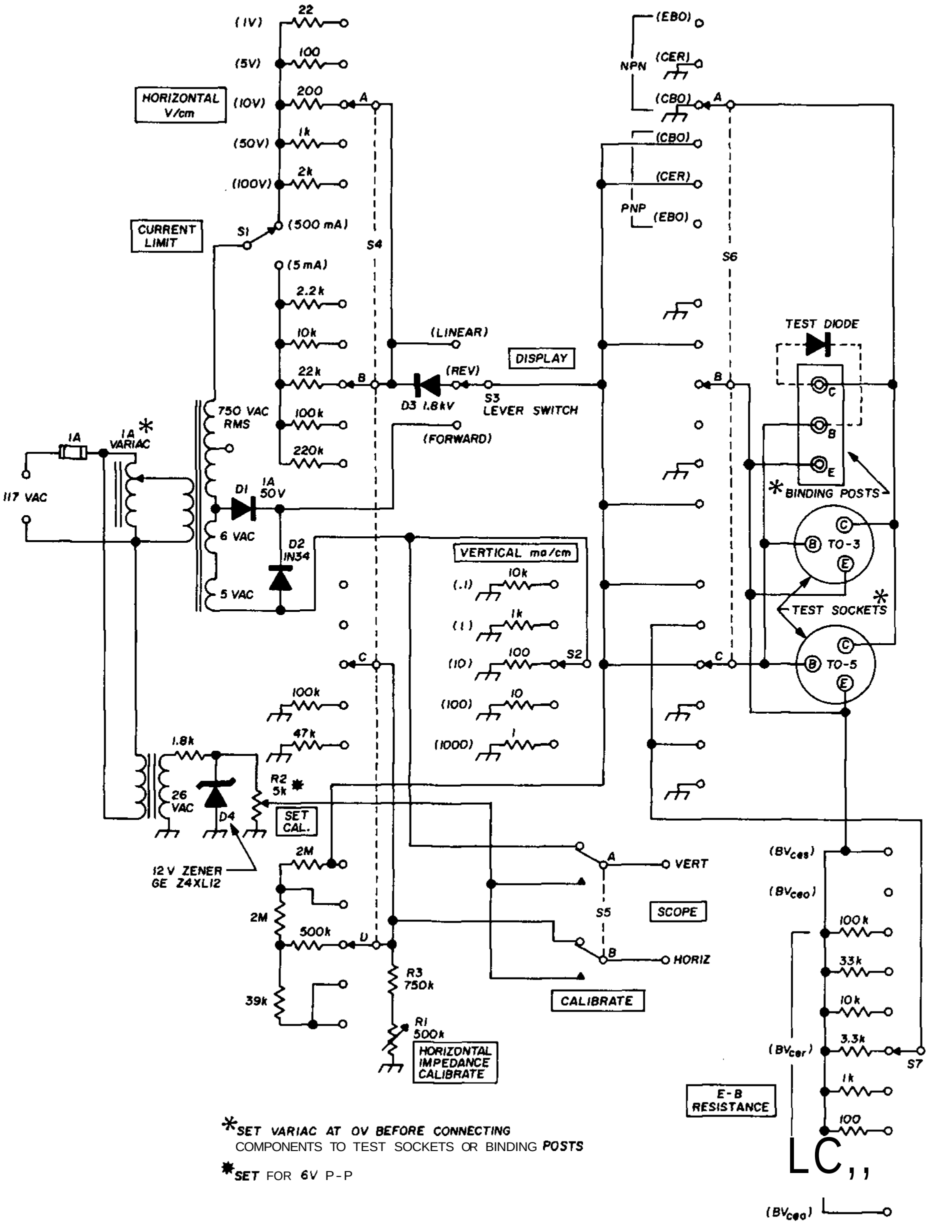


fig. 1. The basic two-terminal curve tracer and several different types of traces which may be obtained with it.

construction

There is nothing critical about the layout. Just remember that up to 1 kV appears across the switches and test jacks, so everything should be well insulated. Diode D3 must be rated above the maximum test voltage and may be made up of several 400- or 600-volt TV-type power diodes in series. D1 is a 50-volt or better diode rated at several amperes. D2 bypasses any reverse leakage from D1. The 750-Vac transformer may be any small receiver transformer with a secondary voltage of at least 750 volts (1 kV peak). All low-voltage windings are series connected to supply the low-voltage, high-current, forward sweep. If the secondary voltage is too low,



S1 **SPDT** toggle switch (Current Limit).
S2 **SP5T** rotary shorting switch (Vert. **mA/cm).**
S3 **SP3T lever** switch (Forward-Revane-Linear).
S4 **4P5T** shorting rotary switch **(Horiz. V/cm).**
S5 DPDT push-button switch (Push to Calibrate Scope).
S6 **3P6T** rotary switch (Breakdown Voltage Circuit).
S7 **SP10T** rotary switch (Base-Emitter Resistor Selection).

fig. 2. Schematic diagram of the **curvemaster.**

use the over-voltage connection on the Variac. The test connections are brought out to five-way binding posts and to both rniniature and power transistor sockets, as shown in the photograph. The 5k set calibration and 500k horizontal impedance pots are screwdriver adjustments mounted on the rear of the chassis.

calibration

The **curve master** can be used with any scope, but for the calibrations to be correct it should meet these specifications:

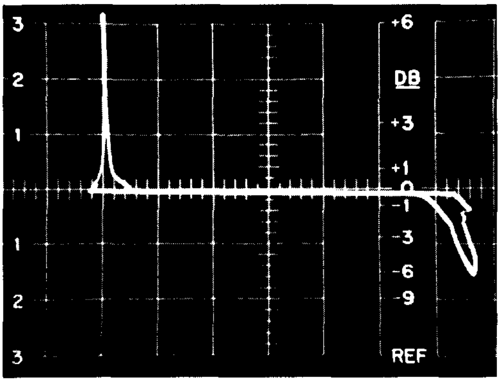
Vertical sensitivity—one volt per major division or better.

Horizontal sensitivity—one volt per major division or better.

Horizontal input impedance—over 500k. The scope does not have to be direct coupled, although this feature is desirable for certain displays.

Only two controls must be calibrated—R1 and R2. R1 sets the horizontal impedance at exactly 500k and is adjusted first. Turn the **Variac** to zero and connect the **curve master** to your scope. Set the scope's vertical gain to zero, the horizontal selector to external and the horizontal gain about half open. Set S3 **(display)** to linear and S4 **(horizontal)** to 1 volt. Advance the **Variac** to obtain exactly full-scale horizontal deflection. Change S4 to 5 volts, and adjust R1 for exactly 1/5-scale deflection. If your scope's horizontal impedance is not around 1 megohm, it may be necessary to change the value of R3 to obtain 1/5-scale deflection. The horizontal multiplier, S4, is now adjusted. This setting must be checked if the **curve master** is used with another scope.

fig. 3. OA2 curve. 50 V/cm H, 10 mA/cm V.*



* 50 V/cm H, 10 mA/cm V indicates the curve-master is set for 50 volts per centimeter horizontal and 10 mA per centimeter vertical on the oscilloscope. The graticule marks on the oscilloscope shown in the photographs are one centimeter apart. The curve to the right of center is the forward characteristic; to the left, reverse. In some cases different sweep widths are used for the forward and reverse traces.

The **curve master** has an internal calibrator to set up the scope gain controls. This calibrator must be set for exactly 6 volts, peak to peak. If you have a calibrator for your scope, it is still convenient to use the **curve master's.** R2 may be set for a 6-volt calibrating waveform in two ways. If you have a calibrator or a scope with a calibrated vertical amplifier, adjust R2 as described in the next paragraph.

Calibrate the scope tor 1 volt per major division. Connect the **curve master** and press S5 **(calibrate).** A sloping line will appear on the screen. Adjust R2 for a vertical distance of exactly six major divisions.

If no external calibrator is available, a good 0A2 voltage regulator tube can be used. Set the scope vertical gain to roughly 1 volt per division. Connect the **curve master** to the scope. Attach the collector jack to pin 7 of the 0A2 and the base jack to pin 1. Set the switches as follows: S1 to 500 mA, S2 to 10 mA/division, S3 to reverse, S4 to 50 volts per division, S6 to PNP-CBO. Advance the **Variac** and the horizontal gain control on the scope to form a trace like the left half of **fig.** 3. The right edge of the base line will be a bright spot. Using the horizontal gain control, set the vertical part of the trace exactly three divisions from the trace's right edge. The horizontal system is now set at 50 volts per division. Turn the vertical gain and the **Variac** to zero, push S5 and adjust R2 for a six-division horizontal deflection. The instrument is now calibrated.

using the curve master

Caution: dangerous voltages may be present at the test terminals and sockets. Always turn the **Variac** to zero before making any connections. It is a good idea to start all tests with the limit switch in the 5-mA position to prevent possible damage to the device being tested.

Let's get used to the operation by checking several devices in some detail, using the photos as examples.* Connect the **curve master** to the scope and set it for external

*These photos were taken with a Polaroid Pacemaker, 3000 speed film, and both plus-4 and plus-2 close-up lenses. The traces were displayed on a Knight KG-2000 scope, using an exposure of 1/4 second at f8.

horizontal input. Push the calibrate switch (S5), and set both scope gains for six divisions of horizontal and vertical deflection.

Connect an NE-2 or other small neon bulb to the collector and base terminals. Two-terminal devices are always connected to these jacks and the BV switch set on CBO-PNP. Changing to CBO-NPN merely reverses the test device's connections and may be switched to reverse a mis-connected device or to examine its reverse characteristic. Since a neon bulb operates at about 60 volts and 1 mA, set the Vertical mA per division switch (S2) to 1, and the Horizontal volts per division switch (S4) to 50. The limit switch (S1) must be in the 5 mA position. Set the display switch (S3) to linear, and advance the Variac to obtain the trace shown in fig. 4 which is displaced below center for a clearer photograph.

Assume the sweep voltage from the transformer is zero, and increasing positively. The trace will start in the center of the screen and move to the right along the baseline. Until the voltage increases to about 75 volts, the neon draws no current and there is no vertical deflection. At 75 volts, the bulb fires, suddenly decreasing the voltage and increasing the current throughout the bulb. The trace moves (too quickly to photograph) down and to the left, joining the lower branch. The bulb is now lighted and conducting the current permitted by the circuit resistance. As the sweep voltage increases, the current sky-rockets downward (positive), reaching the maximum allowed by the Variac setting. In fig. 4 the maximum current is 1.8 mA.

The sweep voltage now decreases. The trace moves back along the same path until the current reaches zero at 55 volts, and the bulb extinguishes. On the reverse cycle the same events occur except that now the second electrode is lighted, and its characteristic is traced. The two electrodes' characteristics are similar but may not be identical. The curve master can compare these two characteristics.

fig. 4. NE-2 curve. 50 V/cm H, 1 mA/cm V.

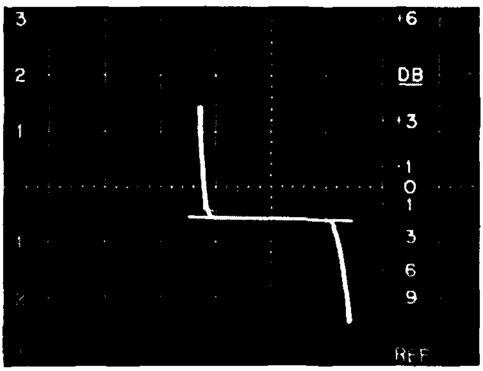
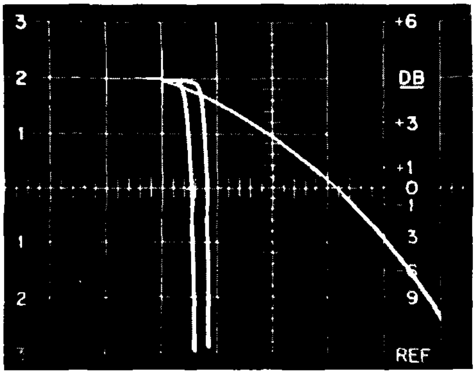


fig. 5. Left to right: high-conductance germanium diode, silicon diode, low-conductance germanium diode. 1 V/cm H, 10 mA/cm V.

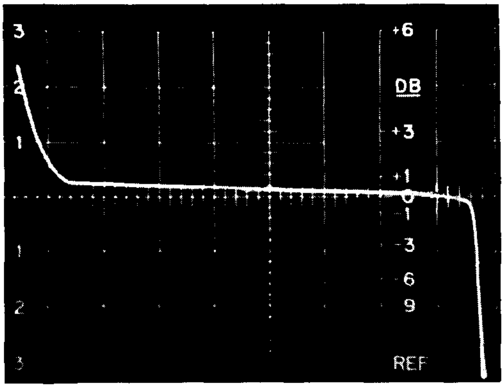


Now change the display switch to reverse, so only the left half of the trace is displayed. The zero-voltage, zero-current point is now the bright dot at the right end of the baseline. This point can be kept in the center even on an ac-coupled scope. Only the second electrode, being displayed, is lighted. If the BV switch is changed from CBO-PNP to CBO-NPN, the first electrode will light and be displayed, but flipped to the screen's left—exactly in place of the previous trace. The two electrode characteristics can be compared by super-position if the BV switch is rapidly alternated between PNP and NPN.

Let's look again at the V-R tube. Connect an 0A2 as described in the calibration section. Use the linear display, 500 mA limit, 10 mA per division vertical and 50 volts per division horizontal to obtain a trace like fig. 3. The left-hand portion of the curve is the normal regulation connection, showing very little voltage variation from 8 to 30 mA. A backward-connected V-R tube would give very little regulation, as shown by the right half of the trace. Try flipping the BV switch to flip the trace. Now try it with the display switch in the reverse position. For an expanded view of the regulating region, set the horizontal switch to 10 volts per division and re-position the trace on the screen.

Now we will look at a typical germanium diode, such as a 1N34. Turn the **Variac** down, the limit to 5 mA, the vertical to I mA per division, and the horizontal to 50 volts per division. Advance the **Variac** to obtain the trace shown in **fig. 1A** using a linear display. Now, change the display to forward and reduce the horizontal to 1 volt per division. The expanded forward characteristics will look like one of the outer traces of **fig. 5**, depending on the particular diode used. Notice that both low and high conduction diodes show conduction starting at about 0.2 volts. The center curve is typical of silicon diodes, which begin conduction at about 0.6 volts. These unique forward curves can be used to identify the type of diode that you are testing.

fig. 6. T1 U213 silicon diode. Forward: 1 V/cm H, 100 mA/cm V; reverse: 10 V/cm H, 10 mA/cm V.



typical curves

Fig. 6. and 7 are double exposures with the forward and reverse curves photographed separately at different settings. The reverse curve of a silicon TV type power diode, the TI U213, is shown in **fig.** 6. Notice the fairly soft breakdown at 660 volts, well beyond the 400-volt rating. Compare this with the very sharp zener diode break in **fig.** 7. The zener diode is designed to operate in the reverse region, the power diode is not. An old-fashioned selenium rectifier stack curve is shown in **Fig. 8.** Compare the high reverse leakage and high forward-voltage drop with the silicon unit.

Now, let's try some transistors. The important voltages here are:

BVcbo Breakdown voltage between collector and base with the emitter open.

BVebo Breakdown voltage between the emitter and base with the collector open.

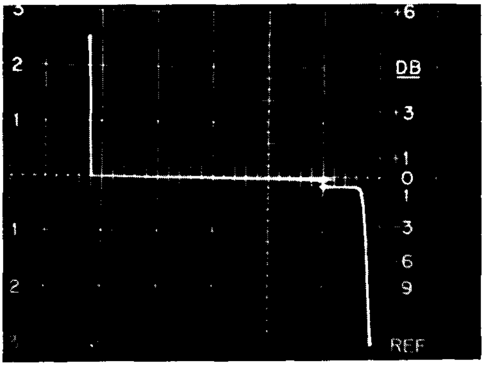
BVces Breakdown voltage between collector and emitter, with the base connected to the emitter.

BVcer Breakdown voltage between collector and emitter, with resistance between base and emitter.

BVceo Breakdown voltage between collector and emitter, with the base open.

For discussions of the exact meanings and uses of these ratings, see the General Electric

fig. 7. Mallory ZA51 zener diode. Forward: 1 V/cm H. 100 mA/cm V; reverse: 10 V/cm H, 10 mA/CM V.



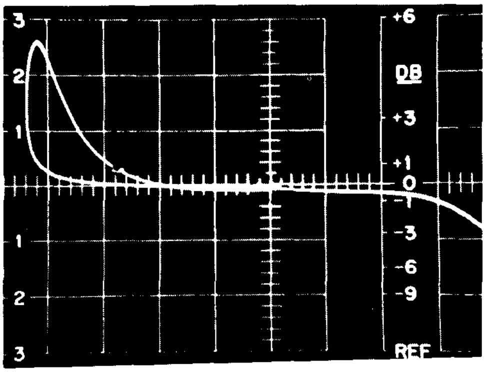
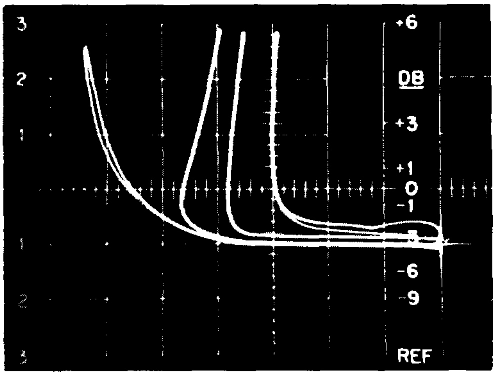


fig. 8. Selenium rectifier. Note relatively poor forward and reverse characteristics. Forward: 1 V/cm H, 100 mA/cm V; reverse: 50 V/cm H, 10 mA/cm V.

Transistor Manual or the Motorola Power Transistor Handbook.

First, a power transistor. Start with these settings: display, reverse; limit, 500 mA; vertical, 10 or 100; horizontal, 50 or 100. Set S7 to BVces and S6 to BVcbo-PNP. Plug a power transistor into its socket and advance the **Variac.** Don't forget to connect the collector to the socket with a screw. You should see the left-hand curve of **fig. 9.** If you only get a vertical trace, either the collector-base junction is shorted, or you have an NPN unit. Try the NPN position of S6. Most power transistors are PNP.

fig. 9. 2N554 power transistor. Left to right: BV_{cgo}, BV_{ces}, BV_{cer} (100 ohm), BV_{cco}. 10 V/cm ti, 10 mA/cm V.

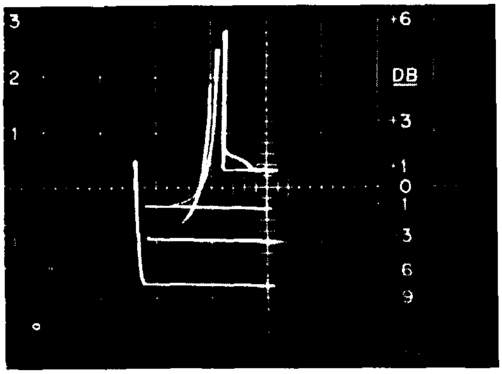


The collector diode rating, BVcbo, is the voltage at the vertical part of the curve—in this case over 65 volts. Not bad for a 2N554, rated at 15 volts! Now turn S6 to BVcer. In this position the collector-to-emitter characteristic is traced, with the base-to-emitter resistance selected by S7. When S7 shorts the base to the emitter, you have BVces. When S7 opens the base, you have BVceo. In other positions of S7 you have BVcer, with S7 setting the value of R. For transistor quality checks, the value of R from the manufacturer's data sheet for BVcer should be used.

For circuit design, set S7 to the value to be used and check the actual curve for the transistor. BVebo is not shown, but will be displayed with S6 set to that position. This is the maximum reverse input (base-to-emitter) voltage that may be used if not already limited by BVcbo. BVebo is important in reversed-bias applications such as blocking oscillators, where a very large reverse-voltage spike can occur.

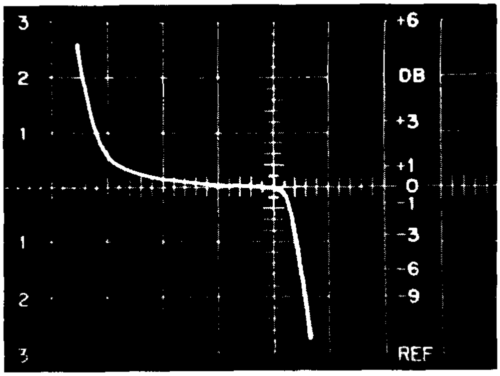
fig. 10. 2N2924 silicon transistor. Top to bottom: BV_{ces}, BV_{cro}, BV_{cer} (33k), BV_{cer} (10k), BV_{cbo}. 50 mA/cm H, 1 mA/cm V.

Traces displaced from origin for clarity.



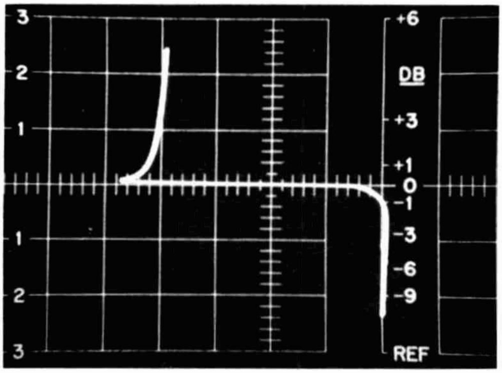
A small-signal silicon planar transistor exhibits slightly different curves. The GE 2N2924 NPN device is traced in fig. **10.** Notice the sharp breakdowns in the two BVcer tests. The lowest break is at 40 volts, while the transistor is only rated at 25 volts. Don't forget to use the 5-mA limit with low-power devices.

Fig. 11. 1N34 germanium diode. Forward: 1 V/cm H, 1 mA/cm V; reverse: 50 V/cm H; 1 mA/cm V.

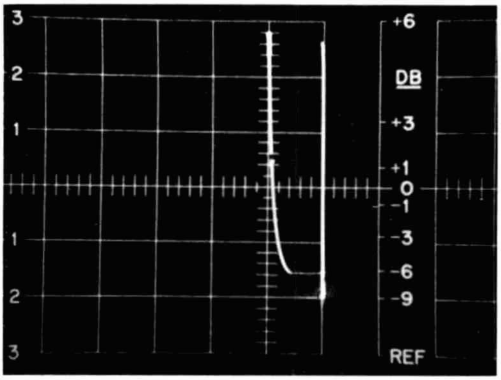


The rest of the photos show traces of less common devices. If an external bias supply is used, the complete collector family curves can be displayed on a dc-coupled scope. With external bias and filament supplies, it is possible to display vacuum and gas-tube family curves as well.

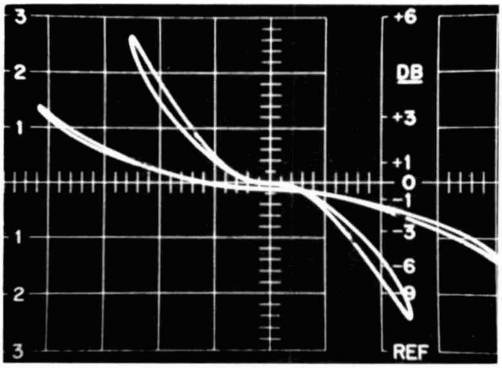
ham radio



10 V/cm H, 1 mA/cm V.



1 V/cm H, 1 mA/cm V.



100 V/cm H, 1 mA/cm V.

Top to bottom: TI-42 trigger diode. 1N653 tunnel **diode,** photocell showing nonlinear resistance at different light levels.

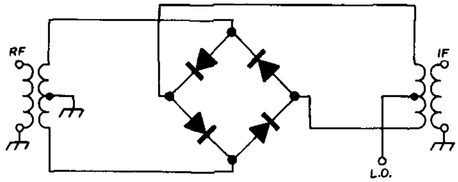


fig. 1. Basic double-balanced mixer circuit. Note that the diodes are not connected in a diode bridge.

double-balanced mixers

how to use this fascinating new electronic stuntbox in amateur equipment

Jim Fisk W1DTY, RFD 1 Box 38, Rindge, New Hampshire 03461.

If you are tired of fooling around with transistors and bored by integrated circuits, you might be interested in a new electronic stuntbox that is currently on the market—the double-balanced mixer. These new gadgets are small (the unit I have in my hand is half an inch wide, $^3/_8$-inch high and about $^7/_8$-inch long), broadband, typically from 500 kHz to 200 MHz, and can perform all kinds of electronic tricks. To say that they are versatile is the understatement of the year.

Broadband double-balanced mixers can be used as suppressed-carrier modulators, deriving sum and difference products of two input frequencies; they can serve as up- or down-converters with appropriate filtering at the output, and they can be used as product detectors, phase detectors or voltage-controlled attenuators. In addition, they may serve as a-m modulators, pulse modulators or spectrum generators. To top it all off, they don't require a power supply. Interested? Read on.

While these broad-band double-balanced mixers have been available for several years, up until now their cost has been so high that they were completely out of the amateur picture. Although they are still quite expensive, the price has dropped to the point where many amateurs will be trying them in their projects.

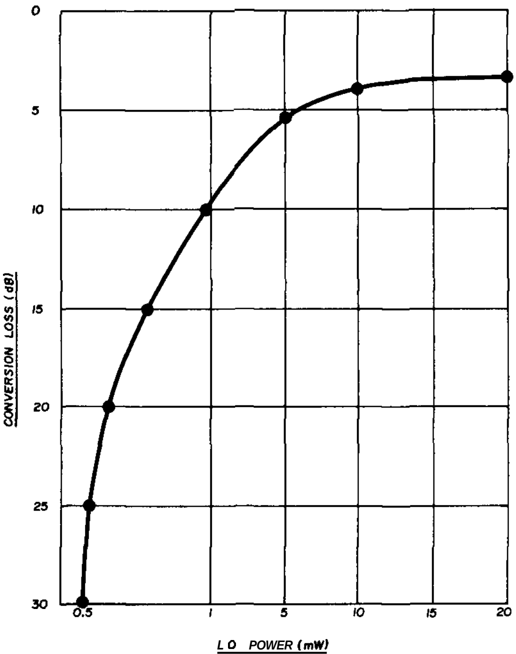
circuits and characteristics

The units presently available actually come in all shapes and sizes, from miniature units for printed-circuit mounting to larger units with built-in coaxial connectors. However, they all have one thing in common — their circuit. Evidently the manufacturers are afraid someone is going to copy their designs, because most of the units are hermetically sealed. However, the circuit shown in **fig. 1** is probably what you would find if you opened the thing up.

Schematically, it isn't that *impressive* — two transformers and a diode ring. Each of these components must work together to provide the desired isolation, frequency range, and conversion loss. The diodes aren't any garden variety units — often they're Schottky barrier diodes; and they are very closely matched. The transformers, usually toroidally wound, must be precisely balanced to obtain the desired operating characteristics. All these things contribute to the cost. For example, in a typical double-balanced mixer, up to 50 MHz the isolation between the rf and i-f inputs is 25 dB, isolation between rf and local oscillator is 40 dB, and isolation between the local oscillator and i-f is 35 dB.

Internal construction and engineering is also important to the frequency characteristics and conversion loss of the unit. The frequency range varies from manufacturer to manufacturer, as shown in table **1**. These are the fre-

fig. 2. Conversion loss versus local-oscillator power for the Ultramatic Systems UM-1 double-balanced mixer in a 50-ohm system.



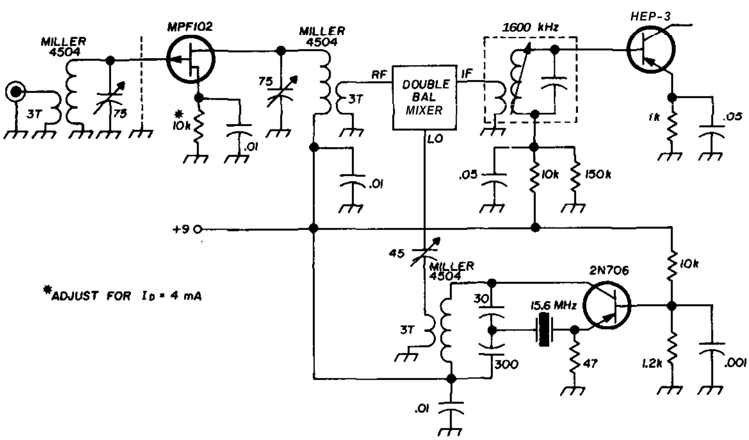


fig. 3. A 20-meter converter using the double-balanced mixer. The 75-pF capacitors in the rf stage may be ganged and tracking obtained by adjusting the two slug-tuned coils.

table 1. Double-balanced mixers suitable for amateur use.

Manufacturer	Model No.	Frequency	Insertion Loss	LO Isolation	Noise Figure
Adarns-Russell'	MLFJ	.5-200 MHz	8 dB	25 dB	—
Adams-Russell	MHF-3	10-500 MHz	8 dB	25 dB	—
Comdel[2]	CM101	1-30 MHz	6 dB	30 dB	6.5 dB
Comdel	CM102	20-120 MHz	7.5 dB	30 dB	8.0 dB
Comdel	CM103	2-200 MHz	7.5 dB	30 dB	8.0 dB
Ultramatic Systems[3]	UM-1	.2-200 MHz	6 dB	40 dB	—

1. Adams/Russell-Anzac, 121 Water Street, Norwalk, Connecticut 06854.
2. Comdel, Beverly Airport, Beverly, Massachusetts 01915.
3. Ultramatic Systems Laboratory, P. O. Box 2143, Sunnyvale, California 94087.

quencies at which the unit is guaranteed to work within the limits of the other specifications. Usually they may be used at somewhat lower frequencies. For example, if the low-frequency limit is listed as 1 MHz, in all probability the unit will work satisfactorily down to several hundred kilohertz. Also, the i-f terminal is usually dc connected; this is very important. In some units you will find a series capacitor between the external terminal and the transformer center tap. With a straight dc connection as shown in **fig. 1**, the versatility is increased tremendously. Applications as a voltage-controlled attenuator, a-m modulator, phase modulator and phase detector are impossible without it.

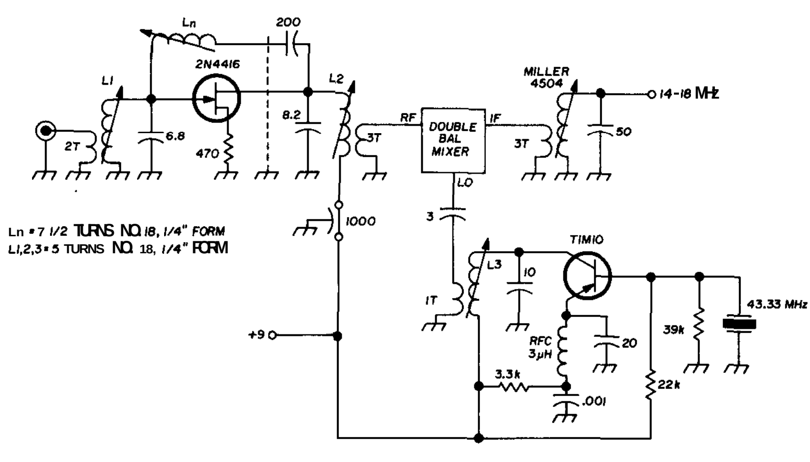
If you look closely at the specifications for these units, you'll find that most of them are designed for 50-ohm systems. However, this doesn't mean that the input and output impedance have to be precisely matched. The major effect of increasing the load impedances will be to raise the low-frequency cutoff of the device.

As shown in **fig.** 2, the conversion loss of the double-balanced mixer is very dependent upon the local oscillator drive that is supplied. This curve is typical of the Ultramatic Systems UM-1 double-balanced mixer, but units from other manufacturers have similar curves. In this case, when the local oscillator power is less than about 5 mW, conversion loss increases very rapidly. As local-oscillator power is increased above 5 mW, conversion loss does not decrease significantly. For most circuits, optimum operation will occur somewhere in the neighborhood of 5 to 10 milliwatts of local-oscillator drive.

frequency converters

One of the most important applications for double-balanced mixers is as a mixer or frequency converter stage in a communications receiver. In this case **(fig. 3)**, the rf

fig. 4. The 144-MHz converter with a double-balanced mixer stage. For more gain, another rf stage may be added.



signal from the first rf amplifier is fed into the rf terminal, the high-frequency oscillator is connected to the local oscillator terminal, and the i-f output is taken off the i-f terminal. One of the big advantages of this circuit is that very little local oscillator power reaches the i-f terminal — it is very highly suppressed. In the circuit in **fig. 3**, about 6 mW of power at the local oscillator terminal provided the best results; this is adjusted by the 45-pF padder in series with the local oscillator link.

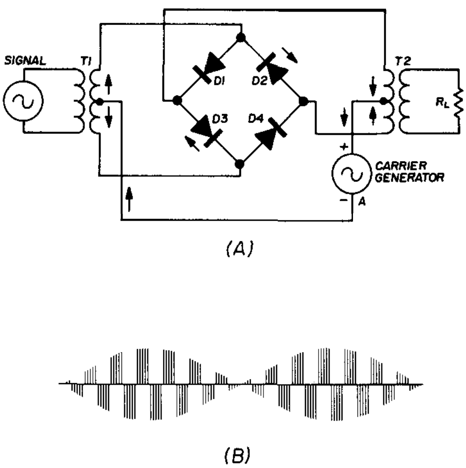
Another very important characteristic of the double-balanced mixer is reduction of spurious mixer products which often result in birdies. If, for example, you were to compare the spurious mixing products from the first six harmonics in a standard single-diode modulator, you would find no less than twenty-six separate spurious internal signals that could (and would) cause birdies. The standard series two-diode mixer that is often used in amateur equipment helps some — it only generates eleven spurious signals. The double-balanced mixer only generates five.

Although sheer numbers of internal spurious signals provide a certain guideline, what is more important is their magnitude. In both the single-diode and series two-diode mixers, there are first magnitude signals that can cause birdy problems. In the double-balanced mixer the strongest spurious signal is of the third order.

Another application for the double-balanced mixer is the 144-MHz converter shown in **fig. 4**. If you break this circuit down into its individual parts, it's not much different from **fig. 3**. Although different transistors are used, the main differences are in the tuned-circuit values. Similar converters could be built for any of the amateur bands from 160 meters through 432 MHz or so, depending on the frequency characteristics of the unit you are using.

Double-balanced mixers are also highly desirable in the frequency-converter stages of single-sideband transmitters. Their high dynamic range, carrier suppression characteristics and low spurious output provide clean ssb signals. Also, third-order intermodulation distortion products are typically suppressed by more than 50 dB.

fig. 5. Since the currents which pass through the transformers of the double-balanced mixers are 180° out of phase, they cancel out. The output waveform is shown in B.



theory of operation

Before we discuss any further applications of the double-balanced modulator, it might be interesting to find out how the thing really works. Consider the circuit of **fig. 5** with the signal voltage disconnected; with only the output of the carrier generator connected to the double-balanced mixer, there is no output across the load resistor R_L. When the A side of the carrier generator is negative, current flows through T1, diodes D1 and D4, transformers T2 and back to the generator as shown by the arrows. Since the currents on each side of center tap are 180° out of phase, they cancel, and there is no output. When point A is positive, current flows through T1, diodes D2 and D3, and transformer T2, again with no output.

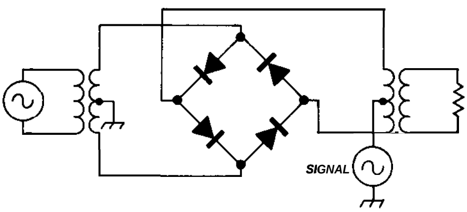
If a signal is applied to the primary of T1, the currents in each half of the secondary are unbalanced, and an output voltage appears across the load resistor R_L. In actuality, the carrier generator switches the signal voltage on and off — nearly rectangular pulses controlled by signal amplitude are found across the load resistor at the output. To obtain good switching efficiency and maintain minimum distortion, the frequency of the carrier signal is usually many times greater than the frequency of the signal at the rf terminals.

With a properly designed and constructed double-balanced mixer, carrier suppression on the order of 40 dB is not difficult to obtain. With this type of mixer, third-order distortion products are typically suppressed by 50 dB. Since the even harmonics are inherently suppressed by the operation of the circuit, the only spurious signals we have to contend with are those created by odd-numbered harmonics. In other types of mixers we have to contend with even-numbered harmonics as well.

An interesting modification to the basic double-balanced mixer circuit may be obtained by exchanging the carrier and signal generators as shown in **fig.** 6. In this case, when the carrier is positive on the upper side of the input transformer T1, the signal passes through the upper half of T2. When the carrier goes negative, the signal passes through the lower half of T2. The output waveform with these connections is exactly the same as shown in **fig. 5B,** with the signal current being switched back and forth through the output transformer by the carrier.

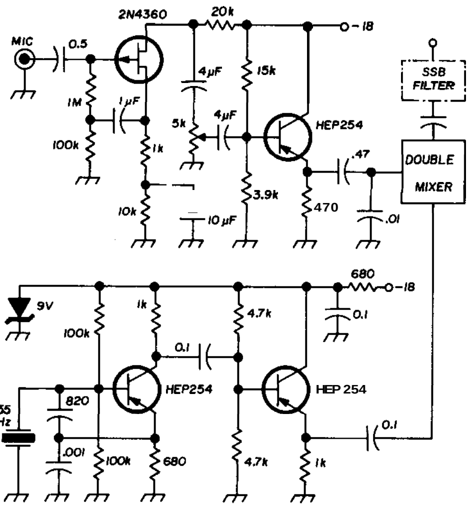
There are several advantages to be gained when the double-balanced mixer is operated in this manner. First of all, in many presently available commercial units, the input transformer T1 cannot be used efficiently below about 500 kHz. If only the standard connection shown in **fig. 5A** were available, this would limit the unit's use as a sideband generator since our signal is an audio voltage.

fig. 6. Alternate connection to the double-balanced mixer. The output waveform is the same as shown in fig. 5B.



Since the carrier input is typically characterized with a dc to 200 or 300 MHz bandwidth, it is perfectly suitable for audio or video inputs. (Has anybody tried ssb for ATV video transmission?) With the alternate connection, the audio driver may be single ended. Also, it eliminates the need for balancing capacitors across the input transformer that are needed in other circuits.

fig. 7. A 455-kHz ssb generator using the double-balanced mixer. With proper shielding and grounding, carrier suppression of this circuit will exceed 50 dB.



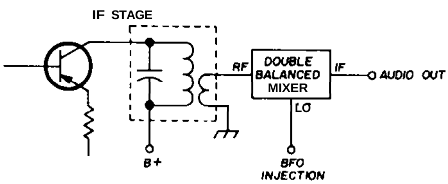
sideband generator

One of the most appropriate applications for the double-balanced mixer is in a sideband generator. This is where it really shines. There are no touchy controls to adjust and no power connections. Just connect local-oscillator power, audio and presto — double sideband! With the filter across the output to remove one of the sidebands, you have a clean single-sideband signal with excellent carrier suppression.

At the input side of the filter, with good construction, grounding and shielding, carrier suppression of the double-sideband signal will be on the order of 40 dB. With the additional suppression afforded by the sharp skirts of the filter, carrier suppression of the ssb signal should be well in excess of 50 dB.

A ssb generator circuit using a double-balanced mixer is shown in **fig.** 7. This circuit is straight forward, small and performs very well. It could easily be built on a small printed-circuit board two inches square. By adding another crystal in the local oscillator circuit with a switch, selectable sideband would be possible at little extra cost. For proper operation, the local oscillator should provide about 10 mW of power at the local-oscillator terminals (2 V p-p into 50 ohms).

fig. 8. A product detector incorporating a double-balanced mixer may be added to most receivers.



For maximum linearity, the audio signal should not exceed 1 mW or 225 mW rms.

product detector

Another application for the double-balanced mixer is as a product detector. Although there is no significant advantage to using a double-balanced mixer as a product detector, it is an interesting application for the device. The product detector circuit shown in **fig. 8** requires a minimum of components, but does an excellent job. It has very high dynamic range and does not overload with large input signals. The local oscillator power should be from 5 to 10 mW, and the rf signal, about 1 mW.

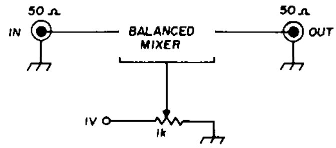
voltage-variable attenuator

Here is a very useful job for the double-balanced mixer. As an attenuator, it may be used with signal generators, in the input to your receiver, or as a handy gadget around the shop. All that you need is a variable source of control voltage from zero up to about 1 volt, either positive or negative. As the voltage is varied, the attenuation through the double-balanced mixer will vary from 2 to 40 dB. Make sure that the input current to the device does not exceed 60 mA. More current than this will irreparably damage the diodes. The easiest way to limit the current is to put four series-connected forward-biased silicon diodes across the voltage source. They will limit the input voltage to approximately one volt; the 60 mA point will be above this.

summary

These applications are just typical of the many things that can be done with these new double-balanced modulators. In addition, of course, they may be used in a-m and pulse modulators, spectrum generators and phase detectors. If you have considered building a frequency synthesizer or a phase-locked receiver, the double-balanced modulator phase detector is an ideal choice.

fig. 9. This voltage-variable attenuator will attenuate the input signal from 2 to 40 dB as the control voltage is varied from zero to 1 volt. The input current should be held below 60 mA to prevent damage to the double-balanced modulator.



For circuits that use transistors and integrated circuits, the small size of the double-balanced mixer is ideal. When you go to buy one, they may seem a little expensive, but in terms of carrier suppression, dynamic range and spurious outputs, their performance is pretty hard to beat.

ham radio

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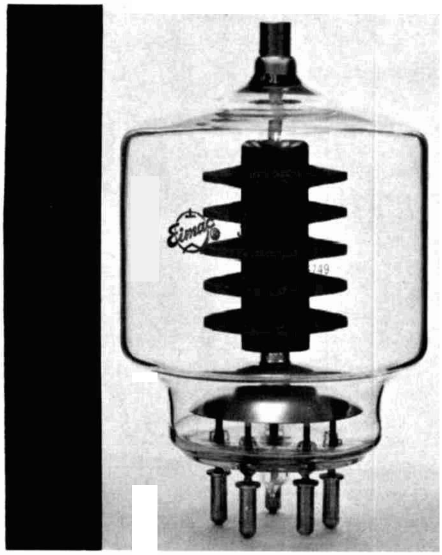


the 3-500Z in amateur service

Here's a new zero-bias triode from Eimac that features increased plate dissipation

William I. Orr W6SAI, Eimac Division of Varian, San Carlos, California 940

The 3-5002 is a heavy-duty power triode of 500 watts plate dissipation. It is exceptionally well suited for use as a class-B amplifier in rf or audio application. It may be used in zero-bias linear-amplifier service at plate potentials up to 3000 volts, eliminating bulky and expensive screen and bias power supplies.

Of particular interest to the radio amateur is the use of the **3-5002** as a grounded-grid (cathode-driven) amplifier for ssb service. One **3-5002** is capable of a PEP input of over 1100 watts, requiring only **30** watts PEP drive power. Intermodulation distortion products at this power level are **30** dB or more below one tone of a two-tone test signal. At 2000 volts, moreover, over **500** watts of power output are obtainable with distortion products better than **38** dB below one tone of a two-tone signal. Typical operating characteristics for the 3-500Z are listed in table 1. A data sheet covering operation of the 3-500Z may be obtained at no cost by writing to me.

In cases requiring additional plate dissipation, the 3-500Z may replace the **3-4002.** The forced-air requirements for the two tubes are approximately equal, and a blower capable of **13** cubic feet per minute at a back pressure of 0.2 inch is satisfactory for a single 3-500Z. (Use blower size #3 at 1600 rpm. For two 3-500Z's, use blower size #3 at 3100 rpm, or size #2½ at 6000 rpm.)

The zero-signal plate current of the 3-5002 is somewhat higher than that of the 3-4002. When the 3-5002 is used to replace the 3-400Z, a means of reducing the zero-signal plate current is recommended, particularly if the equipment is power-supply limited. Only a few volts of bias from a low impedance source are required. The simplest way of obtaining well-regulated bias voltage is to place a zener diode in the filament return circuit of the 3-5002 **(fig. 1).**

The 1N4551 zener diode has a nominal voltage drop of 4.7 volts and an impedance of 0.1 ohm, making it ideal for this service. At this value of bias, the zero-signal plate current of the 3-5002 at a plate potential of 3250 volts is reduced from 160 to approximately 90 milliamperes.

The zener diode may be bolted directly to a cool area of the chassis which will act as a heat sink. Additional VOX-selective bias may be placed in series with this zener diode to reduce standby current of the 3-5002 to nearly zero in order to eliminate "diode noise" during reception and conserve standby power **(fig.** 2).

the grid-current meter

It is advisable to monitor the grid current of the 3-5002 as an indicator of correct drive and antenna loading. Too much grid current indicates underloading or overdriving and too little grid current indicates underdriving or overloading, other things being equal. As the grid must be held at rf ground, the grid meter must be introduced in such a manner as not to disrupt this circuit. A simple grid meter scheme is shown in **fig. 1**. Each grid pin is grounded through a .01-μF mica capacitor paralleled with a 3.3-ohm, 2-watt composition resistor. A small dc voltage drop exists across the resistor under normal tube operation. The voltage drop is read by a simple dc voltmeter (M1) calibrated in terms of grid current.

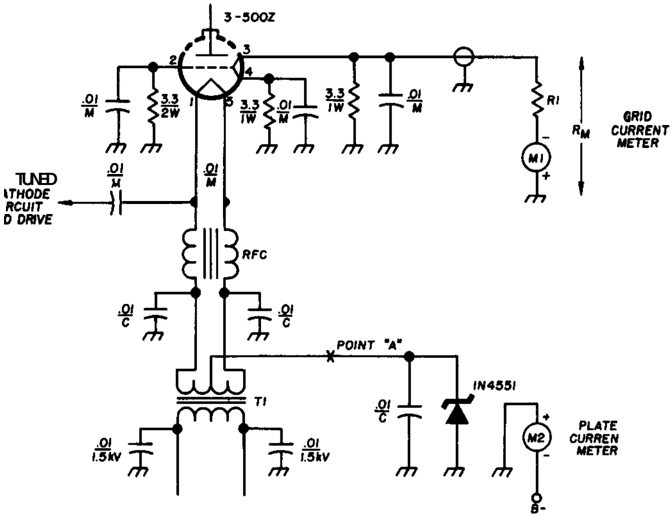


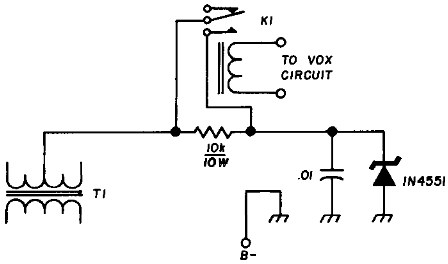
fig. 1. Zener diode bias circuit for the 3-5002. A 1N4551, 4.7-volt, 50-watt saner diode provides cathode bias for tha 3-5002. Meter M1 (0-1 mA dc) reads grid current of the tube in terms of tha voltage drop across the three grid resistors. Meter M2 reads plate current. The multiplier resistor plus internal resistance of meter M1 should total 220 ohms. Grid and filament bypass capacitors are 600-volt mica units (M). Other bypass capacitors are ceramic discs.

In the example shown, it is desireable that the grid meter have a full-scale indication of 200 milliamperes. The dc grid-to-ground resistance is about 1.1 ohm, and, at a current of 200 mA, a voltage drop of 0.22 volts will be developed. The 0-1 dc milliammeter is converted to read 0.22 volts full scale by the inclusion of a series multiplier resistor. The sum

tabla 1. Typical operation of the 3-5002 in grounded-grid rf linear-amplifier service.

DC plate voltage	3000	2500	2000	V
Zero-signal dc plate current	160	130	95	mA
Single-tone dc plate current	370	400	400	mA
Single-tone dc grid current	115	120	130	mA
PEP input power	1110	1000	800	W
PEP useful output power	750	600	500	W
Resonant load impedance	5000	3450	2750	ohms
Cathode input impedance	115	100	100	ohms
Intermodulation products (3rd order)	—30	—33	—38	dB

fig. 2. VOX-selactiva cutoff bias circuit. Additional cutoff cathode bias is added by the VOX relay to reduce standby plate current to near-zero, eliminating "diode noise" in a nearby receiver. The bias is added at point A in fig. 1.



of the resistor plus the meter resistance should total 220 ohms.

3-500Z circuitry

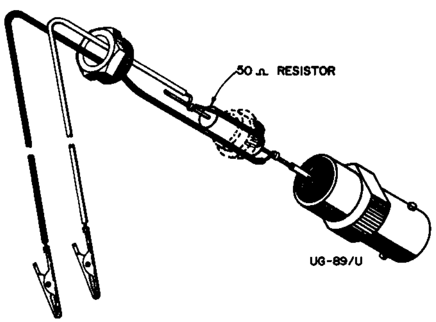
No specific circuits are shown for the 3-5002, since published 3-4002 circuitry applies equally well to this tube. Two 3-500Z's may be used in place of a single 3-1000Z with appropriate corrections in air flow, filament power requirements and zener bias (if necessary).

ham radio

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.

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 ½-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.



Exploded view of the diode mount. The two connector nuts and button capacitor are soldered into one unit. The method of bending the capacitor tabs is also shown.

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 'Id-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

If the DAVCO DR-30 is only the finest amateur receiver made, why have so many professional users bought DR-30's?

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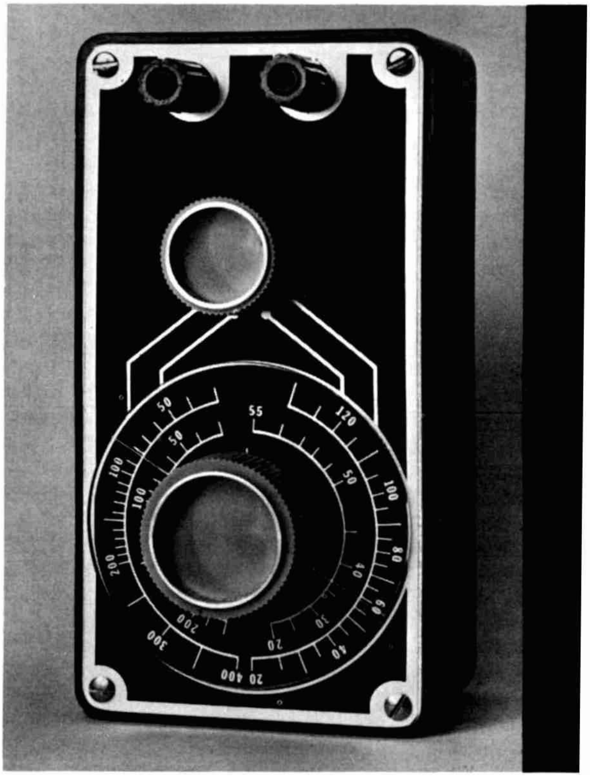
The pre-con, ready to go to work! When used with a grid-dip oscillator, this instrument is very handy in measuring unknown coil values. It is also invaluable as a variable capacitor substitute when designing rf tank circuits.



pre-con

A low cost semi-precision capacitor of simple construction

E. L. Klein W4IBS, 8N Cr n ad
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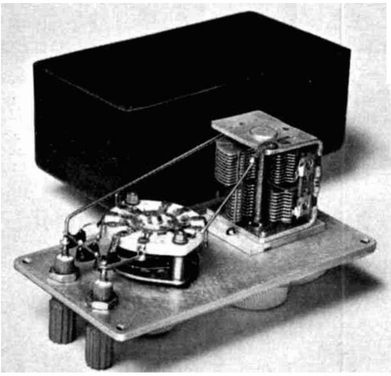
The ham-radio experimenter frequently finds himself in need of a variable capacitor— one which is calibrated to a fairly accurate degree so as to be called a semi-precision capacitance standard. This need was recently experienced when building toroid tank coils for amateur applications.

Very few hams have access to a Q-meter, an instrument which is ideal for proving calibrations and verifying coil construction. But most all shacks have a grid-dipper. Now, by adding the **pre-con** to our collection of most used instruments, we are able to measure unknown coils and capacitors with ease.

'J. Fisk W1DTY, **"A Semi-Precision Capacitor", 73 Magazine, February** 1967.

component parts

The accuracy of a precision capacitor is directly related to the quality of its component parts, as well as the quality of workmanship which goes into its construction. However, if our calibrated capacitor is to be used in conjunction with a grid dipper, we might reason that the capacitor's accuracy need not be significantly greater than that of the grid-dip meter.



Construction of the capacitance standard is simple and straight forward. All components are mounted on the aluminum front panel. Copper bus wire (#16 AWG) is used because of its rigidity for making interconnections.

Therefore, a common midget-radio broadcast-tuning capacitor was selected as the main tuning element. These are available from Lafayette for as little as 39c. All variable mica trimmers should be removed and discarded. The unit pictured here was a junker, but did have the added advantage of a planetary vernier drive built into its 1/4-inch diameter control shaft. This proved to be most convenient. It permitted the scribed pointer to be fastened to the direct-drive portion of the shaft while the knob was fastened to the high-speed end.

The time proven device of inserting a fixed capacitor in series with the tuning capacitor is employed to obtain electrical band spread. This in effect makes it appear that a smaller tuning capacitor was being used, and thence a lower range. Three different capacitors having the values of 510, 180 and 47 pF provided three suitable ranges on the author's pre-con as shown in fig. 1.

The dial pointer is made from 'Is-inch thick lucite. A 2 3/4-inch diameter fly-cutter is used to cut halfway into the lucite from each side. Its cutting bit should be reversed so that a straight-sided cut is made around the periphery of the pointer. A hairline is made by scratching into the underside of the lucite. India ink, carefully ruled into the scratch, will provide a permanent, sharp black hairline.

construction

A small bakelite instrument box was used for the housing of the unit shown in the photographs. These are available from

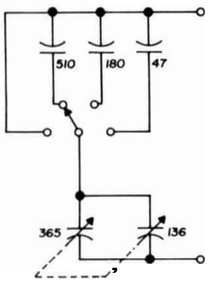
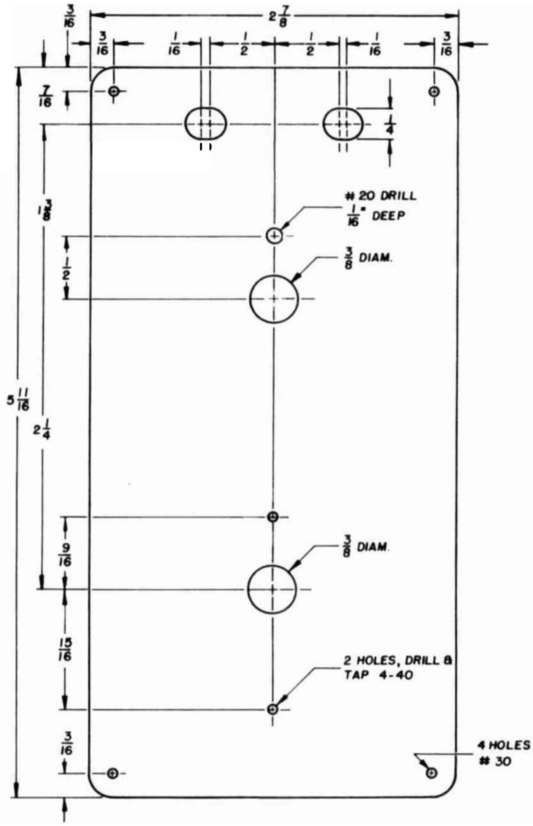


fig. 1. Schematic of the semi-precision capacitor. Dipped silver-mica capacitors are used for the three series capacitors mounted on the switch. A midget broadcast tuning capacitor provides better than 400 pF maximum capacitance.

fig. 2. Panel for the precision capacitor is made from 1/8-inch aluminum. Thinner materiel, such as 1/16 inch, may be used with some sacrifice in rigidity.



several surplus houses. The reader might have shared our surprise in having seen their cost recently rise from 59c to 95c to $1.59. An aluminum mini-box may be the better bet because it provides a ready-made panel, as well as affording shielding around the capacitor. If the plastic box is used, its panel should be cut from 'Is-inch thick aluminum in the interest of rigidity. The vital panel dimensions for the **pre-con** are shown in **fig. 2.**

In the interest of minimizing hand capacitance, the tuning capacitor is insulated from the panel, and the panel is left as a floating shield. If a totally enclosed metal housing is used, it could be secured to the capacitor frame. **Fig.** 3 is a drawing of an insulating mounting bracket which is used to hold the tuning capacitor away from the panel. After wiring up the **pre-con** with stiff #16 AWG copper bus wire, the instrument is ready for calibration.

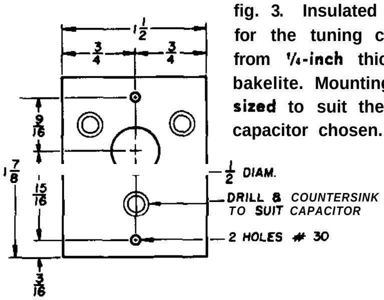


fig. 3. Insulated mounting bracket for the tuning capacitor is made from ¼-inch thick polystyrene or bakelite. Mounting holes should be sized to suit the particular tuning capacitor chosen.

calibration

Each different instrument must be separately calibrated. This is particularly true when you use available tuning capacitors and other odd parts. The full-scale pattern shown in **fig. 4** illustrates a typical panel design. The constructor should check his instrument against another known standard. Placement of the individual dial-calibration points may then be verified.

The first step in calibrating the **pre-con** is to cement a piece of bond paper onto the panel face. Rubber cement works best because it can be readily peeled off later. With the dial pointer and knob installed, the maximum values for each of the four ranges are measured and recorded on a separate piece of paper. Knowing the limits of each range, we can then decide what major scale divisions will be used on the dial.

I had temporary access to a precision General Radio capacitance bridge which made calibration a breeze. Lacking this fine equipment, the amateur constructor should purchase a few 10-, 50-, and 100-pF silver-mica capacitors of the closest tolerance (5% or better) that he can afford. If a capacitor checker such as made by Heathkit is available, these capacitors will serve to double check the accuracy of its dial. If only a grid dipper

fig. 4. Full-scale pattern of the panel name plate. Drawing may be made with pen and ink and mounted on the panel with epoxy cement or it may be photo-engraved into the aluminum itself.

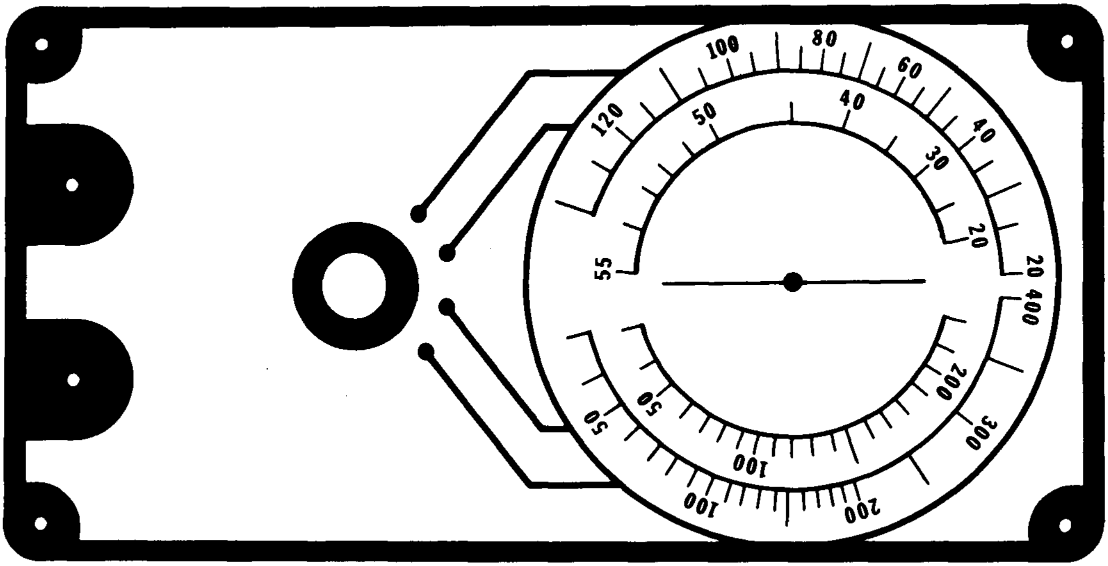


table 1. Material required for the pre-con.

1 Bakelite instrument box. 3" x 6" x 2½" (Lafayette)
1 Panel. 2⅞" x 5-11/18" x ⅛" aluminum
1 Capacitor, miniature BC (Lafayette)
3 Dipped mica capacitors. 510, 180, 47 pF
1 Wafer switch. SP4T (Mallory)
2 Terminal posts
2 Knobs

is on hand, it may be used in conjunction with an arbitrary coil to cross-check between your fixed capacitors and your **pre-con.**

As each major scale calibration point is determined, a pencil mark is made on the temporary bond-paper panel face at the outer edge of the pointer. Different colored pencils are recommended to avoid confusion between the several ranges. Having identified the major dial markings, the intermediate points may be established by interpolation. A careful eye and steady hand will produce a nice-looking job.

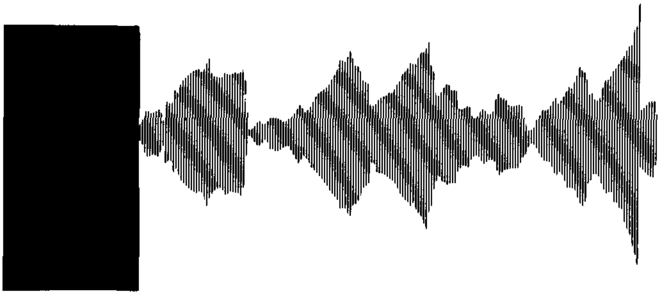
making the panel face

Dial calibration points may be permanently marked on the aluminum panel in a number of ways, India ink will take well on aluminum which has been given a satin finish with Drano or some other caustic. The builder may desire to transcribe his markings in ink onto a fresh piece of bond paper and cement it to the panel. In either case, it is a good idea to spray the finished markings with a clear plastic coat. The panel face illustrated in this article was photoengraved on pre-sensitized and blackened aluminum stock.

using the pre-con

Principle use of this calibrated capacitor will be found in designing rf tank circuits. Of course, its degree of accuracy will in no way match that of the semi-precision capacitor described recently by Jim Fisk. However, when used with a grid dipper, it is most convenient in determining the number of turns required to resonate an experimental coil at a proper frequency and in the correct band. I have found the **pre-con** to be very valuable at W4BRS, and its ease of construction should make it an interesting week-end construction project for many amateurs.

ham radio



a beginner's guide to single sideband

A candid look at what makes single sideband tick

Forest H. Belt, 119 Baker Avenue, South Plainfield, N. J. 07080

There's been plenty of discussion during the past few years about the rag-chewers' migration to a form of voice transmission that is growing in popularity. The "new" mode—called single sideband or just ssb—has for many hams largely supplanted the older amplitude modulation or a-m. Fans of ssb claim superior communication.

Single sideband isn't really new. It's been around since long before World War 2. Telephone and military uses of single sideband antedate its use for amateur radio by several years. Expense was what finally brought it up to date. Once crystals and other parts became available at reasonable cost, and more efficient circuits were designed, single sideband moved into the price range a radio amateur could afford.

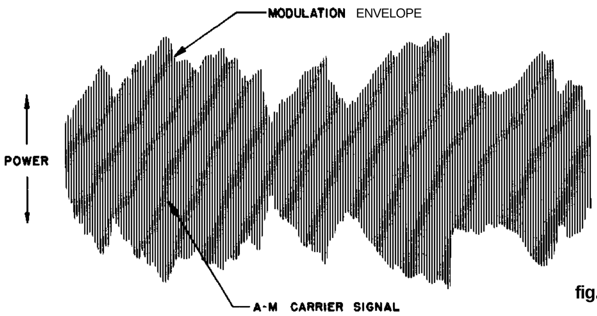
With ssb transmission popping up all over the bands, and even some old timers preferring to switch rather than fight, questions keep coming up: What is single sideband? Why is it better (or worse) than a-m? How is it done? How do you use it? You may have wondered some of these things yourself. The answers lie in the very nature of single sideband. That, then, is a good place to begin.

what is a sideband?

There was a time when sidebands were thought of as a useless byproduct of amplitude modulation. That just isn't true. The sidebands are the most important part of the a-m signal; without them, voice transmission would be impossible. Two of them are produced by ordinary amplitude modulation —one on each side of the carrier.

To understand how they are produced, you must first understand exactly what happens in the process called **modulation.** Reduced to simple concepts, modulation is merely the mixing of two electronic signals of different frequencies. Consider amateur-band communications, for example. One of the signals is at voice frequencies, audio signals generally between 100 and 3000 Hz. The other may be at any radio frequency assigned to ham radio. For this explanation, we'll pick one in the 20-meter band: 14.25 MHz.

fig. 1. Modulation envelope of an a-m signal.



Though the 14.25-MHz radio-frequency (rf) signal is called a **carrier,** it isn't really needed to "carry" anything. It is needed only in the modulator, to help form the sidebands. It will be needed again later for the demodulator in the receiver—but that's getting ahead of the story. In the transmitter modulator, the 14.25-MHz signal is merely a vehicle with which the voice signal can mix. The two are mixed in such a way that two things happen:

1. The power of the carrier signal is varied in exact step with the audio signal. This is called **amplitude** modulation because the voice-varied carrier signal forms a modulation envelope that is a graph of its varying amplitude. **Fig. 1** shows this modulation envelope.

2. The signals heterodyne, as any two electronic signals will do when mixed in a circuit like this. The products of heterodyning are, in addition to the original signals, their sum and their difference. The voice signal is a complex one, as you can see from the modulation envelope in **fig. 1.** It contains many frequencies between 100 and 3000 Hz; for easy calculation, they can be considered as one block of frequencies, called merely the **voice** or **audio frequencies.**

The **sum** of the carrier and voice signals is a block of signals: 14,250,000 Hz **plus** the 100-to-3000-Hz block of voice frequencies; the sum is 14,250,100 to 14,253,000 Hz. This block of sum frequencies is called the **upper sideband,** and is shown in a spectrum graph in **fig.** 2.

fig. 2. The upper-sideband of a signal centered on 14.25 MHz.

fig. 3. The lower sideband of a signal centered on 14.25 MHz.

The **difference** between the two signals is also a block of signals: 14,250,000 Hz **minus** the 100-to-3000-Hz block of voice frequencies; the difference is 14,249,900 to 14,247,000 Hz. This block of difference frequencies is called the **lower sideband,** shown in another spectrum graph in **fig.** 3.

where the power goes

When you look at a spectrum graph of the full voice-modulated a-m signal **(fig. 4),** you begin to see the whole picture of what's happening with amplitude modulation. An astonishing fact is that 66% of the total rf output of the transmitter is in that carrier, and only 17% in each of the sidebands—even at full modulation. This is not by design; it is merely a result of the modulation system.

What makes it astonishing is that the

carrier is not really needed in the output. Once it has done its job of heterodyning in the modulator, it can be eliminated completely as far as transmission is concerned. The voice power is in the sidebands alone, and they are rf signals that propagate just as well as the carrier does.

This brings out why proponents of single-sideband transmission are so critical of a-m. An a-m transmitter with rf output of 750 watts puts 500 of those watts into the carrier. The 250 watts in the two sidebands (125 in each) is the only power that is really of any value to communication. As a matter of fact, you can see from **fig. 4** that either sideband is a mirror of the other, so it is actually a waste to use both sidebands. When the signal reaches a receiver, the information contained in one sideband can, correctly processed, develop just as much recovered audio as the whole double-sideband-plus-carrier signal.

The conclusion? A 125-watt single-sideband transmission can produce at least as much intelligible radio communication as a 750-watt amplitude-modulated transmission. Imagine what you can do with a full kilowatt concentrated in one sideband!

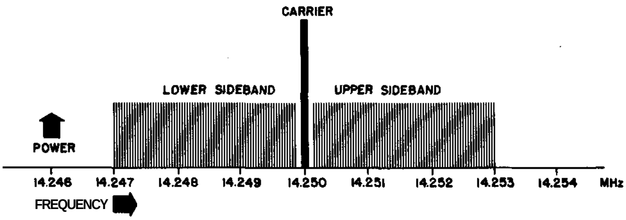
recovering the voice signals

The only reason a carrier is needed at the receiver is for the demodulator. It isn't needed for tuning, because the receiver can be tuned to 14.25 MHz merely by reference to either group of sideband frequencies. In a conventional a-m detector, the carrier and the two sidebands are fed to a nonlinear detector: usually a diode. This causes heterodyning between the carrier and the two groups of sideband frequencies. The difference "group" comprises the original 100-to-3000-Hz voice signals.

Without the carrier, the upper and lower sidebands have nothing to heterodyne with, so the voice intelligence can't be recovered. A single-sideband signal, which is transmitted without any carrier signal, can't be demodulated in an ordinary a-m detector. The answer is to re-insert a carrier of some sort at the receiver.

Since superhets are the only practical receivers for this kind of communications, the carrier-reinsertion problem is simplified. The rf sideband is converted to an i-f sideband. An i-f carrier—which is easier to generate accurately than a 14.25-MHz rf carrier—is added, and the two are fed to a diode detector where normal demodulation takes place.

fig. 4. Spectrum graph of an a-m signal showing the carrier at 14.25 MHz and the upper and lower ride-bands.



generating an ssb signal

Now that you understand the rudiments of single-sideband communication, you probably wonder about some of the more technical details. Knowing them will help you understand ssb even better. For example, there is the question: What's a practical way to develop a single-sideband signal?

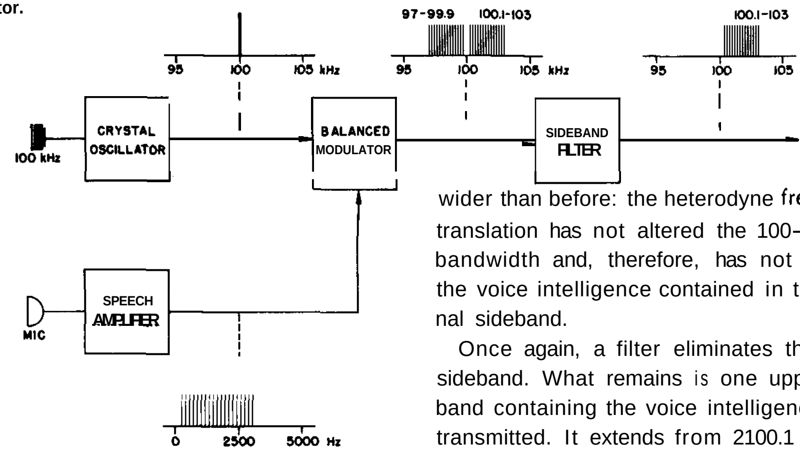
Without worrying about specific details of circuit analysis, take a look at **fig. 5** for an explanation of how a simple single-sideband transmitter generates its signal. A crystal-controlled oscillator, usually at a very low radio frequency, supplies the carrier for modulation. The oscillator frequency may range from 50 to 500 kHz, and occasionally higher. A common one for this purpose is 100 kHz.

The operator's voice is picked up by a microphone, and the voice signals are amplified by the speech amplifier. Voice frequencies below 100 Hz and above 3000 Hz are not needed for intelligible communication, so they are filtered out by circuits in the speech amp. What's left is the 100–3000-Hz block of natural voice frequencies.

Both the carrier signal and the voice signals are fed to a balanced modulator. There are several different kinds of circuits used for this stage, some with diodes and others with tubes or transistors. There are even special switching tubes that make excellent balanced modulators. The balanced modulator mixes the voice and carrier signals in such a way that both sidebands are produced normally, but the carrier is balanced out. **Carrier suppression,** the process is called.

The output of a balanced modulator is technically named a **double-sideband suppressed-carrier** signal. With the 100-kHz oscillator signal and the usual block of voice frequencies (100 to 3000 Hz), the upper sideband extends from 100.1 to 103.0 kHz; the lower sideband extends from 97 to 99.9 kHz. The two sidebands are shown in the spectrum graphs in **fig. 5.** The carrier, which served its purpose in the modulator, has been eliminated in the output of the balanced modulator.

fig. 5. Block diagram of a simple single-sideband generator.



it is in the next operation that the signal first becomes single-sideband. The method is deceptively simple: a filter is used which passes the desired sideband and blocks the other. In some ssb transmitters (variously called **sideband generators** and **ssb exciters),** an electromechanical filter is used. In others, a "lattice" of quartz crystals gives the filter the wideband response curve needed. You see, if the transmitter is to operate on the upper sideband, the filter must pass the frequencies 100.1 to 103 kHz, while blocking 97 to 99.9 kHz. The output of the sideband filter in **fig. 5** shows the lower sideband reduced drastically.

the correct output frequency

Of course, the single-sideband signal at 100.1–103 kHz is not to be transmitted directly. Somehow the single-sideband voice intelligence must be translated to a frequency within the Amateur Radio Service. How about 14.25 MHz, the 20-meter ham frequency we used earlier? This is ordinarily done in two steps, both of which you can see in **fig. 6.**

The 100.1–103-kHz sideband is mixed with a 2000-kHz (2-MHz) oscillator signal. The result of their heterodyning is another pair of sidebands—one equal to the sum of the input sideband and the oscillator signal, and the other equal to their difference. Two things are apparent. The sidebands are much further apart than before, which makes them easier than ever to filter. Also, they are no wider than before: the heterodyne frequency-translation has not altered the 100–3000-Hz bandwidth and, therefore, has not affected the voice intelligence contained in the original sideband.

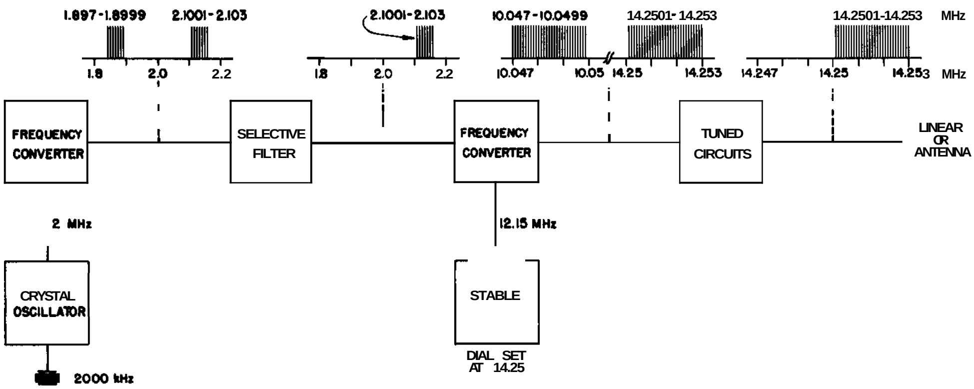
Once again, a filter eliminates the lower sideband. What remains is one upper sideband containing the voice intelligence to be transmitted. It extends from 2100.1 to 2103 kHz. This sideband is ready to be raised to the transmitter output frequency. When that is done, the output signal should be a single sideband extending from 14,250.1 to 14,253 kHz. (That's the upper sideband of 14.25 MHz.)

The second frequency conversion is accomplished in another heterodyne mixer. The rf signal is supplied by a very stable variable-frequency oscillator (VFO). it is variable so the transmitter can operate at other frequencies besides the one in our example.

When the dial of the VFO indicates 14.25 MHz, the stable oscillator furnishes an rf frequency of 12,150 kHz. In the heterodyne mixer, this rf signal beats against the 2100.1–2103-kHz sideband and produces two new sidebands. One of these, the new lower sideband, is the difference heterodyne: 10,047–10,049.9 kHz. The new upper sideband is the sum: 14,250.1–14,253 kHz. The latter is exactly what you need; it is the upper sideband of 14.25 MHz.

Must you have another filter to get rid of the unneeded lower sideband? Not necessary. The two sidebands are so far apart—more than 4 MHz—that one can be easily eliminated now by ordinary tuned circuits. A tank tuned to 14.25 MHz has plenty of bandwidth to include the sideband just above. Therefore, eliminating the lower sideband is no more complicated than merely tuning the output of the second mixer to 14.25 MHz, the "reference" frequency for this single-sideband signal.

fig. 6. Block diagram of a single-sideband transmitter which puts the sideband signal generated in fig. 5 on 14.25 MHz.



For any other transmitter frequencies, the VFO frequency can be reset. At the same time, the tuned tanks following the second mixer must be retuned to the new frequency. As an example, a VFO dial setting of 14.3 MHz causes an oscillator signal of 12.2 MHz. Mixed with the 2100.1–2103-kHz sideband from the second sideband filter, the 12.2-MHz signal heterodynes a new lower sideband at 10,097–10,099.9 kHz and a new upper sideband at 14,300.1–14,303 kHz. The tank circuits following the second mixer, tuned now to 14.3 MHz, eliminate all trace of the 10-MHz lower sideband. The upper one, which is the upper single-sideband signal of 14.3 MHz, is coupled on to the transmitting antenna or to whatever amplifying stages follow.

building up single-sideband power

A complete single-sideband transmitter has another section in addition to those already shown. The final section is the power amplifier. It includes amplification for the sideband signal, plus output coupling to feed the single-sideband signal to the antenna. In both these actions, there is only one difference between the ssb transmitter and any other: the ssb signal must be amplified without the slightest distortion. If the amplifier were nonlinear, the sideband signals would mix and beat against one another, forming an unintelligible mass of frequencies. There is no carrier to maintain a sideband relationship, so nonlinearity in amplification just isn't permissible.

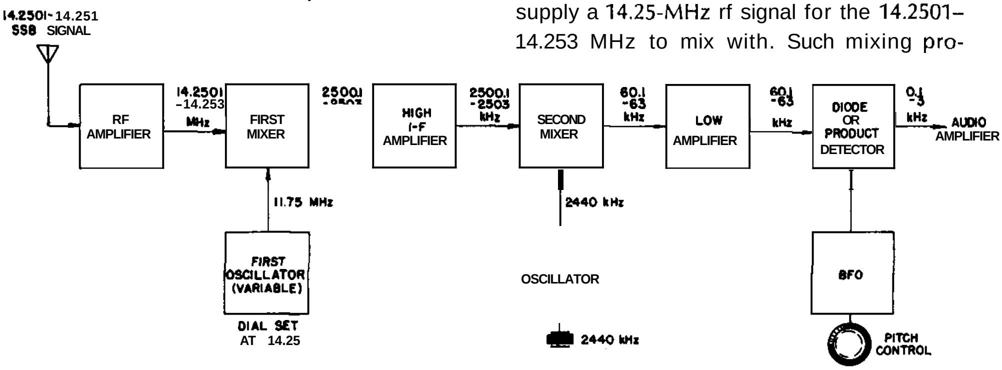
The amplifier, then, must be linear at all cost. The final stage of a single-sideband transmitter is called just that: linear amplifier. It is operated class A for low power, or class AB, or AB_2 for high. The important thing is to make sure the stage (or stages) is adjusted so it generates no harmonics. The sideband must be reproduced intact and unchanged.

Regulation in the voltage supply furnishing dc power to the linear amplifier tubes is critical. Keep in mind that power in the sideband signal follows closely every slight rise and fall in volume of the voice producing it. In fact, when there is no voice, even for the slight instant between two spoken words, there is no power, for there is no sideband signal. This greatly and rapidly fluctuating demand on the dc power supply can be difficult to handle without adequate regulation. Sometimes regulator tubes are used, particularly in the supply that furnishes dc to the linear-amplifier screen grids. Generally, at least with low- or medium-power ssb transmitters, an oversize output filter capacitor is sufficient to handle the sudden load variations.

Measuring the output power of a single-sideband transmitter is more complicated than measuring that of an a-m transmitter. The a-m transmitter produces nearly 70% of its power continuously, in the carrier. Only when a voice modulates the transmitter is the other 30% or so developed, and then only if the voice is loud enough to drive the transmitter modulator to full (100%) modulation. When typical voice modulation is analyzed, its waveforms are found to approach maximum amplitude very seldom. Thus, total output power of a normally modulated a-m transmitter rarely exceeds an average of 10% above its no-modulation power. As a consequence, a-m transmitter power is measured directly, with little concern for modulation.

The power output of a single-sideband

fig. 7. Block diagram of a double-conversion superhet receiver suitable for ssb reception.



transmitter, on the other hand, depends entirely on the voice modulation. As you've already seen, a voice varies the output power considerably, depending on word inflections, voice loudness, even voice timbre. To account for that, power in single-sideband transmitters can best be rated by peak envelope power (PEP).

Peak envelope power is not necessarily a measure of how much power an ssb transmitter is producing at any particular time, but is a measure of its power capability. The wattage figure stated in a PEP rating signifies the output rf power when the transmitter is fully modulated for a period of time sufficient to make the measurement. The truth is that this almost never happens. In fact, the rating is based on a test signal of two sine-wave audio tones; continued stress to that extent can damage most ssb transmitters. Nevertheless, voice peaks do reach the amplitude that generates full power, even if only momentarily. On those peaks the effectiveness of communications rests, and that's why the peak-envelope-power rating is used for ssb transmitters.

at the receiving end

Recovering the voice modulation from a single-sideband signal has only one difference from ordinary a-m. That difference is in the demodulator. As you learned earlier, the carrier that was removed after modulation is suddenly needed again. The sidebands, whatever their frequency, need a carrier to beat against in the nonlinear demodulator circuit. The process is a form of heterodyning.

Suppose the 14.25-MHz upper sideband is to be demodulated directly. It is necessary to supply a 14.25-MHz rf signal for the 14.2501–14.253 MHz to mix with. Such mixing produces a signal equal to their sum, slightly above 28.5 MHz, and another equal to their difference, which is .0001–.003 MHz (100 to 3000 Hz). With the usual rf filtering that follows the diode demodulator, it is easy to see that only the block of voice frequencies will be left.

The accuracy of the reinserted carrier is of considerable importance. Even a few cycles of error can spoil the recovered voice signals by making their frequencies different from those transmitted. At high radio frequencies, like 14 MHz or thereabouts, controlling frequency so closely can be a problem. Fortunately, with superheterodyne receivers, that isn't necessary. The incoming single-sideband signal is heterodyned to the receiver's i-f (or i-f's if the receiver uses double conversion). To re-insert the carrier, all that is needed is an accurate CW signal at the frequency of the receiver's last i-f. With that signal and the i-f sideband fed into a non-linear demodulator, voice-signal recovery occurs without complication. **Fig. 7** shows in block-diagram form just how such a system works in one typical double-superhet receiver.

Accuracy of the re-inserted carrier is still important at the lower frequency of the last i-f of a receiver, but it is easier than at high frequencies. Ham-band receivers use last i-f's from 50 kHz to 500 kHz. A common one is 455 kHz. In **fig.** 7, the low i-f frequency is 60 kHz.

The incoming 14.25-MHz sideband (14.2501–14.253 MHz) is first superheterodyned to the high i-f: 2.5 MHz. That makes the sideband 2.5001–2.503 MHz (2500.1–2503 kHz). Second conversion turns the signal into the low i-f: 60 kHz. That places the sideband at 60.1–63 kHz.

The oscillator that provides the carrier for reinsertion is tuned precisely to 60 kHz. In some receivers designed specifically for single-sideband reception, a phase control is accessible to keep that oscillator exactly on frequency. Any slight alteration can distort the recovered voice signals, and the operator needs some degree of control.

ssb with an ordinary receiver

You don't have to have a special receiver to pick up and listen to single-sideband transmissions. Of course, the specially built ssb set will make the job much simpler, but the fact is that any reasonably stable communications receiver can be used. There is only one requirement: the receiver must have a continuous-wave oscillator (CWO)—the beat-frequency oscillator (BFO) that is used for receiving CW code transmissions (A1). This oscillator can supply the carrier that is needed for reinsertion prior to the demodulator.

You'll need a little practice to become proficient at tuning ssb on your old-faithful ham-band set. First, you must be able to recognize the Donald-duck chatter that is characteristic of ssb transmissions heard on a double-sideband carrier-dependent receiver. When you come across this phenomenon and would like to de-scramble the jargon, take these steps:

1. Make sure the receiver has been on long enough that all its warmup drift has finished.

2. Turn off the avc. Turn the rf gain control to minimum and the audio gain (volume) control wide open. Leave the BFO (CWO) turned off.

3. Advance the rf gain control enough to hear the QRM across the band.

4. Tune in the signal you suspect of being ssb to its strongest spot on the dial. If you have trouble judging strength, "bracket" the signal: tune halfway between its two fade-out points.

5. Turn on the BFO. "Tune" its pitch control for greatest clarity of the voice. This is critical, so turn the knob slowly till the voice sounds normal. You may have to "ride" the BFO pitch control to maintain clarity, unless your receiver is blessed with an unusually stable front end and beat-frequency oscillator.

There are a couple of characteristics that may hamper your tuning-in and listening to ssb signals. One is the fact that many ssb transmitters use VOX transmission—that is, the transmitter is turned on automatically only while the operator is speaking. This may make some ssb signals so intermittent it is hard to get them zeroed-in on the tuning dial. With a little patience, however, you can do it.

Another little problem is the variation in signal strength between transmitters of different powers and distances. With the avc not operating, you may have to juggle the rf gain control a bit between the two sides of an ssb conversation. Don't bother the audio gain control, though. Keep the ssb signal as low as possible in your receiver to avoid overloading; use the rf control for volume adjustment. Again, with practice you will become adept.

That's the ssb story. Make your own decision about whether to go the a-m route or follow the trend to ssb. You can make your decision intelligently now that you know what single-sideband is all about.

ham radio

. . . it may be batter to use several chassis bolted together than one large chassis . . .

for that

professional look,

old man . . .

Save money, impress your friends! Build professional-looking homebrew ham gear

Mike J. Goldstein VE3GFN, 22 Kingswood Road, Toronto 13, Ontario

Are you a builder or an appliance operator? If you're a builder, read on; these words are directed to you. If you're an appliance operator, turn the page quickly, for the secrets to be revealed here could usher you into the art of building such magnificent ham gear that you will end up enjoying the awe reserved by your fellows for the successful constructer.

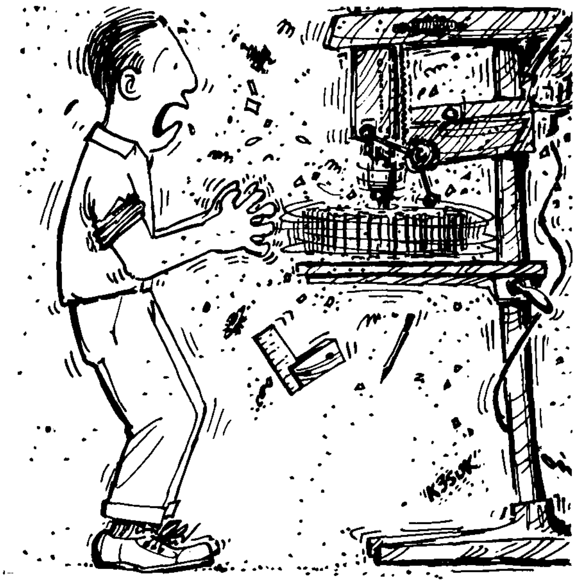
It has been my experience many times in the past few years to have non-hams arrive in my shack, look at a piece of homebrew, and ask where I bought it. Hams who have built equipment themselves have even bought equipment born on my workbench. Since I have been building electronics gear for almost fifteen years (making a living at it at times), I have picked up many ideas and tips that contribute tremendously toward a finished product that is professional looking and a source of pride to me when it is finished. For those of you who are just starting out and don't know where to begin, and for those hardy souls who struggle toward perfection, I submit the following:

The first rule of thumb, and the most important, is to have patience with the job. Think it out beforehand, do it carefully, and don't be satisfied with second best. This is hard when you want to get the job finished, but every rushed job and short cut will hit you right in the eye when you sit back to survey your handiwork. And, you will end up being dissatisfied. Since I have now imparted one-half of my hard-won knowledge, I can proceed to the remainder.

metalwork

The most popular form of ham-gear construction is the "chassis and panel" type. With this type of construction, everything is mounted on a metal chassis or sub-chassis, with the meters, control knobs and switches

Cutting holes in metal can be an awful chora . . .



on the front panel. Small units can be built directly into small utility boxes which are commercially available.

While it is certainly cheaper to scrounge sheet aluminum and bend up your own metalwork, the job will turn out poorly unless you have a proper shop at your disposal. It is far better to buy the proper panel and chassis new and start from there. Since aluminum is very easy to work with, use it as much as possible. Always keep a catalogue of metal parts handy when you're designing a project. Never use painted chassis for building—the grounding problems presented by the paint are horrible, and scraping paint from the inside of a chassis is awkward.

If the job is a large one, it may be better to use several chassis bolted together than one large chassis. This provides shielding between circuits and eliminates flexing which results from a large, flat, metal surface. This can often be a major factor in stabilizing receivers and exciters.

Cutting holes in metal can be an awful chore, or, part of the fun. Let's start with the drill—most of us don't have access to a drill press, but a '14-inch drill and a set of **sharp** bits will usually suffice. Set the work on a firm base before drilling and make sure it won't slip. Before drilling any holes, center-punch them first. This centers your hole in the right place and your drill won't slip and gouge the paint or your finger. When drilling, don't exert too much pressure on metal surfaces that will bend easily.

For larger holes, use a little machine oil on the bit. When working on a painted chassis, cover the entire surface with masking tape or thin paper; lay out the hole centers with a pencil and drill through the protective paper layer. Don't remove the paper until all the holes are drilled. Also, don't put an unprotected painted item on your workbench for drilling—all the metal bits you didn't brush off the bench will pepper your paint with scratches. It takes conscious effort not to make this mistake, and I always have a can of spray paint handy for touch-up jobs.

Try to avoid cutting holes with a drill larger than '14-inch in diameter. If you need a hole larger than this, cut a small hole and enlarge it to the desired size with larger drills. Larger holes should be cut with chassis punches. If you have a set of punches on the bench, life can be beautiful. A complete set is rather expensive, but you can usually find someone who has the size you need. Since they are so handy, a lot of hams get together and buy a set collectively.

If you have access to a machine shop, a circle cutter can be used instead of a punch. However, don't attempt to use a circle cutter with a hand drill— blood makes a poor lubricant. The Adel nibbling tool* is handy for many metalworking jobs. This gadget will cut practically any size hole in material up to 1116-inch aluminum or 18-gauge steel, and is practically a machine shop in itself.

When you use a chassis punch, put a little light machine oil on the bolt threads before each hole is cut. Support the punch in a vise if possible. Cut the clearance hole for the punch bolt just large enough for the threads to clear without scraping; then your carefully-placed centers will not be too far off after the hole is cut. One of the greatest faux pas you can make is to return a chassis punch with a metal ring inside it; the owner will think twice before lending it again.

Chassis punches can even be used to cut 'Is-inch thick aluminum rack panels; this can be done safely, even with the smaller sizes, if you protect the punch. Turn the bolt slowly and use lots of oil — the shearing torque developed on a punch bolt is surprising.

Since painting or retouching metal is an art, I shall desist from comment except to mention that the aerosol-can spray paints do a beautiful job if you follow the directions on the can.

mounting components

Once we have the chassis and panel reasonably prepared for construction, we have to consider the problems of mounting components. It has been my sad experience that predrilled holes never seem to end up where they should, despite the care I have taken to lay them out. Therefore, I always line up my sockets, control holes, etc., with a grease pencil line so I know how things are going to fit. And, I never drill any holes until I am ready to use them. Otherwise (I confess) I end up with at least one hole in the wrong place. This requires disguising, which is difficult; the "ventilating hole" excuse is thin, at best. If you don't pre-drill your chassis, you can change your mind as you build without inviting disaster.

As far as the layout is concerned, everybody says "keep it square"—I second the motion. Draw centering lines on the chassis with a grease pencil (washes off with solvent) so everything fits properly. Even VHF equipment can be laid out neatly, short leads and all. Think the layout out on paper first so you won't forget some essential component. At one club station I know of, the entire rig and control system sit in a six-foot rack — except for the main high-voltage transformer; this sits in a desk drawer beside the operator. Never did figure that one out.

When components are all mounted, nothing should wiggle. Keep the leads short, and mount all components along tag (terminal) strips. You can buy these in all sizes and combinations of terminals, so you should have a supply on hand before construction starts. Build everything as though it were going to be used mobile in a Volkswagen during the Shell 4000 Rally. Use lock washers under all nuts.

Build everything as though it were going to be used mobile in a Volkswagen during the Shell 4000 Rally.

Uniform hardware adds a great deal to the finished appearance of homebrew equipment. Keep a supply of standard nuts, screws, lock washers, and solder lugs on hand. I stock 6/32 screws in 1/4- and '12-inch lengths, with nuts and washers to match, and scrounge the odd sizes as I need them.

Speaking of hardware, never throw any away. When you strip a chassis, throw all the hardware into a box. The gismo you throw

* $4.15 **at Allied Radio, 100 N. Western Ave., Chicago, Illinois 60680. Catalog number 26 B 1153.**

away is exactly the item you'll need next year when Hurricane Whatnot is raging and the world is doomed for the lack of a 3-mil, 1/2-inch white metal bolt with a hex head (thought you'd never need it, hah?). I have saved for years, and haven't been stuck for an odd size yet.

soldering

Ah, yes, soldering. There are solderers, and there are solderers. I have a friend who takes an hour to install one connector on a cable; the man who can tear it off afterwards is a **strong** man. I also know people who finish large kits in a day, but they are not on the air too much. Rig problems, you know.

Except for big jobs that require a lot of heat, the soldering gun is a crude instrument. The ideal soldering iron is between twenty and fifty watts and has a spade bit not more than 1/4-inch in diameter. If it's easy to replace bits on your iron, have a machinist friend turn a couple of bits down for fine work. Note that several of the tips manufactured today have a special coating on them to extend their life and prevent pitting—filing or removing the coating will shorten tip life.

Speaking of hardware, never throw any away.

When you first use a new tip, dip it into acid-free paste and tin it with the solder you intend to use. Don't use a rag to clean the tip; keep a tin containing a wet sponge handy and wipe the iron off on that. If you leave a little solder on the tip when the iron is sitting hot, the tip will not pit nearly so rapidly.

The rules for good soldering—clean surfaces, lots of heat, etc., all apply. Since heat rises, put the iron **under** the terminal being soldered. Use a minimum of solder. Don't depend on solder to hold a wire in position. Each wire should be hooked around the terminal to prevent it from springing off. It is a good idea to wrap the wires securely around a terminal if the wiring is to be permanent. It is my observation that the only permanent things in this world are death and taxes, and the former is being disputed in medical circles. Once you have attempted to remove a wire that has been securely wrapped, you will tend to depend on the solder bond for mechanical strength.

If you have to remove the solder to take off a wire or a component, don't just heat the connection directly. The flux in the solder has long since disappeared, and the hot solder will sit there while components and/or printed board char. Take a length of braid (like the shield on small coax) and dip the end in solder paste. Place the pasted end under the solder connection, put the iron against the braid, and press against the solder. The solder will run up the braid, nothing will overheat, and the wires will be exposed. If all the solder doesn't come off the first time, snip off the solder-saturated braid, redip the end, and repeat the process.

When soldering to rotary-switch terminals, it is sometimes difficult to keep solder from running down the terminals into the switch deck. If you "color" the switch terminal just below the solder point with a soft lead pencil, it isn't possible for the solder to flow down too far.

When you're soldering hookup wire, don't grip the wire just above the stripped section—the insulation will stick to the pliers and peel off when they are removed.

wire

There are many types of hookup wire available, and what you use must be governed only by your finances—buy the best you can afford. The best general hookup wire is stranded, number 20 to 24 tinned copper with Teflon insulation. Teflon insulation will never peel back under heat and is really the best stuff to come along in years. The only drawback is that it is expensive. The other types of insulation should be assumed to be susceptible to heat, and proper precautions should be taken to protect it while soldering. My method is to strip the insulation back about '1s-inch further than I actualy require. After stripping, I twist the strands tightly, and tin to within 'Is-inch of the insulation. The tinning allows such a fast transfer of heat that by the time the heat works up to the insulation, soldering is complete. Incidentally, twist and pre-tin all stranded wire, or you will have ends sticking out all over as soon as you try to bend it.

I have not mentioned solid wire because I find it often snaps off after being bent a few times—especially if you're modifying a circuit. Many of you will disagree with me, but I think solid wire just increases your headaches.

general wiring

Try to cable all the interwiring on a chassis. Use different colored wires so you can easily identify different wires in the cable. Set up a standard color code: green for filaments, red for high-voltage, black for ground, etc., and stick to it. Don't lace your cables—it is difficult enough to modify a cable. If you want to tie them down, tie separate loops at intervals along the cable with waxed lacing cord. You can cut these easily for modifying and then replace. Unlacing a long cable in a tight spot is messy, and while you can sometimes remove a wire, replacing it in a tightly-laced cable is a hairy job indeed!

A better system is to use the small nylon clips which are available in a wide range of sizes. These loop around the cable and are held in place with screws and nuts. They make a beautiful appearance and modifications are a snap.

Two precautions: never run any rf or af signal leads (unless they're shielded) in a cable—the resulting pickup and feedback can run you in short circles for a month. Also, don't solder directly to the chassis—use solder lugs held in place with screws, nuts, and internal-tooth washers.

the finishing touch

When the unit being built has been tested and debugged, the last decals are drying on the front panel, and the TVI tests show negative, the time has not come to place your masterpiece in the console. Not yet.

While it is all open in front of you and your memories are fresh, collect all your notes. Draw a schematic diagram, and make sure all the changes you made along the way are incorporated. Jot down the calibration procedure you discovered was the best. Make a note of the current and voltage readings at pertinent points. Your memory isn't nearly as good as you think, and this information will be necessary if anyone else ever inherits the equipment. Finally, file the information away where you can find it in a hurry.

There it is. You've done a nice job, and it looks as good as it works. It took a little longer, and perhaps cost you a little more, hut, "Say, Sam, where did you buy that classy-looking rig?"

ham radio

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whips and loops

as

apartment antennas

If you're stuck with a poor antenna because you live in an apartment, try one of these recommended by W2EEY

John J. Schultz W2EEY/1, 40 Rossie Street, Mystic, Connecticut 06355

I have lived in a lot of apartments and tried various solutions to the problem of putting up a reasonably useful indoor or balcony-type antenna for 80- through 10-meter coverage. Most apartments have sufficient space to string enough wire around so that a transmitter will load properly, but getting efficient radiation often calls for the experimental approach.

Strictly indoor antennas—types which are erected wholly within an apartment—seem to be the least effective. They often load well, and the SWR may be extremely low when they are carefully cut to resonance. However, much of the radiated energy is absorbed by the surrounding structure and lost. When the same antennas are placed in a wooden frame building they perform fairly well. This is particularly true when they are mounted in the attic away from the wooden structure.

Generally speaking, however, a smaller antenna erected on the outside of the building will perform better. I should point out that antennas which are placed outside a building, but very close to it, don't work too well. An example is a wire from an apartment window or balcony along the side of the building. This antenna can be made to load rather well, but it will couple most of the rf energy into the building. Antennas of this type are generally only useful when local coverage is desired.

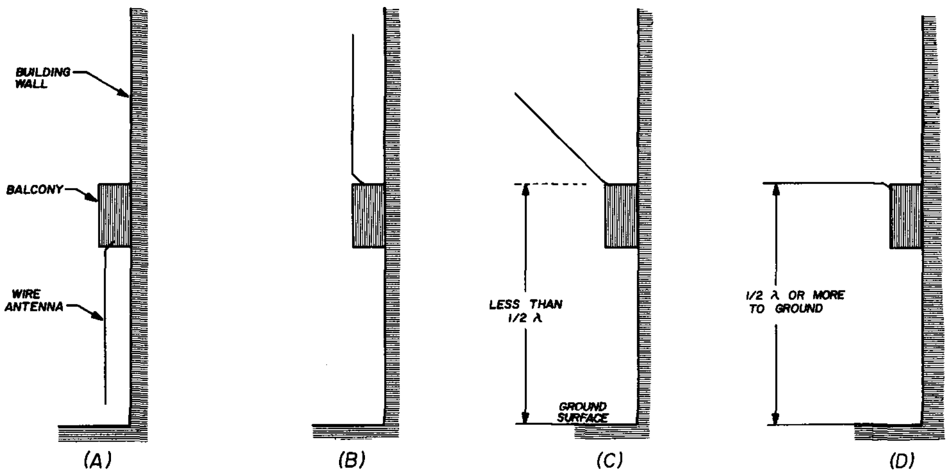
I will describe two antennas here which can be used for this purpose: the loaded-whip and the loop antenna. The loop is particularly interesting because when it is properly fed, it is surprisingly efficient, even when relatively small.

loaded-whip antennas

When you must operate from an apartment, the antenna problem is very similar to the mobile installation. When the antenna is indoors, the structure effectively shields it. Unfortunately, there is usually only limited space available to erect the antenna outdoors. The ground point is an indefinite thing, especially at the lower frequencies where a good electrical ground is desirable.

Since apartment antenna problems are similar to mobile operation, many of the techniques which have been applied to mobile antennas are useful for apartment dwellers.

fig. 1. A wire antenna hung along the wall of a modern apartment building (A) couples most of its energy into the building structure as does a window- or balcony-mounted vertical whip (B). The whip should be installed with as great an angle as possible (C and D) from the surface which acts as its ground plane.



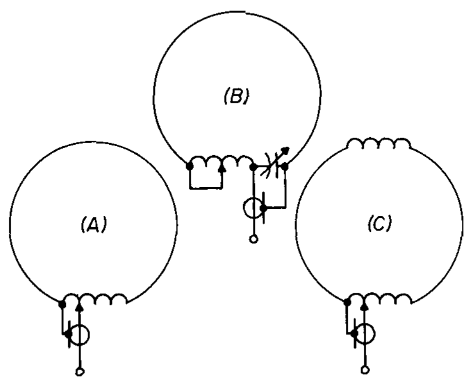
The loaded-whip antenna is not an efficient radiator, particularly on the lower-frequency amateur bands, but it is still one of the best solutions to the mobile antenna problem. Mounted vertically on the automobile, it provides a vertically-polarized radiation pattern of more or less circular shape, depending where it is mounted on the car. The antenna works against the ground plane provided by the automobile and the earth.

When an antenna is used by an apartment dweller, the idea of this **ground plane** is very important. If the antenna is used in an apartment high above the ground, the side wall of the building becomes, in effect, the plane against which the antenna operates. And, for best results, the antenna must be mounted at right angles to this ground plane.

As the antenna is tilted closer to the building surface (toward the vertical), the effective radiated signal decreases rapidly. Although I have made no formal measurements, experience with a ten-foot base-loaded whip in a fourth-floor apartment bears this out. As a balcony-mounted antenna was tilted from a position 90° to the building structure to within about 30" from the structure, signal reports decreased at least 4-6 S-units. The effect is similar to a vertical quarter-wave whip when it is lowered toward the ground.

Because of this effect, recognition of the true ground plane is important. Therefore, a

fig. 2. A simple inductive-loaded loop antenna (A) and variations which may be used for operation from 80 through 10 meters. This type of antenna is inefficient because of the loading.



whip used by the apartment dweller should be placed at right angles to its ground plane, not 90° to the earth. This may change for lower-level apartments because of the effect of the closer ground level. In any case, only a series of actual on-the-air checks can determine the best position for a particular installation.

The effectiveness of the ground connection used in an apartment station is just as important as in a mobile. The mobile ground connection is limited by the fact that the tires effectively isolate the body of the automobile from earth ground. Also, the metal parts of the automobile are not bonded together with low-loss electrical connections. On the lower high-frequency amateur bands it is often necessary to bond the major surfaces of the automobile together with ground straps for best performance.

In an apartment, metal balcony frames and window frames can be used for ground connections but these surfaces must present a relatively large low-loss interconnected mass to be effective. Welded balconies are effective, but screened surfaces which are pressure bonded may present high electrical resistance. Water or heating pipes in an apartment building usually present such a high-loss path to ground that they are useless.

With care in orientation to the building, and a decent ground connection, balcony- or window-mounted whips can be as efficient as a mobile whip. In practice, they are sometimes more effective because of their elevation above ground obstructions. It doesn't seem to matter a great deal whether the whip is base-, center- or top-loaded.

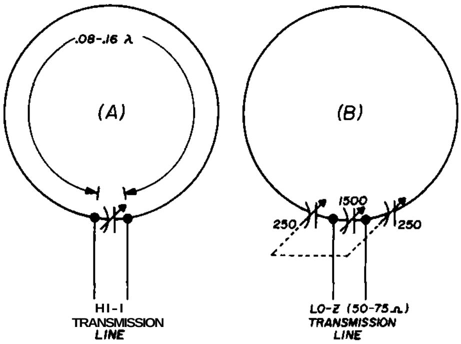
A good ground connection is more readily achieved in an apartment installation than in a mobile. On 10 and 15 meters, where the physical length of a whip can approach a quarter wave, there seems to be a definite advantage to moving the loading coil as far out on the whip as possible. Otherwise, the location of the loading coil on a whip which is less than 10 feet long seems to have little effect on field strength; and, when the coil is mounted at the base of the whip, bandswitching is simplified.

The use of a base-or center-loaded whip antenna in an apartment will generally prove superior to an antenna erected inside the apartment or a wire hung along the side of the building. However, the whip must be properly oriented and a low-loss ground connection or ground surface must be provided.

small-loop antennas

Loop antennas are often used for restricted-space antennas because they may be resonated without the need for a low-loss ground connection. Also, when they are physically small, they are vertically polarized and can be mounted next to any large surface mass. A loading coil is generally used

fig. 3. Because of its inductive property, a loop can be resonated by a capacitor (A) or capacitive divider (B). The scheme in A is useful with high-Impedance feedlines while B is more suitable for 50- or 75-ohm lines. The values shown in B are for a 20-meter loop antenna, 20 feet in circumference, constructed of one-inch tubing.



as shown in fig. 2A to increase the electrical length of the loop to one-half wave, the length of a self-resonant loop.

The radiation resistance presented by loops is very low; their efficiency is dependent upon the losses of the loading coil. Variations of the basic loaded loop are shown in fig. 2B and 2C. In both cases, if the loop is made from one inch or greater diameter tubing, the major losses occur in the loading coil.

The loops I am talking about here are on the order of 0.08 to about 0.16 wavelengths long. These loops are short enough so that the current is in phase and essentially the same amplitude throughout the loop. The radiation is vertically polarized and the directive pattern is similar to a dipole. Maximum radiation is in line with the plane through the sides of the loop.

Inductive loading is not the only method that may be used to make a small loop resonate. Unlike any other short antenna, the loop has a unique, inductive property which is determined by the area of the loop and the type and size of conductor used in its construction. You can take advantage of its inductive quality and use an external capacitance as shown in **fig. 3A** to resonate it. The overall effect is basically the same as inductive loading. Theoretically, you can't obtain more efficient coupling of power by either method. However, in practice, air-variable or mica capacitors have far less dissipative losses than a loading coil; and, the capacitive method of resonating the loop is far more efficient. The efficiency of a 0.16-wavelength loop approaches that of a full-size half-wave dipole antenna. Smaller loops are less efficient, but far better than inductively-loaded, shortened antennas.

You can visualize the loop as a simple parallel-resonant circuit consisting of the inductance of the loop and the resonating capacitor. When a low-impedance transmission line is connected across such a circuit, a high standing wave ratio results. To match the low-impedance line, either an inductive or capacitive tap system can be used. However, the capacitive system is much more practical and easily achieved by splitting up the resonating capacitor as shown in **fig. 3B.** If you are familiar with the **Transmatch*** antenna coupler, you will recognize that this is basically the same matching system.

If we assume that the capacitor losses are fixed, the overall losses will depend upon the conductor losses in the loop. Low conductor losses are achieved by constructing the loop of one-inch or larger diameter tubing. Low-resistance connections between the loop and the resonating capacitors are an absolute necessity. A wide, flat, strip of metal may also be used to form the loop since the objective is to achieve as large a surface area and, therefore, inductance. A square loop with the same maximum radii as a circle will enclose about one-third less area; the inductance and efficiency will decrease in about the same order.

* L. McCoy, "The 50-Ohmer Transmatch," QST, Vol. 45, **July 1961, p.** 30.

Fig. 3B shows typical values of capacitance necessary to resonate a twenty-foot loop (circumference) on 20 meters. Proportionately more capacitance will be required on lower frequency bands and less on 15 meters. The capacitors should be air-variable types for the greatest possible Q. Matching adjustments are basically the same as a Transmatch antenna coupler. First, the center capacitor and then the two outer capacitors are varied (simultaneously) to obtain a low SWR on the transmission line. The adjustment is a back-and-forth procedure from the center to the outer capacitors.

Once the approximate capacitance value has been determined, two of the variable capacitors may be replaced with fixed values —fine tuning is accomplished with one capacitor variable. Bandswitching can be accomplished in the same fashion since the capacitor values for each band have been determined experimentally.

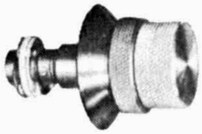
Such conventional qualities as angle of radiation and directive patterns are almost impossible to establish for such an antenna. The directive pattern will usually be entirely dominant in the direction away from the building surface. However, the lowest angle of radiation and best DX will generally be achieved in directions which are an extension of the building plane against which the antenna is mounted. This factor is true for the loop-type antenna as well as the whip.

summary

Apartment antennas present a frustrating but challenging situation as far as antenna erection is concerned. Probably more than any other factor, aesthetic considerations govern the type of antenna which can be used. An outdoor antenna will invariably produce better results than an indoor one.

The most efficient small antenna is the loop; it is also an antenna which can be made small and unobtrusive. If the loop antenna is properly tuned with a capacitive matching circuit and constructed of heavy low-loss "hardware," it will delight the apartment dweller who has given up hope of satisfactory operation on the high-frequency ham bands.

ham radio



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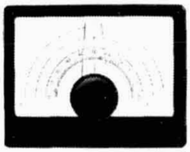

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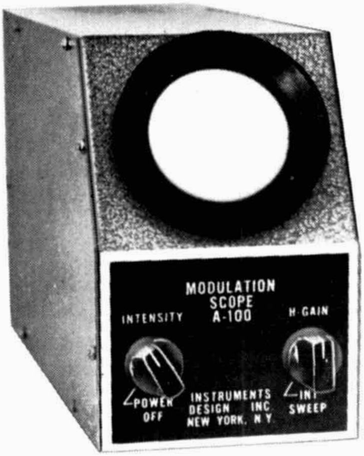

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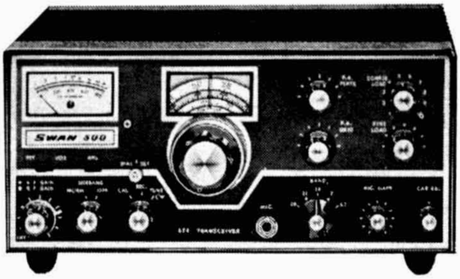

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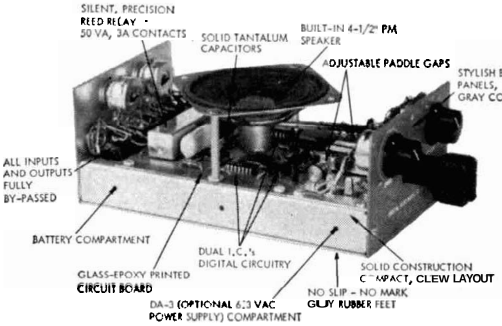

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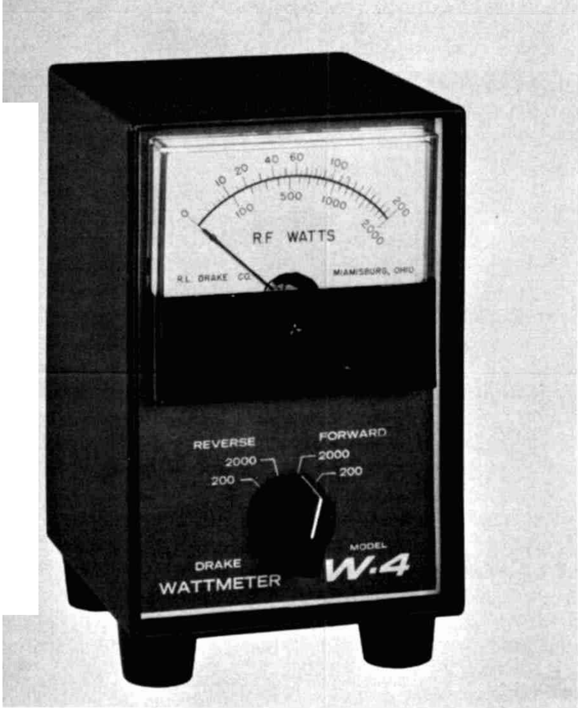


the Drake W-4 directional rf wattmeter

Zero to 2000 watts with good accuracy to 50 MHz

Jim Fisk W1DTY, RFD 1, Box 138, Rindge, New Hampshire 03461.

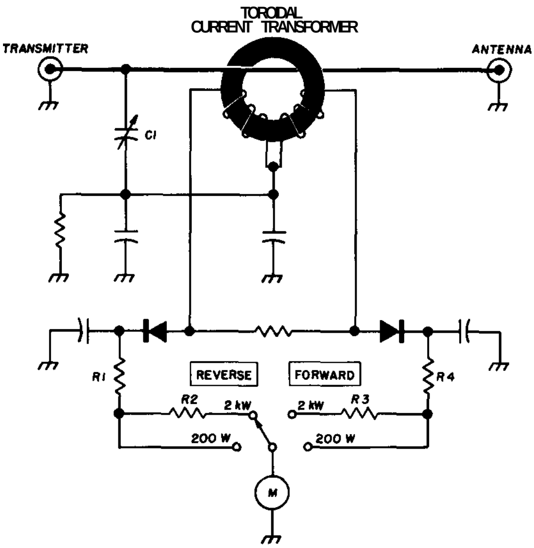
Today, very few amateurs would even consider going on the air without some type of SWR indicator or directional wattmeter in the transmission line. There are many reasons for this. First of all, the popularity of the monimatch-type SWR bridge. Secondly, most ssb transceivers are designed for a relatively narrow range of load impedances. If your antenna isn't resonant, or the matching system isn't adjusted correctly, the transceiver will not load properly.

There are several disadvantages to the low-cost SWR meters that most hams have in their shacks. Most important, their directivity is only about 20 dB, so the best VSWR you can possibly measure with them is 1.7:1. If you use one, and are claiming an SWR of 1:1, you're not kidding anyone but yourself. Another disadvantage of the monimatch-type coupler is its frequency sensitivity. As you raise the frequency of your transmitter, the unit becomes more and more sensitive. Therefore, to use it to measure actual power, it has to be calibrated throughout the frequency range at which you want to use it.

SWR indicators and directional watt-meters both make use of the fact that the forward components of voltage and current along a transmission line are in phase, while the reflected components are out of phase by 180° and cancel. A small voltage derived from the current in the line is added to the voltage across the line. With proper construction and choice of components, this sum represents only the forward power. If the phase of the current sample is shifted by 180°, the forward components cancel and the sum represents only the reflected power.

In the monimatch-type SWR indicator, a small loop is placed next to the transmission line. The current in the line induces a voltage into the loop; the voltage sample is provided by capacitive coupling of the loop to the line. The sensitivity increases with frequency because the current induces more voltage into the loop because its inductance is increasing, and the voltage sample increases because the capacitive reactance is decreasing. Fortunately, the current and voltage ratios remain constant, so the unit will indicate the proper SWR over rather wide frequency ranges. However, since it is frequency sensitive, it is impossible to make power measurements without an involved calibration process.

fig. 1. Circuit diagram of the Drake W-4 directional rf wattmeter.



The Drake* W-4 wattmeter uses a method of obtaining the necessary voltages which is not frequency sensitive in the range of interest, 2 to 30 MHz. In the W-4, the current in the transmission line induces a voltage in a toroidal current transformer **(fig. 1)**. A Faraday shield eliminates any capacitive voltage pickup by the toroid which would affect accuracy. The capacitive voltage pickup is provided by a small variable trimmer (C1) connected across the line. The sum of the two rf voltages is rectified by a diode (one for forward power and one for reverse) and measured by a meter which is calibrated in watts. The variable resistors R1-R4 are used for calibration.

Now, the reason for the W-4's frequency independence. The voltage sample is independent of frequency because it is taken through a capacitive voltage divider. Although capacitive reactance decreases with increasing frequency, the ratio between the two parts of the divider remains constant, so the voltage sample at the tap does not vary with frequency, other things being equal. The current is independent of frequency because the reactance of the toroidal coil goes up at the same rate as the induced voltage.

With this type of coupler, frequency response is limited primarily by the coil. At the

* R. L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

W-4 specifications

frequency response:	2 to 30 MHz, usable at 50 MHz with correction factor
impedance:	50 ohms
VSWR insertion:	1.05:1 maximum
power capability:	2000 watts continuous
accuracy:	± 5% + 2 watts on 200-watt scale, ± 5% + 20 watts on 2000-watt scale from 2-30 MHz
connectors:	SO-239
size and weight:	6" H X 3-11/16" W X 4" D; 2 pounds
price:	$49.50

lower frequencies, response is limited by the point where the reactance of the coil causes a noticeable phase error. The upper-frequency end is limited primarily hy the series self-resonance of the coil, although lead length of resistors and capacitors is also important.

The design of the W-4 is such that its frequency response is down by 2% at 2 ancl 30 MHz. Within this range the accuracy of the instrument is ± 5% of the reading +2 watts on the 200-watt scale, and ± 5% of the reading +20 watts on the 2000-watt scale. Above 30 MHz slight errors began to creep in because of phase-shifts, but correct readings may he obtained on the 50-MHz band by multiplying the indicated power by 1.111.

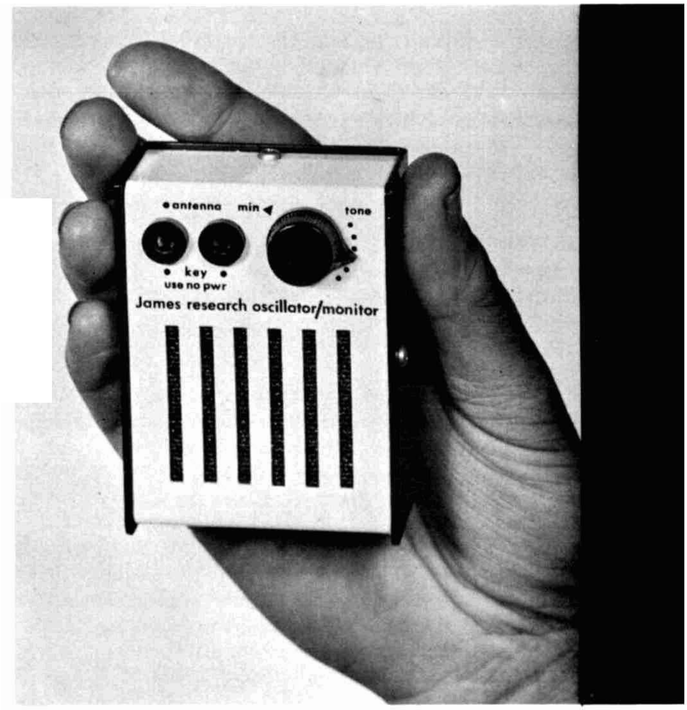
Since it absorbs negligible rf power, it may be left in the line at all times. One nice feature of the W-4 is the removable coupler. It may he removed from the main cabinet and put in a remote location, up to three feet away from the box containing the meter and function switch.

To put the W-4 wattmeter into operation, all you have to do is insert it in the transmission linc and set the function switch to the appropriate range. In the "forward" position, it measures the power that the transmitter is putting into the line. In the "reverse" position, it measures the power that is reflected from the antenna. To determine how much power is actually being radiated (assuming a lossless transmission line), all you have to do is subtract the reverse power from the forward power.

Since the W-4 is primarily a wattmeter, it doesn't measure SWR directly. However, with the forward and reverse power measurements, it is relatively easy to determine. You don't have to make any calculations either— Drake has thoughtfully included a nomograph with the unit that does all the work for you. It's printed on plastic, so it won't get dog-eared with age, and there's a convenient storage place for it in hack of the meter box.

Most amateurs are interested in how well their antenna is matched, and even more interested in how much power they are getting into the transmission linc. The Drake W-4 wattmeter provides both of these functions with good accuracy up to 50 MHz.

ham radio



the James Research oscillator/monitor

Jim Fisk W1DTY, RFD 1, Box 138, Rindge, New Hampshire 03461

At first glance the James Research* oscillator/monitor appears to be just another code-monitoring gadget. Not so. It's a whole lot more. It may be used for checking continuity, testing resistors or even checking components. It is a very sensitive rf detector, providing an audible output with inputs down to 10 milliwatts with an eight-inch pickup antenna. By connecting it directly to the rf source, sensitivity is increased considerably. In addition, of course, it will provide a CW sidetone of your transmitted signal without direct connection to your transmitter or serve

* **James Research Company, 11 Schermerhorn Street, Brooklyn, New York 11201.**

as a code-practice oscillator.

If you're doing any trouble shooting, it gives rapid indications of continuity and open components. It's a lot faster than the old ohmmeter approach, because you don't have to look up at a meter as you go from point to point. You can even check electrolytic and coupling capacitors down to about 0.5 μF. If the capacitor is good, you'll get a short tone burst as it charges; if it's open, no tone; shorted, steady tone. You can even guess capacitor values by the length of the tone burst, but it takes a little bit of practice.

The oscillator/monitor is also great for checking semiconductors, either transistors or diodes. In addition to indicating open and shorted junctions, it will show polarity, anode or cathode, NPN or PNP. If you've ever graded semiconductors with an ohmmeter, you know how long it takes. With the oscillator/monitor it takes a matter of seconds. In fact, I just went through an assortment of 100 diodes and checked them all in less than two minutes. Transistors take slightly longer because you have to check two junctions instead of one, but it is still very fast.

Because of the high-gain dc amplifier which is incorporated into the unit, it is a very sensitive rf detector. With an eight-inch length of number 14 bus wire as a pickup unit, it will provide an audible output when placed near rf sources down to 10 milliwatts or so. If you want to check lower-powered rf sources than this, it may be connected directly. However, don't connect it directly to an rf source greater than 100 milliwatts—you're liable to burn out the input circuit. However, in amateur applications, power levels will usually be above the 10-milliwatt level, even with transistor circuits.

To use the oscillator/monitor as a CW monitor of your transmitted signal, simply place it in close proximity to your transmitter. As the transmitter is keyed, it will emit an audio tone. It may also be used to tune your transmitter for maximum output. In this case, put the oscillator/monitor just close enough to the transmitter so the oscillator just barely triggers. Then, as you tune the transmitter up, and the power increases, the tone will change.

For use as a code-practice oscillator, just connect your bug or key across the input terminals. The audio output can be heard throughout a rather large room. You can also set up two units for code practice between two different locations. Because of the low current involved at the input terminals, resistance of long connecting wires has little effect on their operation, and the oscillator/monitors can be separated by quite long distances.

The circuit of the lames Research oscillator/monitor is really quite straight-forward—a broadband rf detector coupled to a high-gain dc amplifier which triggers an audio tone oscillator. Since the input circuit is untuned, it will respond to rf sources from 100 kHz to 1000 MHz. Also, resistance from zero to 100,000 ohms will trigger the oscillator directly from the internal 1.5-V battery. Trigger current varies from 50 to 100 μA, depending upon the resistance across the input terminals.

The unit is well packaged—16 gauge aluminum. A lasting finish is provided by black and white epoxy resin. For convenience, a magnet is attached to the back. In addition to holding the oscillator/monitor to any steel surface, it grounds the unit for better rf pickup. All things considered, a versatile, well-designed and useful little package.

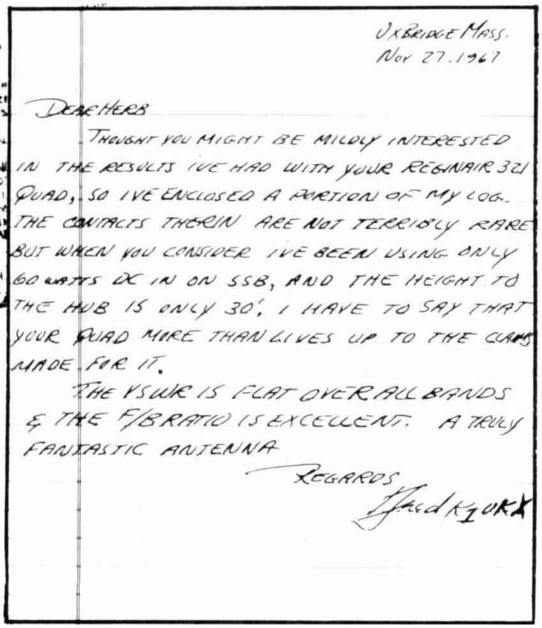
ham radio

oscillator/monitor specifications

frequency range:	100 kHz to 1000 MHz
sensitivity:	10 mW without direct connection to the unit
applications:	rf detector, continuity, component, and semiconductor tester, CW monitor, code practice oscillator
semiconductors:	4 transistors and 2 diodes
power required:	1.5 V AA battery (furnished)
size and weight:	3½" H x 2¾" W x 1¼" D; 8 ounces
price:	$12.95 postpaid

advertisers' index

One of Our Many Testimonials Concerning the Reginair Quad



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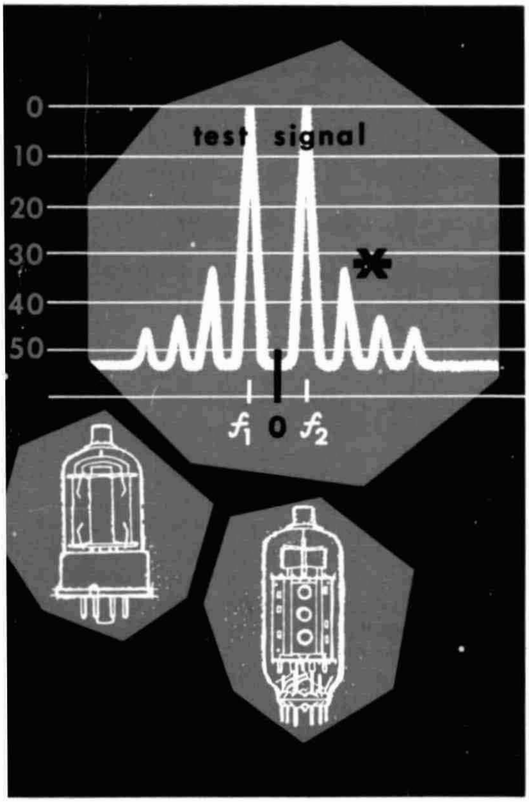

this month

special section

NEW Drake R4B Receiver

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General: With integral VSWR meter and RF wattmeter. Matches 50 ohm resistive transmitter output to coax antenna feedline with VSWR of up to at least 5:1 whether resistive, capacitive or inductive. Covers ham bands 80 thru 10 meters. Has alternate output for tuning up into external dummy load. Meter reads forward power directly and VSWR directly, or can be calibrated to read reflected power directly in watts. Size: 5½" H x 10¾" W x 8" D. Matching network can be switched in or out with front panel switch.

Continuous Duty Output: MN-4, 200 watts; MN-2000, 1000 watts (2000 watts PEP).

Meter reads forward power directly: MN-4, 300 watts full scale with accuracy ± (5% of reading ± 3 watts): MN-2000, 2000 watts full scale with accuracy ± (5% of reading + 20 watts), and 200 watts full scale with accuracy ± (5% of reading + 2 watts).

MN-2000 only: Up to 3 antenna connectors can be selected by front panel switch.

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april 1968

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ham radio

staff

editor
James R. Fisk, W1DTY

roving editor
Forest H. Belt

associate editors
A. Norman Into, Jr., W1CCZ/K2IEO
Nicholas D. Skeer, K1PSR
James A. Harvey WA6IAK

art director
Jean Frey

publisher
T. H. Tenney, Jr., W1NLB

offices

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

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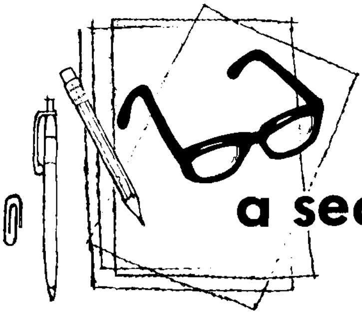

a second look

by jim fisk

Where is amateur radio heading? This is not a new question; I imagine the spark operators on the air before the first big war often wondered the same thing. And down through the years as we make new improvements, we wonder what is ahead.

I'm sure that right after the second war, when the hams had assimilated all of the available technological information that had been generated by the war effort, they thought they had reached the pinnacle of achievement. But, in 1947 came the birth of the transistor. There weren't any big announcements because it was a well-kept secret until the patents had all been secured.

In the early 1950's when the hams started to hear about the transistor and the things it could do, they probably thought that they had really reached the end of the road. But as we can see now, transistors have improved. Many field-effect transistors are now the same price as most of the older bipolar types, and Integrated circuits are taking over many of the jobs that were relegated to vacuum tubes less than 20 years ago.

What will we be using for communications next year? Or ten years from now? Do you have any idea? I don't. Next year, of course, won't be that much different from this. Most hams will still be using their sideband transceivers on the hf bands. On vhf we'll be doing more work with moon-bounce, satellites, meteor scatter and other esoteric modes of communication.

Moonbounce communications, out of the question for most amateurs five years ago, have grown to the point where the EME path can be conquered by any serious experimenter. For example, a few week5 ago, on February 12th, Henry Theobalt, KØIJN, worked VK3ATN in Australia on 144-MHz moonbounce. What makes this QSO interesting is the fact that this was Henry's first moonbounce schedule, and he had **not** received his own echoes from the moon until 15 minutes before schedule time! In addition, although he clidn't have a pre-arranged schedule, he called K6MYC and Mike heard him.

Although we depend primarily on the state of the ionosphere for long-distance communications on our high-frequency bands, and there is no reason to think that it will change in the future, right now we have to wait for the sunspots to come around every eleven years. Fortunately, in the future, we may not be quite so dependent upon the sun. Scientists feel that they can produce an artificial ionosphere. This isn't something we will see next month, or even next year, but in ten years it could very possibly be a reality.

With stationary communications satellites, the demand on high-frequency space should be less. With stationary satellites, the large point-to-point communications and foreign broadcast stations could get better coverage and more reliability by using vhf and uhf. If and when that happens, the high frequencies will probably become the domain of the radio amateur and other less-critical users. The vhf and uhf bands will be the targets of international conferences.

We have to think ahead **now** to reserve and save our vhf and uhf bands for the time when they will be even more valuable than they are now. The plight of business radio today will give you an idea of the number of users who will be demanding vhf and uhf space in the next decade. If amateurs don't use the vhf and uhf bands they have, it will be very easy to lose them. They are presently being used on a shared basis with the government, and the government has priority!

There are some bands in the vhf/uhf range that see very little amateur habitation. Consider 220 to 225 MHz for example. On this band you won't hear a single station on in most parts of the country except during the vhf contests in January and September. If we don't populate these frequencies, there are other users who need them desperately.

I have heard a number of amateurs in metropolitan areas complain about congestion on our two-meter band. FM repeaters are in the vogue now and every metropolitan area has at least one FM repeater in operation. If the congestion is that bad, why not put some FM repeaters up on 220 megahertz? This wouldn't pose any problem that I can see, since you have to make slight conversions to surplus commercial equipment anyway.

Rather than decreasing the frequency, all you have to do is raise it a little bit. I will admit that the percent of frequency change would be greater to put the equipment up on the 220-225 MHz band, but it's still within reason, and within the capability of the commercial units presently being used.

When the amateurs conquered the two-meter band, for some reason they went directly to 432 and didn't stop at 220. I suppose that 220 is too much like two-meters and really didn't seem to offer that much of a challenge. Now is a good time to start thinking about getting some equipment on 220.

How about other modes of communication for the future — new devices and components? Well, it's hard to foresee what we will be using then — who would have predicted the varactor, the laser or the parametric amplifier 15 years ago? Yet today, in some amateur installations, these are rather commonplace. There **are** a few recent advances that will find their way into the ham shack in the next decade, including some of the new diodes, the Gunn oscillator, the plasma amplifier and miniature antennas.

Like I said before, it's hard to make any firm predictions, but based on our previous history and performance, you can bet your bottom dollar that it's going to be a very, very interesting decade.

Jim Fisk, W1DTY
Editor

note to VHFER subscribers

As you were told in the last issue of the **VHFER,** you will now receive **ham radio** magazine for the balance of your subscription. We have tried to catch all of you who have subscribed to both **ham radio** and the **VHFER** so we could combine both subscriptions into a longer one for **ham radio.** If you get two copies of **ham radio,** please send us both mailing labels and we will make the necessary corrections in our records.

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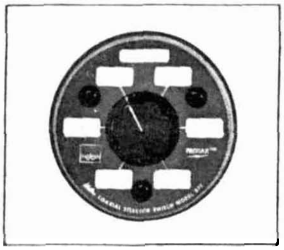

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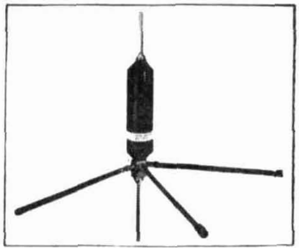

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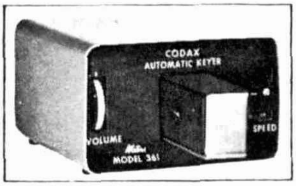

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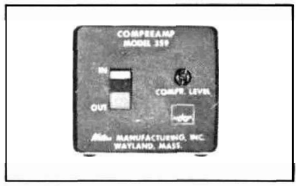

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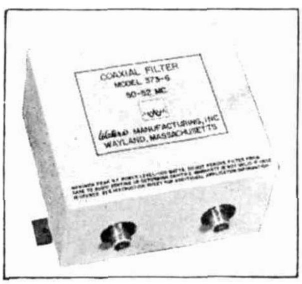

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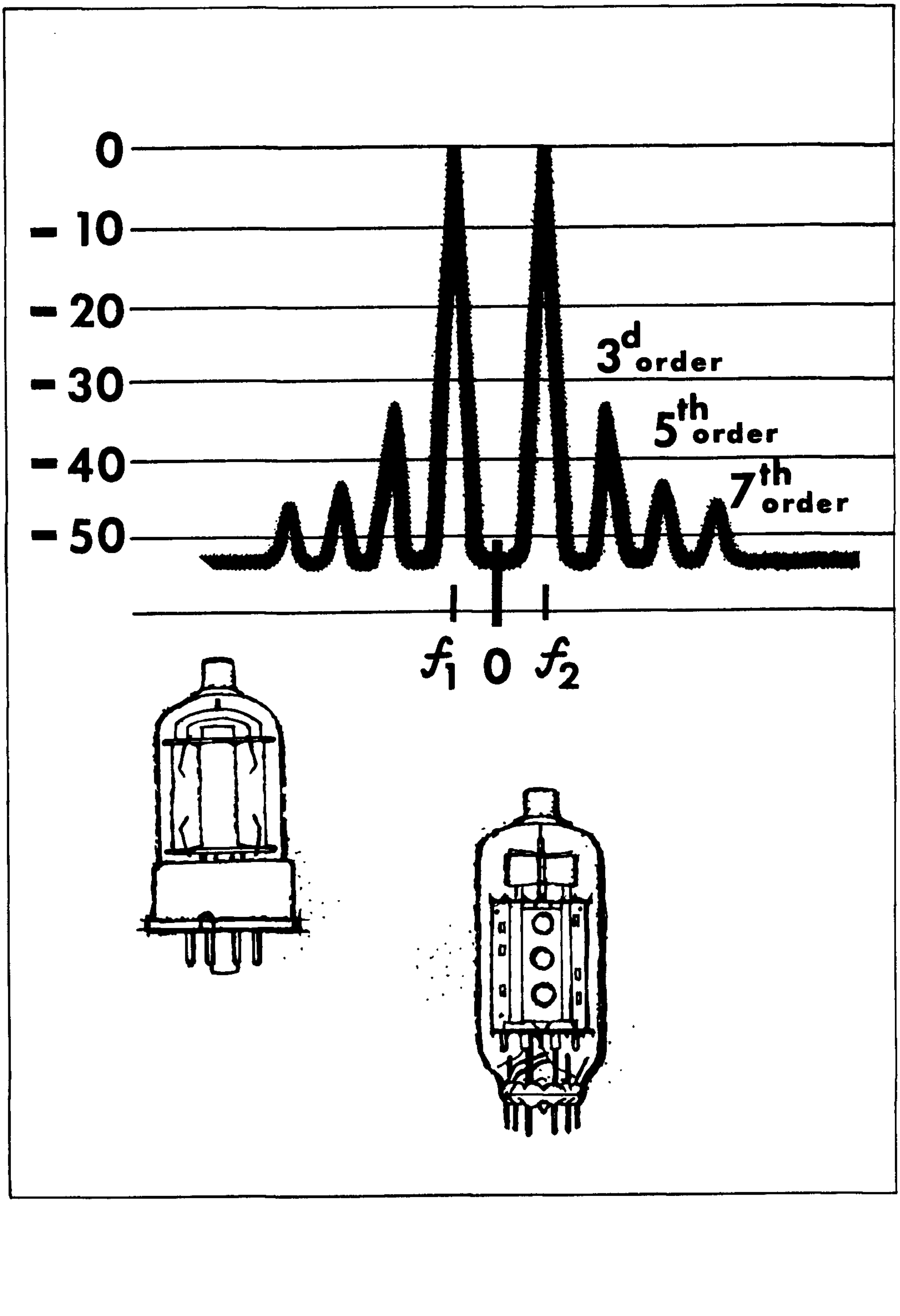
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'full-blast' operation of tv sweep tubes in linear service

Although TV sweep tubes are used extensively as linear power amplifiers in amateur equipment, there has been little published data on their use at rf. Here, W6SAI and W6UOV discuss the intermodulation-distortion characteristics of these tubes.

William I. Orr, W6SAI, 4 Campbell Lane, Menlo Park, California
Robert I. Sutherland, W6UOV, 4040 Fernwood Street, San Mateo, California

During **the** past few years, it has become popular to use TV-type deflection-amplifier tubes as linear amplifiers in amateur sideband gear. These rugged, low-cost pentodes and tetrodes provide amazing PEP capability along with reasonable tube life. However, limited operating information is available on the use of the "sweep tube" in amateur service. Most available data is based upon heuristic (trial-and-error) experiments. It is the purpose of this article to provide some meaningful data covering the use of TV sweep tubes in linear service and to examine the intcrmodulation distortion characteristics of some of them.

Modern horizontal deflection amplifier tubes, while not originally designed for rf work, have several attractive characteristics. There are power-supply economies because they are capable of high peak currents at low plate voltages. Further, wide-spread use in TV sets contributes to low cost and general availability.

Sweep tubes can be divided into two general categories, those designed for black and white television, with plate-dissipation ratings from about 17 to *20* watts, and larger 25- to 35-watt tubes designed for color applications. To satisfy specific design requirements, variations occur in electrical characteristics, internal connections and basing.

the linear amplifier

Modern ssb transmitters generate intelligence at a low level; it is increased to the operating level by means of one or more linear-amplifier stages. The linear amplifier is a device with an output envelope amplitude which is directly proportional to the input envelope amplitude. In other words, the linear amplifier has constant gain, independent of signal amplitude up to the point of overload. The perfect linear amplifier, of course, does not exist; to a greater or lesser extent all linears exhibit amplitude distortion and gain variations with changes in signal level.

A previous article[1] discussed envelope or **intermodulation distortion** (IMD) tests run on high-power linear-amplifier tubes and how the tests were made. Carrying this investigation a step further, we have made intermodulation-distortion measurements on various small TV-type tetrode and pentode tubes, particularly under "full-blast" operation commonly used in amateur-type ssb exciters and driver stages. The results of these tests are discussed and tabulated in this article.

intermodulation distortion

Intermodulation distortion is a particularly noxious form of amplitude distortion found in linear amplifiers driven by a complex signal having more than one frequency. Speech, for example, consists of a multiplicity of tones and is susceptible to IMD in a nonlinear system. Intermodulation distortion of course exists to a degree in all ssb amplifiers. The overall excellence of a linear amplifier may be expressed in terms of the level of intermodulation products as compared to that of the output signal.* The distortion products consist of spurious signals, some of which fall close to, and sometimes within, the operational passband of the amplifier. These spurious signals cannot be removed from the signal by the simple tuned circuits of the equipment. These unwanted emissions are called **odd-order products.**

A clean ssb signal generated by a well-designed and intelligently-operated transmitter occupies little more spectrum than the passband of the intelligence. On the other hand, a poorly-designed or badly-adjusted ssb transmitter with high-level odd-order distortion products (splatter) can smear a swath of frequencies many times wider than that required for transmission of intelligence.

The amount of intermodulation distortion a given signal may possess without creating intolerable interference in an adjacent channel or degrading the transmitted intelligence is subjective and debatable. Some forms of transmission, such as multiplex, require extremely low IMD to faithfully preserve the complex intelligence transmitted.

When the intelligence bandwidth is somewhat less, as is the case with voice, a higher level of IMD may be acceptable. This depends upon the degree of fidelity required, the masking-channel noise level, and the degree of interference tolerated in adjacent channels. Severe intermodulation distortion of a voice signal is characterized by "gravelly" audio and excessive adjacent channel splatter.

As an example, a 1000-watt (PEP output) ssb transmitter with an IMD level of −40 dB in a two-tone test will produce 0.1 watt of power in each of the third-order products. A 1000-watt transmitter, on the other hand, having an IMD level of −20 dB in third-order products will have 10 watts power in the third-order products falling outside the intelligence passband. While 0.1 watt may seem miniscule, 10 watts of unwanted signal may be intolerable, especially when it falls on top of that S-1 DX signal you are listening to!

Studies are presently underway to formalize distortion testing techniques for linear amplifiers. Criteria will be established so linear systems may be evaluated in respect to

* For convenience, the ratio between one of the test signals and one of the IMD products is read as a power ratio expressed in dB below the test signal. Other methods of expression can make the IMD products seem as much as 6 dB better than they actually are.

table 1. 6146/6146B, class-AB, service.

test #	dc plate voltage	dc grid voltage (1)	dc screen voltage	zero signal dc plate current (mA)	maximum signal dc plate current (mA)	maximum signal dc Screen current (mA)	resonant load impedance (ohms)	plate input power (watts)	plate output Power (watts) (2)	third-order IMD products (dB)	approximate plate dissipation (watts)
1	600	-46	200	25	103	9	3570	61	41	−25	16 (CCS)
2	750	−51	200	25	118	7	2825	88	55	−22	28 (ICAS)
3	800	−69	290	30	125	10	3620	100	59	−24	35 (4)
4	800	−77	290	25	180	13	2300	145	91	−19	45 (3)

1. Adjust grid bias for stated zero-signal dc plate current.
2. Does not include tank-circuit losses (about 10%).
3. Applies only to 6146B.
4. Data taken from Collins S-line. Measurements made without rf feedback.
5. Maximum plate and screen currents are listed as single-tone values; voice peaks will run 1/3 to 1/2 this value.

distortion levels acceptable for various transmission circuits. In the meantime, distortion levels are set primarily by the limitations imposed by the state-of-the-art.

The present state-of-the-art in commercial and military ssb equipment calls for third-order IMD products with power levels better than − 40 to − 60 decibels below one tone of a two-tone test signal. The latter degree of linearity may be achieved by proper choice of low-distortion tubes operated in conjunction with rf feedback circuitry. Amateur requirements, as we shall see, are less restrictive by several degrees of magnitude. With an optimum choice of tubes and operating voltages, plus the addition of feedback equipment, designers and users obtain IMD levels (without expensive test equipment) which are acceptable in today's amateur gear.

intermodulation measurements

One industry-wide technique of measuring the intermodulation distortion characteristics of a vacuum tube operating in a linear mode is to run the tube in a two-tone rf test under laboratory conditions where all the parameters are controlled and observed. Various IMD products may be noted on a panoramic analyzer or tunable voltmeter. The circuit parameters and electrode voltages of the tube under test are changed at will to facilitate a search for a condition of low IMD distortion. A typical IMD presentation on the screen of the analyzer is shown in **fig. 1**.

Power tubes up to the 100-kilowatt level or so have been examined in this fashion; a considerable body of literature and test data exist on the linear characteristics of large ceramic and "hard glass" tubes with plate-dissipation ratings of 250 watts and higher. Little data, however, has been formally accumulated on the linearity characteristics of low-power driver tubes. These tubes are usually "soft-glass"** tetrodes and pentodes that are versions of inexpensive audio or television sweep tubes.

In present design practice, high-power amateur ssb gear is physically divided at the 100- to 200-watt PEP level into an exciter and a high-power linear amplifier. The majority of exciters, moreover, use one or more receiving-type tubes of the 6L6 or 6146 family or TV sweep tubes as a linear amplifier.

the 6146 family of tetrodes

The 6146 family of tetrodes is a descendant of the 1936 "grand-daddy" 6L6 beam tetrode. Of convenience to the radio amateur and engineer is the fact that the 6146 family is rated for rf service. Also, application data covering various modes of rf operation are readily available. This popular family of small tetrodes is characterized by short, low-inductance structures which perform well in proper circuitry up to 150 MHz or so.

** Soft glass refers to lead-silicate glass which normally limits the envelope temperature to 240° C or less. Hard glass (Nonex for example) permits envelope temperatures over 300° C.

table 2. ICAS sweep-tube ratings, class-AB, service.

test #	tube type	dc plate voltage	dc grid voltage (1)	dc screen voltage	zero signal dc plate current (mA)	maximum signal dc plate current (mA)	maximum signal dc screen current (mA)	resonant load impedance (ohms)	plate input power (watts)	plate output power (2) (watts)	approximate plate dissipation (watts) signal
1	6GJ5	500	—43	200	30	85	4	3000	42.5	17.5	22
2	6HF5	500	—46	140	40	133	5	1900	66.5	28.8	35
3	6JB6	500	—42	200	30	85	4	3000	42.5	17.5	22
4	6JE6	500	—44	125	40	110	4	2300	55.0	23.4	30
5	6JG6	450	—35	150	30	98	5	2200	44.0	18.9	21
6	6JM6	500	—42	200	30	85	5	3000	42.5	18.3	22

1. Adjust grid bias for stated zero-signal dc plate current.
2. Does not include tank-circuit losses (about 10%).
3. Maximum plate and screen currents listed are single-tone values; voice peaks will run 1/3 to 1/2 this value

A series of intermodulation distortion tests*** run on the 6146/6146B show that this tube exhibits an intermodulation distortion figure in the area of —22 to —25dB for third order products when operated within its published specifications (table 1). When external rf feedback is used (such as employed in the Collins S-line), equipment using the 6146/6146B achieve IMD levels of —30 dB or better (test #3).

Attempts to drive the 6146 beyond its maximum power capability, of course, represents a simple exchange of tube life and substantially higher distortion levels for more output power (test #4).

The peak ICAS plate current capability of the 6146B in class AB, intermittent service is limited to about 125 mA dc. At a reasonable plate potential (750 volts or so), the PEP input level runs about 94 watts. At an efficiency of 65%, plate dissipation is on the order of 34 watts or so, just within the upper rating of the tube.

sweep tubes

Several generations of high-transconductance beam-forming tetrodes and pentodes of 15- to 30-watts plate dissipation have been created for use in TV-deflection circuits. Examples of this family are the 6DQ5, 6HF5, 6JE6, 6GB5 and others. All of these husky low-cost tubes are descendants of granddaddy 6L6, having more cathode emission, greater gain and higher transconductance than their worthy ancestor.

Generally speaking, these sweep tubes are very nearly identical in overall electrical characteristics, varying mainly in physical configuration, pin connections and power capacity. A study of some of these tubes shows that in many cases the various internal parts of the tubes (plates, grids, cathode and supporting structure) are virtually identical. Perhaps the variations in ratings are determined by the applied pulse voltage and sweep parameters of the TV receiver in which the tube is to be used. In all cases, maximum capability of the tube is limited by glass-envelope temperature; 240" Centigrade or less.

Most of the common TV sweep tubes used in radio amateur ssb linear-amplifier service are not rated for this use by the manufacturer. Furthermore, not all of them are rated for audio service, from which ssb ratings may be derived. Class-AB, and class-C operating data for some TV sweep tubes is given in "Sylvania News"[2] (see table 2). According to the author, W. D. Murphy, maximum plate dissipation for some sweep tubes in intermittent ssb linear amplifier service is estimated to be about 1.25 times that value given for TV service.

Several years experience with sweep tubes in linear amplifier service by various manufacturers of amateur gear proved Murphy's data was conservative. In fact, TV sweep tubes may be deliberately subjected to high

***** All tests described in this article were conducted by the authors using the Intermodulation Distortion Analyzer in the Eimac Power Grid Laboratory. All tests were run at 2.0 MHz.**

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peak overloads under the proper circumstances wrthout objectionable loss of life when used in intermittent voice operation. A "full-blast" rating for various tubes was derived (mainly by experience) that permitted PEP input levels of 150 to 200 watts to be achieved, still allowing a good balance between power input, tube cost, and tube life.

This power level is based upon the intermittent nature of amateur transmission, plus the high ratio of peak to average power in the human voice. The factors are hard to pinpoint in actual numbers, but a round, vague figure of 6 dB for the peak to average power ratio has been widely used in designing sweep-tube ssb gear for the amateur market.

With a PEP input of 250 watts under these conditions, the average input power is estimated to run about 62.5 watts over a period of time. If we figure that average efficiency runs about 60% or so, average plate dissipation will be about 26 watts. In most instances, the full-blast rating is further restricted by limiting maximum full-power tuneup periods to 30 seconds in each time period of 2 minutes.

While some rather drastic assumptions are made, practice has shown that full-blast ratings are not unrealistic and good tube life may be achieved (a year or so in normal amateur use). This is provided the operator does not "cook" the tubes during tuneup.

Full-blast operation of sweep tubes in this manner may exceed maximum glass temperature for short periods of time and may eventually lead to seal fractures. When this happens, the tubes go "gassy" after a few hundred hours of operation. Such use, while decidedly uneconomical when applied to a twenty-dollar transmitting tube (especially in commercial gear that stresses reliability), may possibly be considered in a different light when applied to an inexpensive sweep tube that probably will be used only a hundred hours or so during the year. It remains to be seen what happens to the intermodulation clistortion level of the small tube when it is subjected to such overload conditions.

Acting upon the assumption that it is economically feasible to operate a sweep tube at a full-blast 100- to 250-watt PEP input level, we ran a series of tests on various types of tubes to determine their linearity characteristics under duress. In order to hold glass temperature to reasonable values, in all cases cooling air was passed over the tube envelope. Typical rf ratings for intermittent voice linear service were tested. In some cases, operating parameters were duplicated from amateur equipment using the tube. Other working data were derived in the laboratory.

Interestingly enough, certain models of the sweep tube, such as some versions of the 6HF5, were found to have the internal cathode connection at the top of the element structure rather than at the bottom. This

table 3. Full-blast ratings for sweep-tubes, class AB, service.

test #	tube type	dc plate voltage	dc grid voltage (1)	dc screen voltage	zero signal dc plate current (mA)	maximum signal dc plate current (mA)	maximum signal dc screen current (mA)	resonant load impedance (ohms)	plate power input (watts)	plate power output (2) (watts)	third-order IMD products (dB)	approximate plate dissipation (watts)
1	6DQ5	500	—46	150	48	170	17	1800	85	54	—28	27
2	6DQ5	500	—46	150	48	182	13	1625	91	56	—26	29
3	6DQ5	700	—49	150	35	182	11	2210	127	78	—23	41
4	6DQ5	800	—67	180	30	250	13	1710	200	121	—19	70
5	6GB5	600	—41	200	23	192	14	1900	115	80	—18	27
6	6GE5	600	—45	200	30	132	15	2500	79	51	—22	23
7	6GE5	800	—61	250	25	172	18	2750	138	90	—19	39
8	6HF5	800	—45	125	30	197	7	2170	158	100	—21	48
9	6JE6	750	—63	175	27	218	15	1850	163	102	—20	51
10	6KG6	800	—53	160	30	200	5	1000	160	91	—19	60
11	6LQ6	750	—60	175	25	215	9	1850	161	102	—18	49
12	6LQ6	800	—69	200	25	242	13	1850	197	124	—18	60

1. Adjust grid bias for stated zero-signal dc plate current.
2. Does not include tank-circuit losses (about 10%).

extra-long cathode lead provided built-in degeneration that made the tube difficult to drive on 10 meters.

A summary of operating parameters and intermodulation distortion for various sweep tubes is given in **table** 3. It can be seen from this tabulation that the TV sweep-tube family exhibits IMD figures a bit worse than the 6146 group, being relatively constant in the range of – 18 to – 23 dB for third-order products. Correspondingly large values of fifth and higher order intermodulation products were also noted. Reasons for the relatively high level of intermodulation distortion products for this class of tubes are complex, but are probably based upon a combination of low-plate dissipation capability (which restricts zero-signal plate (current) and nonlinear geometry of grid and screen structures.

The degree of intermodulation distortion in the sweep-tube family does not seem to be a direct function of signal level at which the tube is operating as a class-AB, amplifier. Reducing the input level of the sweep tube does not cause a corresponding improvement in intermodulation distortion products; the IMD level holds rather constant as power is reduced.

color sweep tubes

Recent demands for heavy-duty sweep circuits in color television receivers have produced some truly heroic tubes capable of delivering unusually large values of cathode current under conditions of low plate and screen voltage. In particular, the 6KG6, 6KD6 and 6LQ6 seem to be well qualified to deliver large amounts of raw rf under a chosen set of operating conditions. Accordingly, these tubes were examined for IMD characteristics in linear-amplifier service. The results of the tests are tabulated in **table** 3.

The 6LQ6 is rated at 30-watts plate dissipation and the 6KG6 is rated at 34-watts dissipation for TV service. Using the Sylvania rule-of-thumb mentioned earlier, an intermittent rating of 38- and 43-watts dissipation may be expected for intermittent service. Based upon experience with black-and-white sweep tubes, it is reasonable to estimate that the larger sweep tubes may withstand full-blast bursts in excess of twice these values in intermittent amateur voice service. The 6LQ6, moreover, has an additional interesting rating of 200-watts temporary plate dissipation for periods of 40 seconds or less. This allows some latitude in the tuneup process when excessive values of plate dissipation are likely to occur.

The power capability of these compact and inexpensive sweep tubes is impressive. While the plate dissipation under single-tone test conditions is grossly exceeded, if cooling air is circulated around the tube, it is reasonable to assume it is within prudent limits under voice modulation.

The 6DQ5 and 6LQ6 tubes provided the highest level of peak plate current, power output and plate efficiency—about 120 watts PEP output at a PEP input level of approximately 200 watts. Third order intermodulation products ran about – 18 or – 19 dB below one tone of a two-tone test signal. The physically smaller 6GB5 also gave good account of itself, providing a power output of about 80 watts PEP.

It can be seen from the chart that the family of sweep tubes provides a continuum of conforming data, much alike in important aspects. In some instances, maximum-signal plate current is limited by screen dissipation, but in all cases plate dissipation vastly exceeds the published maximum values.

It should be noted that these tests were conducted at a frequency of 2 MHz. It has been reported to us from another source that some 6KG6 tubes failed in linear-amplifier service at 14 MHz at a power input of about 250 watts PEP. Examination of the damaged tubes showed that the internal connecting lead from the cathode base pin to the element structure had melted. It was conjectured that this lead may be made of some type of resistance wire to inhibit "snivets" (vhf parasitics sometimes found in sweep-oscillator service). No tube failures were encountered in the 2-MHz intermodulation tests.

Finally, it should be mentioned that the efficiency of the 6KG6 ran somewhat lower than that predicted by examination of the constant-current characteristics of the tube. Conjecture on this point leads to the thought that the inductance or resistance of the in-table 4. 6550, class-AB, service.

test #	dc plate voltage	dc grid voltage (1)	dc screen voltage	zero signal dc plate current (mA)	maximum signal dc plate current (mA)	maximum signal dc screen current (mA)	resonant load impedance (ohms)	plate input power (watts)	plate output power (2) (watts)	third order IMD products (dB)	approximate plate dissipation (watts)
1	680	—39	340	48	140	20	3010	95	67	—32	21
2	800	—33	290	45	127	15	3920	102	70	—30	25

1. Adjust grid bias for stated zero-signal dc plate current.
2. Does not include tank-circuit losses (about 10%).

ternal cathode lead of the 6KG6 inhibits high-frequency operation of this husky sweep tube. This is a good example of the risk you run when you operate tubes or components in a manner not specified by the manufacturer.

The newly announced 6LQ6 deflection-amplifier tube seems a good candidate for full-blast linear operation, especially in view of the intermittent plate dissipation rating of 200 watts for 40 seconds or less. The 6LQ6, while physically less robust than the 6KG6, provides a good account of itself, as shown in the tabulated data (tests #11 and #12).

It must be noted that these mass-produced tubes have a normal production spread in electrical characteristics. When they are used in parallel, they should be hand-selected to obtain two tubes of approximately the same dynamic characteristic. This can be approximated by comparing the zero-signal resting plate current of a number of tubes and choosing a pair whose currents are closely matched under a given set of operating conditions, or by pairing the tubes in a mutual conductance tube checker.

a "linear" linear-amplifier tube

The only beam tetrode found during these tests capable of relatively high PEP output and low intermodulation distortion in linear-amplifier service was the type 6550, usually employed in hi-fi audio service. The 6550 is capable of a power output of about 67 watts in grid-driven, class-AB_1 service with third-order products – 32 dB down from one tone of a two-tone test signal. Operating data is summarized in **table 4.** In comparison with TV sweep tubes, the 6550 suffers from low transconductance and limited peak plate current. In addition, the interelectrode capacitances are quite high when compared to those of a 6146. Nevertheless, the use of this tube is nearly mandatory if the equipment designer tries to approach an IMD figure better than – 30 dB without the use of feedback. With feedback, an IMD figure better than – 40 dB should be realizable.

summary

While the use of TV type sweep tubes as linear amplifiers in amateur ssb exciters and transceivers may be justified on an economic basis, putting a high-power linear amplifier (even one having negligible distortion levels) after such gear is bound to result in an increase in the level of the various distortion products. In many cases, the distortion products add in voltage amplitude; in all cases, the sum of all components of each frequency product in the output represents the overall distortion level.

Signal distortion, at least to the listener, is a highly subjective thing. To date, the use of sweep tubes in amateur equipment, regardless of the relatively high distortion level, may perhaps be justified when equated against power output and tube life on a dollar-and-cents basis. In any event, until a low-power driver tube in the five-dollar price range comes along to deliver 100-watts PEP at a magnitude of improvement of intermodulation distortion, we'll just have to make-do with the tubes at hand!

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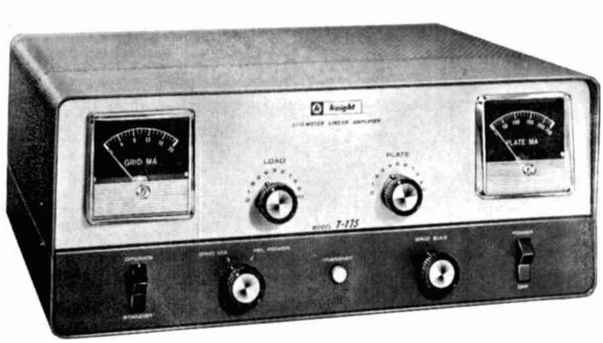
ham radio



One kit-form linear amplifier for 6- and 10-meter ssb uses triode-connected pentodes in grounded-grid class-8.

linear power amplifiers

What they are and why they're necessary for ssb operation

Forest H. Belt, 119 Baker Avenue, So. Plainfield, New Jersey 07080

Single-sideband signals are the most complex in voice communications. The frequencies in them bear a whole range of relationships to one another, and the last thing they can bear is to have those relationships upset. Once the frequencies in a single-sideband signal get messed up, there is no hope of unscrambling them properly at the receiver.

Once a clean ssb signal has been formed by a transmitter (or by an **exciter,** as a low-power ssb transmitter is called), amplifying its power isn't as simple as with an ordinary a-m signal. There is no dominating carrier in the ssb signal to maintain relations among the sideband frequencies. A class-C rf amplifier, the kind used for heavy power amplification of a-m signals, is very nonlinear. The tuned output tank circuits do a good job of restoring the balance, but they can do it mainly because of the strong carrier against which the sidebands (on both sides) can beat to keep their "positions." A single-sideband signal, without a carrier, must be power-amplified in a stage that has virtually no nonlinearity. The sideband frequencies must all keep their positions, with no extraneous frequencies developed from beats among the sidebands or added by the stage.

The power-amplifying stage that accomplishes this feat is called a **linear amplifier.** It may contain more than one tube, to develop the power required by the demands of communications. Or, it may use a single high-power tube, driven to full output by the signal from the ssb exciter. (When a ham has decided he wants the distance-shattering "push" of really high power, his ordinary transmitter becomes the exciter. His new linear amplifier is a completely separate unit with the heavy-duty high-voltage power supplies that are necessary.)

efficiency vs. linear operation

As you probably know, truly linear operation of an amplifier is class-A. The bias for the tube is chosen to place the operation on the linear portion of the grid-voltage-plate-current (E_g-I_p) characteristic curve. Fig. 1 is the graph of E_g-I_p in one tube, and an arrow points out the spot on the curve where bias sets class-A operation. As long as the drive voltage doesn't vary the bias beyond the limits of the straight (linear) part of the operating characteristic, the class-A amplifier introduces no distortion.

The only trouble with class A is its inefficiency. At best, it can never exceed 50%; normally, 35% is pretty good. Even with the advantages of ssb, a 1000-watt dc input would get you little more than 500 or 600 watts of peak envelope power (PEP). That's not enough. What you need is real power amplification, not merely high-power voltage amplification.

Class B is one answer, even though there is quite a bit of distortion. Bias is set near the cutoff point of the tube (fig. 1). You can develop a lot of power amplification with a tube operating at this point on its curve, because it draws plate current only half the time. With that same 1000 watts of dc input power, you can develop up to 2000 watts or so of output PEP. That's good, but what about the distortion? There are ways to reduce it, and so there are several good linear amplifiers using class-B power amplification. Before we study them, though, there's a compromise mode of operation to be considered.

Class-AB operation has one important advantage over class B: less distortion. The improvement is obvious, since distortion can wreck an ssb signal. As you might expect, the operating point for class AB is somewhere between A and B (fig. 1). Class-AB operation is right at the knee of the tube's E_g-I_p characteristic. Almost half of the input waveform runs the tube over the linear portion of its curve. This, and the fact that drive can be quite high in amplitude, make AB especially attractive.

There are two modes of class-AB operation, too. One of them—class AB_1—takes full advantage of linear operation over half of each input cycle. The input (drive) signal is kept low enough that it never swings the grid voltage into the positive region, so grid current (which upsets tube operation) never

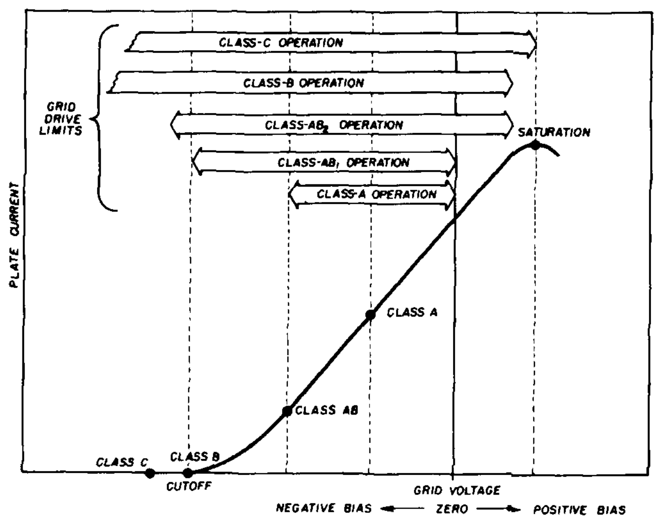


fig. 1. Characteristic curve of one tube type, showing operating conditions of various classes of power amplifiers.

gets a chance to flow. The result is low distortion, while still enjoying the efficiency of operating at class AB. When the drive-signal voltage is raised to a level that causes grid current during a short portion of each positive input cycle, operation is called class AB_2; distortion is much higher.

For efficiency, then, along with linear (distortionless) operation, a good compromise is to run a tube in class AB_1. This mode is popular among commercial linear amplifiers. There are even special power tubes designed for this type of operation. Nevertheless, the attraction of even greater efficiency (more output power from less dc input power) leads some designers back to the class-B power amplifier, using special circuit designs (and special tubes) to overcome the distortion drawback.

designs for linearity

When the need for efficiency overrides the advantages of operating a tube at low distortion, special steps must be taken to combat nonlinearity. Under all circumstances, the ssb signal must not be made to introduce spurious frequencies during power amplification.

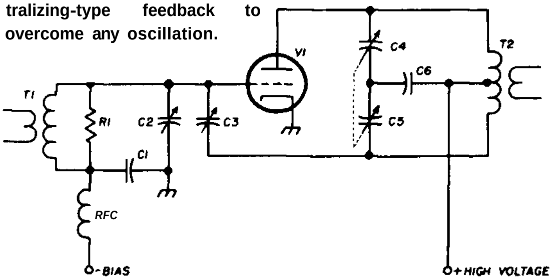
One common way to smooth out distortion is through careful choice of the L and C values in the output tank circuit. The Q of the tank should be 12 or more, because the flywheel effect in the tank returns the undistorted sine shape to each cycle of the signal. If the Q is too high, bandwidth is sacrificed, and distortion occurs from that source. Between 12 and 15 is best for tanks in most class-B or class-AB_2 linear amplifiers.

The most popular tubes for high-efficiency class-B rf amplification are triodes. Operating triode tubes at high power levels introduces new problems. There is a tendency to self-oscillation caused by some of the output signal getting back to the grid, in phase. Fig. 2 shows a way this is combatted in some linear amps. Capacitor C3 from the bottom of the output tank feeds enough output signal back to the grid, out of phase, to prevent any oscillation.

One problem with tubes driven in class B or AB_2 arises during high excitation, when a small part of each positive input cycle momentarily drives the grid positive. That quick burst of grid current can really mess up the output signal for that short time. The reason is traceable to the sudden change of input impedance when the grid draws current. One way around this problem is to keep the input impedance low at all times. Resistor R1 across the input coil (fig. 2) loads it down and overcomes this difficulty to an acceptable degree.

Self-oscillation in triodes is caused by Miller **effect,** a result of inter-electrode capacitance between the plate and grid. The effect occurs most readily when grid and plate are tuned to the same frequency. One cure is to operate the triode with its grid grounded (fig. 3). The cathode is driven, and the grid makes a handy shield between input and output circuits. Furthermore, any plate–grid capacitance now feeds the signal back in wrong phase to regenerate. Most present-day triode linear amplifiers use the grounded-grid circuit configuration.

fig. 2. Grounded-cathode triode power amplifier with neutralizing-type feedback to overcome any oscillation.



Tetrode rf power tubes eliminate Miller effect. The screen grid isolates the input grid from the output plate, at the same time greatly improving power gain in the tube. Introducing this additional element creates the problem of another power supply. The screen grid in most rf power amplifiers takes several hundred volts of positive dc. Because of the dynamic effect the screen has on plate current through the tube, it is important that the screen supply be well regulated. Virtually no fluctuations can be allowed under load, or else that old bugaboo of distortion will rise up to plague the single-sideband signal.

There are three ways to accomplish good regulation: use an electronic regulator of some kind, "tune" the screen-supply filter choke, or put such a low-value bleeder across the voltage supply that the screen current is a light load by comparison. Each is found in modern tetrode-tube linears. As one example, the screen supply in the Collins 30S-1 linear amp is nearly 500 volts, bled by a 5000-ohm resistor. That makes the bleeder draw 100 mA, while the screen (of a 4CX-1000A tetrode) averages 15 or 20 mA.

Some caution has to be used with shunt-type regulator tubes: the range of screen-current variations is wide in ssb linears: the range of firing voltages for the regulators may be too narrow to cover the changes adequately. Much care goes into the design of the screen power supply in a single-sideband linear amplifier.

One effective distortion-limiting device is used in both triode and tetrode linear amplifiers. That is **feedback.** Negative, or degenerative, feedback is probably the best single deterrent to distortion used in any of the linear power amplifiers. You can see the feedback configuration of one tetrode stage in **fig. 4.**

This stage is an exceptionally stable one anyway. The screen is operated at dc ground (with negative of the screen supply going to cathode); thus it makes an extremely effective shield against any possibility of Miller-effect oscillation. The grid is at rf ground—an even further assurance of stable operation at high frequencies and high power. The cathode is the input circuit, driven through a broadband pi-network to keep the input impedance low for stability. An output-tuning system with Q of about 14 assures a clean sine-wave output. The whole stage is operated class AB_1, eliminating distortion problems that might arise from class-B bias levels or class-AB, drive.

Yet, on top of all that, a small amount of negative feedback is included. The 220-pF capacitor (C4) at the grid naturally doesn't bypass rf completely. Instead, C3 is part of an rf voltage divider with C4, coupling some out-of-phase rf signal to the grid. The net effect, even though the grid is not the driven element, is to oppose any distortion that has been introduced by amplification in the tube. Coil L3 and resistor R1 are a suppressor to keep parasitic oscillation from forming in the high-energy feedback circuit from the plate.

The stage in **fig. 4** is a simplified version of that used in the Collins 30S-1. It was chosen as an example because it uses so many of the designs already mentioned as useful in keeping a linear amplifier true to its name. Other commercial linears use many of the same techniques. Triodes, being less expensive, are used more than tetrodes; but both types can furnish a powerful boost to a single-sideband signal without too much distortion, if the proper correctives are designed into the circuit.

tuning up a linear

With modern designs, the guesswork is already taken out of tuning up a linear. Still, most hams like to know what it is they're really doing when they're twisting those knobs.

Only in elaborate linear amps is the procedure much different than in an ordinary power amplifier. So, let's start simple. Tune up the exciter first, and set it for cw operation; you need an rf signal for tuning up the linear. Set the exciter to drive the linear as lightly as possible at first, and set the linear's output loading knob (if it has one) for minimum loading.

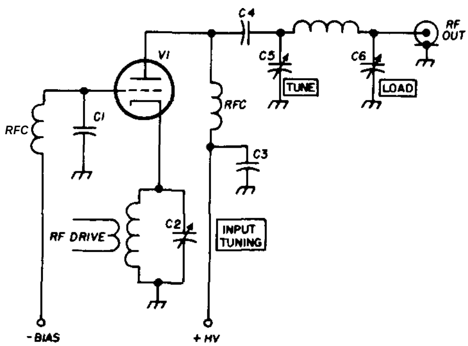


fig. 3. The grounded-grid configuration is one way to overcome Miller-effect oscillation in triode power amps.

Note the plate-current meter reading of the linear, and slowly increase drive from the exciter until the plate current reaches about double its idling value.

Adjust the plate-tuning knob for minimum plate current. This adjustment is easier if there is an rf-power output meter to watch, in which case the adjustment is for maximum output.

Next, tune the loading knob for maximum plate current or maximum rf output. Retune the plate-tuning knob, again for a plate-current dip or for maximum output.

Now that the linear is tuned and loaded, increase the grid drive by raising the output of the exciter. If the linear has a grid-drive meter, use it to determine proper excitation. If not, use the plate-current meter to judge drive; increase it only until plate current reaches the operating level recommended for that tube or by the manufacturer of the linear amplifier.

Redip the plate-tuning knob. Recheck the output loading. You may have to find a happy medium between drive and loading, since both can raise the plate current. However, there's a rule of thumb: increase the drive until increasing it doesn't raise the plate current as fast, then back it off at least 10%. Then, with the plate tuning dipped, if the current is still higher than the ratings suggest, reduce the loading.

The very simplest linear amplifiers don't have a loading adjustment, or a plate or grid meter. One model docs have an rf output meter which you can use to judge the effect of its one adjustment: plate tuning. Just tune for maximum power output.

One elaborate model has grid-circuit tuning, even though the input circuits of most linear amps are rather broadbanded. This control is adjusted for maximum grid-current reading.

A few linear amps require neutralizing, which is done the same as with any other rf power amp. Once set, neutralization can generally be left alone. The object is essentially to stop interaction between the input and output circuits. The screen voltage can be removed, which makes the indication more sensitive; in a triode, plate voltage should be greatly reduced.

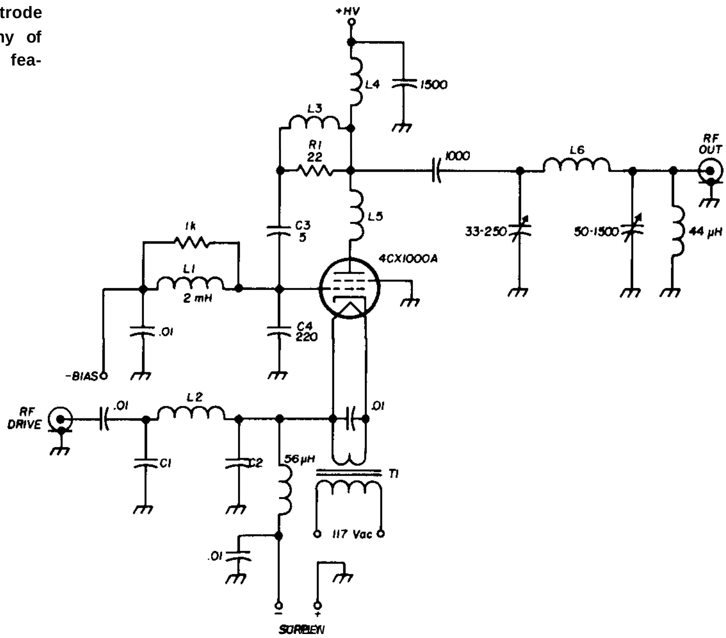


fig. 4. This stable tetrode linear amplifier has many of the distortion-reducing features described in text.

The object of one adjustment method is to keep plate-tuning adjustments from affecting grid-current readings. Monitor the grid current while tuning the plate-tank capacitor. Keep tightening the neutralization trimmer until the effects of plate tuning can no longer be noticed on the grid-current meter. Then refine the adjustment with full plate and screen voltage.

Another neutralization method involves checking for excitation power being fed through the output tube with all plate and screen voltage removed. A sensitive rf output meter, coupled to the output of the linear, is the monitoring device. With the transmitter all tuned up, remove plate and screen voltage, run the drive as high as it will go, and adjust the neutralization trimmer for minimum output. This method is useful if the linear is a simpler one without a grid-current meter.

protecting the equipment

Linear amplifiers need protection from two things: overloads in the circuit, and overheating. The latter is easiest: a blower in most linear amps keeps air circulating over all the components that develop much heat. If the equipment is operated within design limits, there is little fear of deterioration from too much heat.

Circuit .overloads are something else. An overload or a breakdown in one spot may damage an expensive part in another. Protection is necessary. As an example, if plate voltage is lost for some reason, there should be some automatic provision to remove screen voltage; otherwise, the power tube will quickly be damaged. To prevent this, a plate-and-screen overload relay may be provided, especially in the higher-powered models.

Really, though, protection begins when an elaborate and powerful linear amplifier is first turned on. Fuses are included in each important primary circuit. Then, so no plate voltage can be applied to the power tube until its cathode is good and hot, a time-delay relay turns on at the same time as the filaments and blower. When the 3 to 5 minutes have passed, a set of contacts close, and the main dc power supplies are ready to be turned on.

The voltage must be applied to the power tube in proper sequence. In one high-power model, the grid bias is applied to the linear power tube as soon as the main power goes on. After the time delay, the "plate-on" switch can be pushed. Even then, only reduced voltage is applied to the screen and plate. The tube doesn't get full dc power until another time-delay relay has gone through its cycle.

A thermal sensor may shut down the plate and screen supplies if the power tube gets too hot. If the blower quits, the lack of cooling air triggers the sensor. If the tube exceeds its dissipating rating for any reason, the sensor removes power. Indeed, the expensive power tube in a well-designed linear is thoroughly protected.

protecting the user

Safety-consciousness is necessary around equipment with voltages as high as those used in linear amplifiers. Interlock switches on all the covers disable the power-supply primaries whenever the equipment is opened up for inspection or servicing. Never operate the unit with interlocks cheated.

The filter capacitors across the high-voltage supply can store a charge that will kill, even with the supply turned off. In a properly designed supply, the charge is drained off by bleeder resistors. Nevertheless, DO NOT trust them. Indeed, some linear amplifiers have interlock switches that discharge the capacitors just in case a bleeder hasn't done its job. One model has double interlock switching to do this job. Don't trust those, either. Always clip a set of jumper leads to ground *first;* then attach the other ends to each power supply point. Leave them in place until you're through inside the chassis.

Watch out for rf, too. Even though the peak envelope power of a single-sideband signal isn't as strong as a steady cw signal of the same power, that word **peak** means just that. There's a powerful lot of rf at the output and in the power stage when the linear amplifier is operating. Keep away from that rf power. Even when it isn't fatal, it can make a nasty, hard-to-heal burn.

Safety is paramount. Study the operating manual of the linear you're going to work with. No newcomer (nor anyone else) has any business on the receiving end of a nasty jolt or burn.

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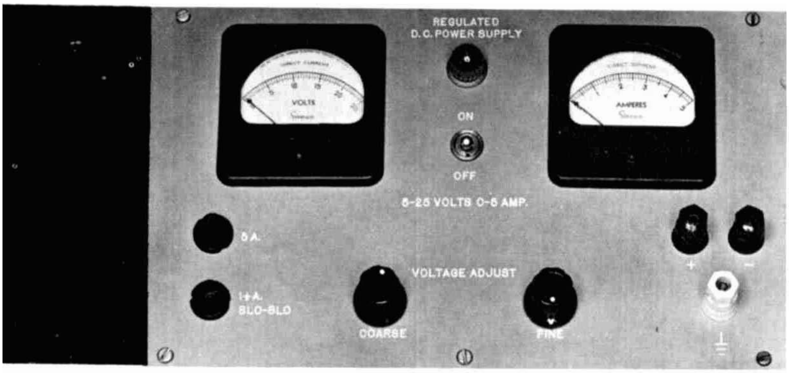
2. Harry D Hooton, W6TYH, Single Side-Band, Editors and Engineers, Indiana, 1967

ham radio



Five to twenty-fiva volts at 5 amps with built-in short-circuit protection.

a modern low-voltage power supply with built-in short-circuit protection

This variable low-voltage power supply provides up to five amps output and won't burn out if you accidentally put a short across the output

Don Nelson, WB2EGZ, 9 Creel Ridge Road, Ashland, New Je 0 34

Without question, the most useful piece of equipment in my shack is the regulated low-voltage power supply. Although my wife teases me about being the only family in the neighborhood with such an elaborate battery charger, nonetheless, it is quite practical in that application. You can see that some parts of this unit are rather old, while others are very new. The explanation is simple: the control circuitry was rebuilt using silicon semiconductors. I have been using the original supply since **1959,** but occasional problems prompted a change. Tests I have run on this new circuit and its components lead me to be very optimistic about their reliability.

circuitry

There is nothing new or unproven in the circuit. It was my privilege to use some new low-cost RCA plastic-encapsulated silicon power transistors which are still relatively unknown. Despite their novelty, you shouldn't have any trouble buying them. Of course, other transistors may be used, but you can't mount them on this size heat sink.

The series-regulating element consists of two RCA 2N5034 power transistors connected in parallel. The gain of this series regulator is multiplied by two Darlington-connected stages using an RCA 2N5295 and an RCA **40311.** RCA has only recently announced the

2N5295, but it has been available to commercial users since April, 1967, as the TA7156. The remaining transistor complement consists of two RCA 40311's in a differential pair.

To reduce the dissipation of the power transistors, a three-position switch applies various input voltages to the circuit according to the output voltage that is desired. To give optimum useful rotation of the fine-voltage control, biasing of the control amplifier is also switched. The three positions are set for: less than 13 volts; 13-19 volts; and 18-25 volts.

A crowbar circuit using an SCR assures that the fuse blows before the transistors. Without some type of current-limiting circuitry or a crowbar, there is a rather large crisis every time a transistorized power supply is short-circuited. I have blown as many as eight components in the old circuit, and the fuse was still as good as new.

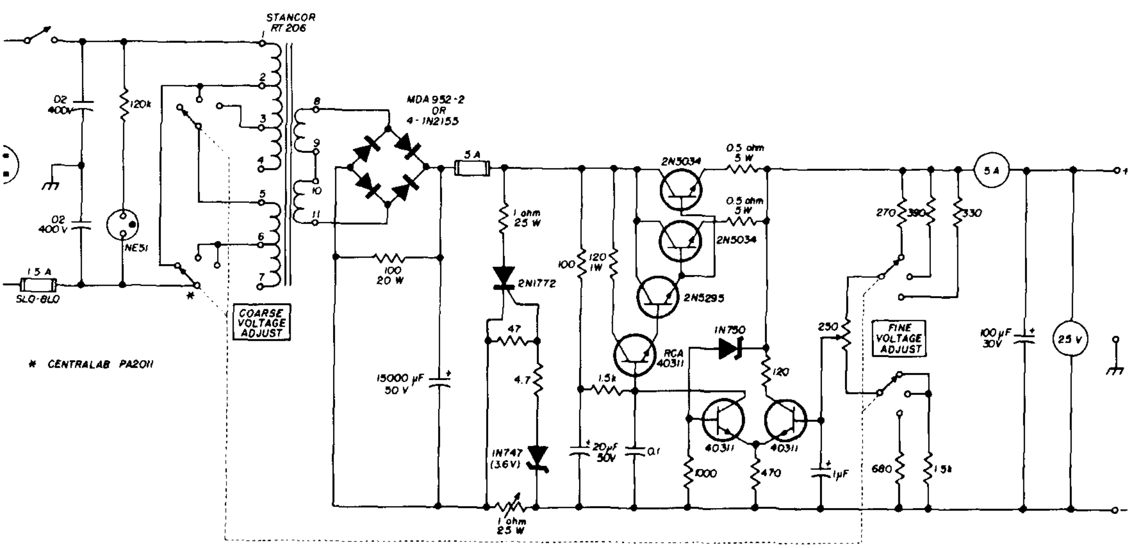
Fuses are much too slow, and the semiconductors are unforgiving. In a low-current supply, automatic current limiting might be preferable. However, at the 5-ampere level, current limiting techniques become impractical, so the simplest alternate, a crowbar, was chosen. The purpose of a crowbar is to melt the fuse by applying a heavy load across the fuse, but not the regulator. This is accomplished by turning on the gate of the SCR when 5 to 6 amperes is flowing through the adjustable l-ohm resistor in the negative side of the supply.

In turn, current flows between the anode and cathode of the SCR until it is interrupted by the fuse. Adjustment of the resistor determines the firing point and is necessary because of differences in components. Don't think you can skip this sophisticated fuse blower–you can't really afford the havoc it prevents.

the crow-bar circuit

A few words on the intricacy of the circuit may help the builder who has trbuble. The SCR will fire when the gate-to-cathode voltage is about 1 volt. According to the data sheet, current shouldn't flow through the 1N747 zener until there is 3.6 volts across it. This would seem to indicate that a 4.6-volt drop across the l-ohm resistor would be needed to fire the SCR. Actually, low-voltage zeners have very high leakage at lower voltages than their rated breakdown. If the leakage is great enough, the SCR will fire at lower power-supply current levels than desired. The solution I used was to have a low-resistance return to the cathode of the SCR. This is also useful in keeping it from firing independently of the gate control–a malady of SCR's and thyratrons. Should your labors evolve a giant

fig. 1. Schematic of the five-amp, low-voltage power supply.



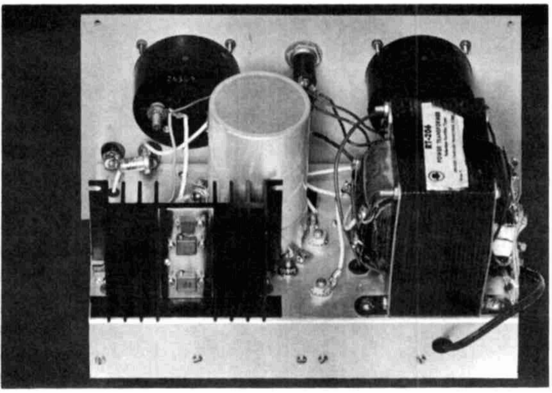
fuze zapper, look for a bad 1N747.

I wish I could use a less-expensive SCR. Unfortunately, experimental results indicate the i^2t* surge rating of the SCR should be at least 75. It is difficult to find the exact rating to use because the i^2t surge rating usually applies to a surge of less than 8.3 milliseconds duration; in this circuit the fuse requires 80 to 600 milliseconds to blow. To compound the problem, I have had a few tenacious fuses that required noticeably longer to melt. Although I shouldn't brag, I have attained a high degree of skill in fuse blowing while perfecting this circuit.

construction

An aluminum chassis was built to fit within a Premier PAC1276 aluminum chassis box. All the large parts are mounted on this chassis. The meters mount directly to the front panel. The voltage controls, fuse holder and chassis binding post secure the chassis and panel by mounting through both.

Inside the power supply. The series-regulating transistors are mounted on a Wakefield NC413K heat sink.



The main chassis acts as a heat sink for the rectifier diodes. Teflon feed-thru insulators are provided with the diodes and must be used. While the 1N2155's do a fine job, a Motorola diode assembly, MDA952-2, would be less expensive if you are buying new parts.

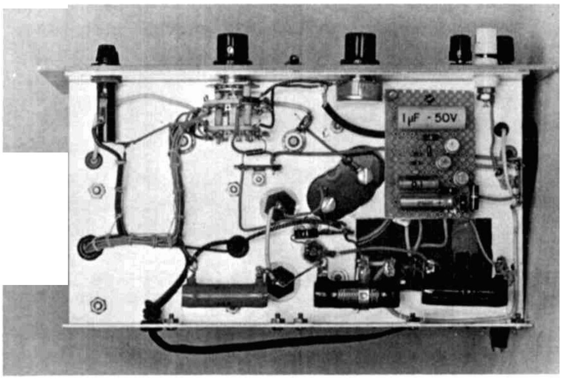
The power transistors are heat sunk to a Wakefield NC413K sink. Since the collectors are connected to the case and insulating washers were not used, the heat sink is electrically isolated from the main chassis by feed-through washers. The regulator amplifier assembly, which is on a phenolic board, is bolted to the main chassis after it is wired. Although I have no plans for heating it up, the SCR is also bolted to the main chassis with insulating hardware.

*The i^2t rating of an SCR is the rate of rise of current when the device is turned on and is measured in A^2s (amperes squared seconds).

some final thoughts

Because of variations in parts, it may be necessary to adjust the values of resistors on the coarse control to place the regulator in the optimum range of control. The general idea of operation of the fine and coarse controls is to allow good regulation with a minimum voltage drop across the power transistors.

Bottom view of the power supply. The control-amplifier board is in the upper right-hand comer.



Excellent regulation and very low ripple characterize the output of this power supply. At certain voltages and high-current levels the heat-sink will get rather hot. Prolonged operation at high temperatures should be avoided. A good rule of thumb: "Keep your heat sink cool enough that you can hold your finger on it."

Much to my dismay, my low-powered transmitter blew fuses when I first hooked it up to this power supply. The problem was remedied when I found that a high value capacitor across the transmitter power input had a charging current far in excess of 5 amperes. In this case, a smaller capacitor was adequate; another application may require adjustment of the 1-ohm resistor which regulates trigger sensitivity.

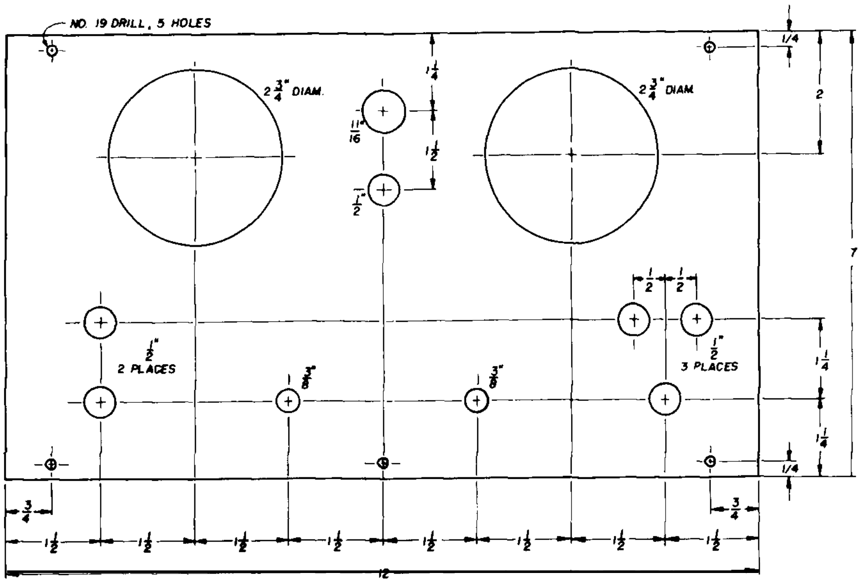


fig. 2. Panel **layout** for the low-voltage power supply.

You will be as enthusiastic as I am when you complete this supply. Have a good stock of fuses for demonstrating. By the way, if you run out of the 5-A variety, the crowbar will melt to 10-A fuse with only a 5-ampere load–just a little more slowly.

ham radio

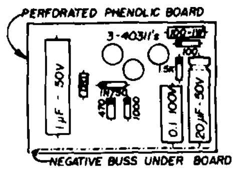
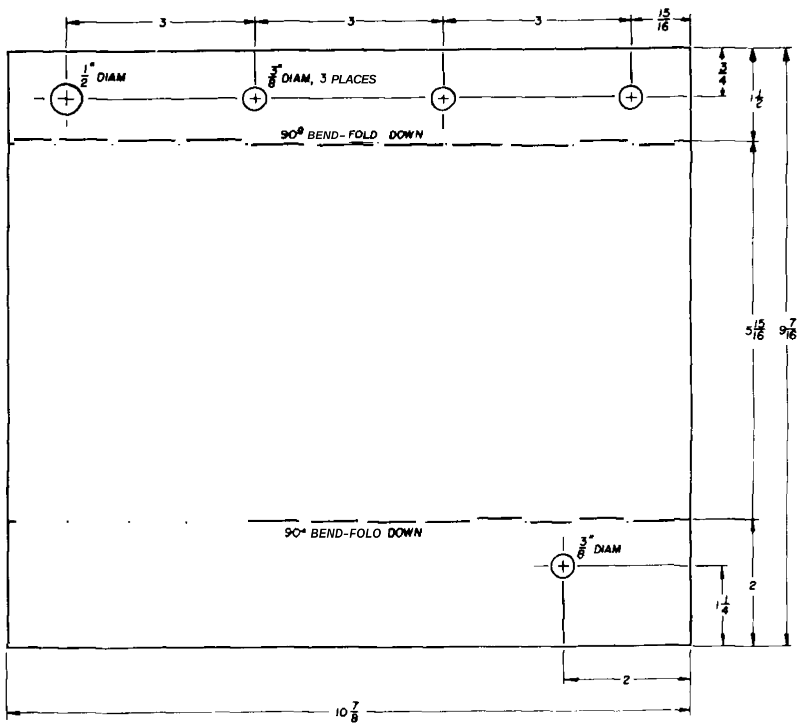


fig. 3. **Layout** of **the** control-amplifier board.

fig. 4. **Power-supply chassis.**



a grand-daddy station-control center

For the discriminating ham who wants the most in station control

Howard S. Pyle, W7OE, 3434 74th Avenue, S. E. Mercer Island, Washington 98040

For the average ham, accumulation of various items of equipment is inevitable as he progresses up the ladder of hamdom. As time goes on, it becomes increasingly awkward to manipulate a dozen or more individual switches to turn the various gear on or off as the need warrants. All too frequently something is overlooked when you close the station after an operating session. This means that the equipment is going to sit and cook for many hours—even overnight! In addition to the wear and tear on the gear, quite often it creates a considerable fire hazard. The thinking ham then commences some thoughtful pondering: what can he do to eliminate the hazard, and, at the same time, improve his operating convenience?

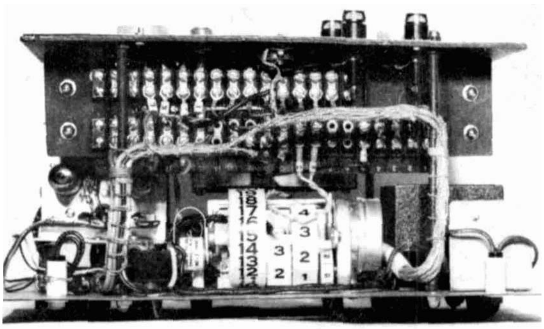
The obvious solution is to take a leaf from a hook of commercial or military installation practice: provide a single, foolproof central control point to insure that every item of equipment in the shack is cold and dead by simply flipping one main switch. In addition to the ac supply, you should provide individual controls for any and all accessories which are ordinarily switchable—ac power circuits, audio input and output levels, CW monitoring—in short, anything that requires frequent switch manipulation. A versatile station-control unit is the answer. I'm going to describe one here which I've been using for several years with complete satisfaction.

the control unit

I'm not going to present this in the form of an actual construction article for a very good reason. No two hams will have identical control problems. Their requirements will vary just as their station equipment does, and as we all know, no two ham shacks are alike! So, I will describe what I've been using successfully to meet my requirements. You can make modifications both in the mechanical layout and wiring to suit your station.

There is probably no other item of ham gear which offers the opportunity for individual design initiative to the extent that a control unit does. You can't buy one on the open market for the same reason I mentioned above; no two ham stations have the same requirements! But, from the photos and schematic of my unit, you can design one to meet what you think will serve your purpose.

Top viaw of the control center showing the digital clock and terminal strips with removable lugs.



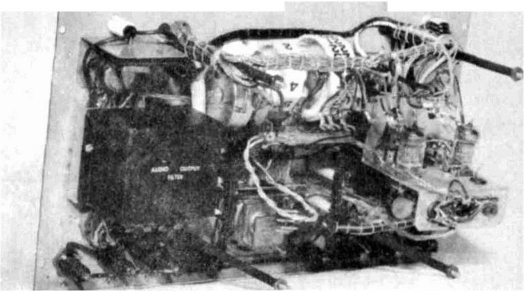
station control

First, what do you want to control from a central position in your shack? What else should you put in the unit for added operating convenience and/or safety? A third point deserves equal attention; where do you propose to locate the control center?

Suppose we take a look at this last point—location. If you want the control unit to be convenient to the operating position, it will probably be next to the actual operating area. For example, in most ham shacks, the transmitter and receiver are located on a desk along with the operating accessories—key, microphone, perhaps a few miscellaneous switches and minor equipment items. Everything is placed so it's convenient to the operator's hand and vision. If the control center is to serve a number of functions as mine does, it must be conveniently located. If it's positioned between the transmitter and receiver, it will he within easy reach of the operating position.

Let's assume that this is where we put it. When it's within easy sight and reach, it can house the items we consider of secondary importance. For example, when the control unit is easily visible to the operator, the station clock could be included. Also, it is desirable to include a series of pilot lights which will serve to indicate at a glance what equipment is "on". This saves a lot of time as you don't have to look at each and every piece of equipment to see what's turned on. A monitor speaker could also be installed in the unit unless adequate speaker response is available from the speaker in your receiver.

Raar side of the front panel with the subpanel removed.



Now, let's go back to the **primary** object of the control center; to provide **control.** By control we mean the ability to not only turn them on and off individually, but to provide additional functions such as switching receiver outputs from phones to speaker, choose the output of several receivers, switch external accessories such as notch-filters, preselectors, converters and similar items in or out of the system. If you use a CW keying monitor, it too can be incorporated within the control unit. Its output may be fed to a speaker or phones at will by suitable switching.

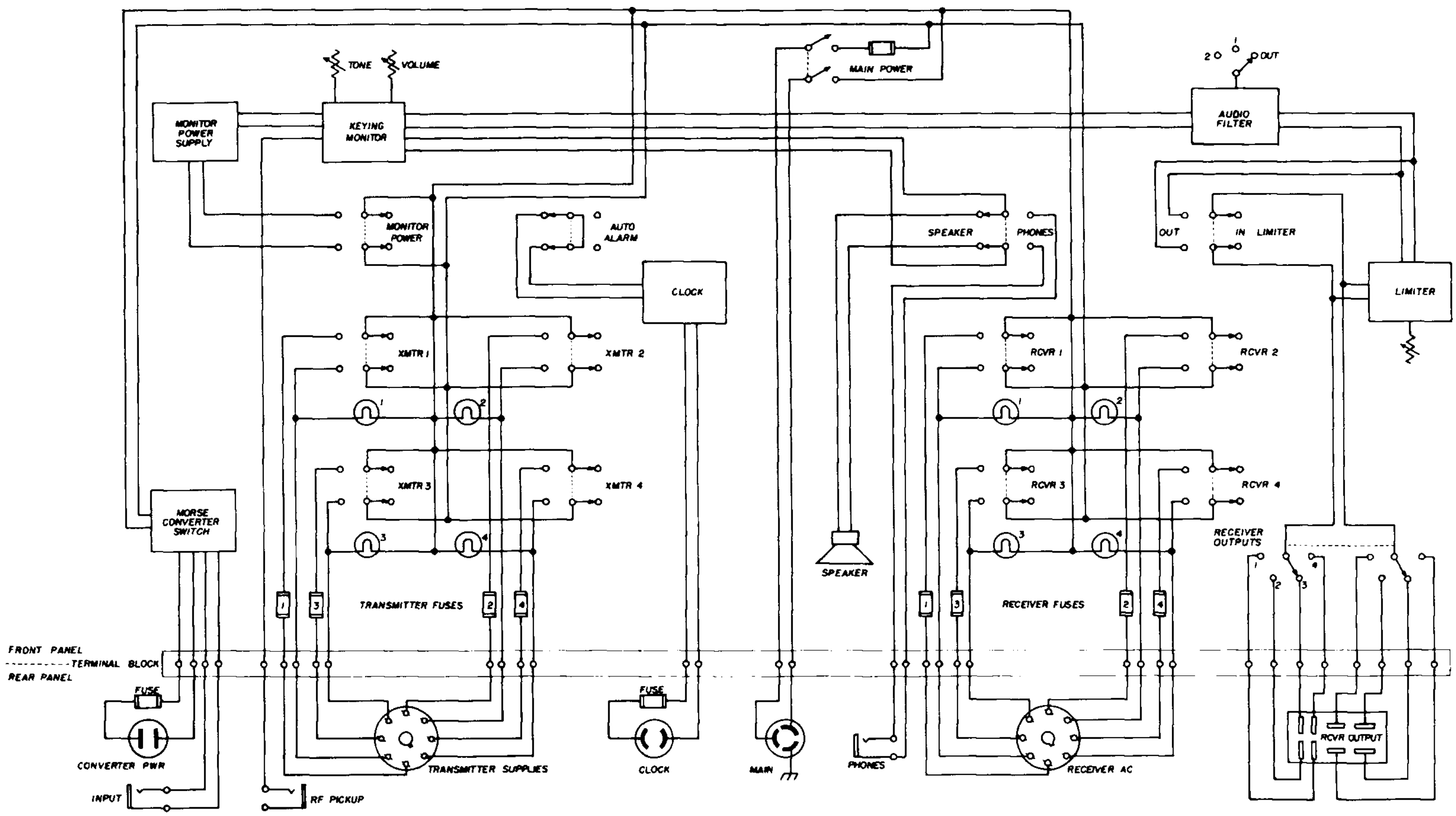


fig. 1. Subpanel and terminal block wiring diagram of the W7OE station-control center.

Turning the equipment on or off from a central point means, of course, that the ac powering the station must pass through the control center. This provides an ideal opportunity to insert a main ac switch in the circuit so that when it's off, power is removed from all equipment. Except, of course, the station clock, if it is one of the electric variety!

Appropriate fusing of all circuits should be provided, and these fuses should be in the control center, not scattered around in the equipment. Since most equipment is fused internally, an over-size fuse should be put in as a replacement with the proper size fuse in the control center. Then, in the event of a blown fuse, it will he in the control center, convenient to replace, and not buried deep in some awkward spot in the chassis!

the *grand-daddy* control center

Now, let's look at the control center which I have been using for several years. This has proven to be the most convenient operating aid in my station.

Suppose we start with the front panel photograph. The panel itself measures 9 x 15 inches and is part of a Bud CU882 steel utility cabinet, with removable back and front panels. These cabinets are available in either black or gray wrinkle finish; I chose black to match the other equipment in the shack.

A 24-hour digital electric clock (Call-Ident Tymeter by Penwood Numechron) takes the top center of the panel. This clock has a buzzer which will sound an alarm every ten minutes as a reminder to identify while you're in QSO. A slide switch on the panel lets you cut this feature in or out as desired. The lower center of the panel has a 2-inch hole backed with screen and grille cloth for the small 3-inch monitor speaker on the rear of the panel.

Slide switch panels (Allied #16B2197) are mounted to the extreme left and right center. Each panel provides six switching operations. The four lower switches on the left-hand panel control the ac supply to four transmitters; the lower four switches on the right-hand side supply ac to four receivers. The upper left-hand switch on the left-hand panel controls the ac supply to the CW keying monitor, and the upper right-hand switch on this panel is the "on-off" switch for the clock alarm.

On the right-hand panel, the switch in the upper left selects either the internal monitor speaker or a head-phone jack on the rear subpanel. The switch on the upper right cuts a diode noise limiter in or out of the audio circuit. Just above this switch is the potentiometer for controlling the audio limiting. To the left is the 10-ampere main fuse for the control center.

To the left of the clock face is a conventional DPST toggle switch for the incoming ac mains. The knob to the left of this switch controls the volume on the CW keying monitor which feeds its output into either the monitor speaker or the headphones through the "speaker-phones" switch. Below the slide-switch panel on the left is the tone control for the CW monitor. The pointer knob to its right selects a Morse-telegraph sounder converter or the speaker or headphones. In the Morse telegraph position, this switch also turns on the ac supply to the converter.

The vertical row of red-jeweled indicator lamps to the left indicate which transmitter is in use; an identical row to the right serves the same purpose for the receivers. The row of fuse holders across the bottom of the panel at the left protect the transmitter ac supplies; a similar row at the bottom right does the same for the receivers. Just above the receiver fuses, the bar knob controls a four-position rotary switch which selects the output of any of the four receivers and feeds it to the monitor speaker (or headphones) after passing through the various audio filters. The bar knob to the right provides a choice of three values of audio filtering in the output.

rear panel

Next, let's look at the rear panel of the control center. This is not the full-sized panel supplied with the cabinet. A chassis bottom plate, 7 x 13 inches, was used here and is supported from the rear of the front panel by four 10-32 threaded rods six inches long. The threaded rods are run through metal sleeves (copper tubing) for greater rigidity and improved appearance. The rear panel is essentially a subpanel and not a part of the cabinet.

Although the photograph of the subpanel should be self-explanatory, there are several points of interest. At the top is an ac socket for the separate incoming ac line feeding the clock circuit. To the right is a chassis receptacle for the rf monitor external pick-up lead. The ac circuits to all four receivers are wired to the octal socket on the left. The "audio-outputs" socket carries the outputs of all four receivers through an appropriate external terminal strip. Next in line is a receptacle to receive the ac supply plug from the Morse-telegraph converter.

The octal socket on the right carries the ac supply to all four transmitters. A jack in the lower right-hand corner takes the audio input of the selected receiver to the Morse-telegraph converter. The 3-pole polarized male plug in the lower center carries the main ac supply for the entire control unit.

internal construction

The space between the two panels accommodates all of the internal components of the center. As these will vary widely from ham to ham, the components I used can only be used as a guideline. Internal parts layout will be different of course, if you use different parts than I did. This is where ingenuity in design and layout will pay off.

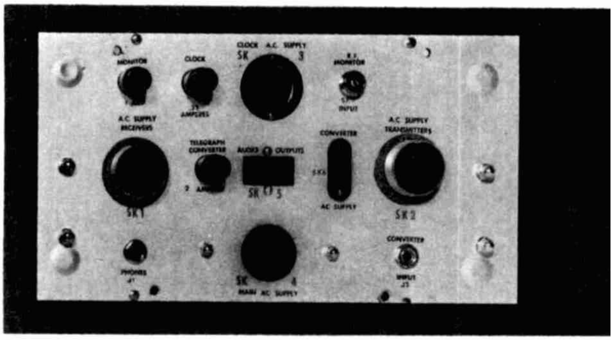
The location of slide-switch panels, rotary switches, potentiometers and other panel-mounted equipment depends somewhat on the interior of the unit. No attempt was made to miniaturize any components; I used parts that I already had on hand as much as possible. Although compactness and a symmetrical panel layout were achieved, there was no overcrowding.

With all panel and subpanel equipment mounted, the remaining space must be used to best advantage to accommodate the rest of the control center package. In my unit, barrier-type terminal strips (Cinch-Jones series 140), permitted breaking all wires between the panel and subpanel. When repairs or modifications are necessary, the two panels can be completely separated by loosening the terminal strip screws on one side and lifting the spade lugs used on each wire. This arrangement also provides more elbow room during installation and wiring of the various components. The internal arrangement I used is shown in the photographs.

wiring

Let's discuss the various parts which make up the unit. First, the main ac input. As shown on the subpanel schematic (fig. 1), a 3-pole male twist-lock connector is used for ac entrance; the third contact is grounded to the subpanel for safety. The ac line passes through the main fuse and switch on the front panel; a "main power" indicator light is provided on the front panel.

The grand-daddy control center. Subpanel face. The various parts as discussed in the text.



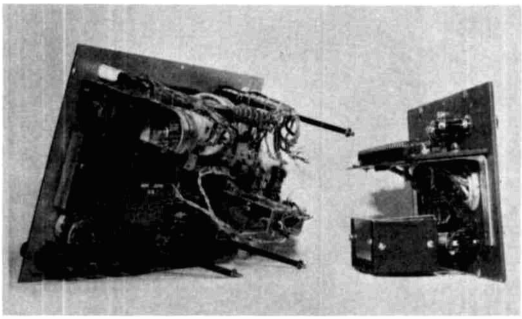
The ac line between the main and subpanel is broken by terminals on the Cinch-Jones terminal strip; from there it is connected to terminals on other strips for distribution to various pieces of equipment. This includes wiring through the slide switches on the front panel to the two octal sockets on the subpanel which supply power to the transmitters and receivers.

While I use four transmitters and four receivers, the average amateur station has only one or two of each. Although you can buy slide-switch panels with fewer switches, I feel that switch panels and sockets should be provided with sufficient capacity to accommodate future additions. Play safe and provide for a minimum of four pieces of major external equipment.

Next, the eight-contact socket in the center of the subpanel. This carries the audio output circuits of all four receivers and is internally wired to one of the terminal strips. From here, the circuitry follows through the noise limiter and audio-filter circuits, speaker-headphone switch, phone jack and the output selector switch on the front panel.

The CW keying monitor in my unit uses rf pickup to activate it. Although I use a **Johnson Signal Sentry,** any suitable keying monitor which you may happen to have may be substituted. Any of the conventional code-practice oscillator-monitors available can be adapted for this use or you can build one if you prefer. A **Cordover** solid-state CW-monitor module occupies less space and re-

The front panel to the left; the rear subpanel to the right.



quires no external ac supply source if you'll be satisfied with an internal battery. I don't favor this approach since it's awkward to open the control-center cabinet periodically for battery replacement. Of course, a small ac supply can be suhstituted.

The noise limiter is a simple diode type which I built and has proven very satisfactory. The audio filter is a surplus military unit originally designed for aeronautical use. This is useful for narrowing the audio channel during extreme interference conditions for CW reception. If you have suitable audio filtering in your receiver, either or both of these audio devices can be eliminated. Also, unless you are a Morse telegrapher, you probably won't want the controls for the telegraph converter.

summary

This pretty well covers the **grand-daddy** control center, hut a few additional tips may be helpful. For instance, the switches and fuses I used may not have aclequate carrying capacity if your transmitter(s) run at very high power. Check the ampere capacity of any switches you want to use and put in fuse values which agree with your current demand. Since my maximum transmitter input power is only sixty watts, the values shown here are more than adequate. I would suggest too that you use **shielded** wire for all of your audio circuitry; this will prevent hum pick-up from adjacent ac wiring in the cabinet. And, for a really professional job, lace your wiring runs into cabled harnesses.

This should give you a number of ideas for increasing the operating convenience as well as the safety of your ham shack. Carry it even further—don't stop with a nice control unit and then connect the external wiring by cords and cables, straggling all over the shack. Do the **external** wiring job neatly as well: make straight wire and cable runs with rounded right angle turns. Whenever possible, use small cable clamps to strap cable runs to the wall, table or bench. You'll get a great deal of pride and operating pleasure during your on-the-air sessions and you can point to your installation with pride when visitors arrive!

ham radio

next month in ham radio magazine:

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Simple Transistor Checker
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Mini-Monopole Antenna
How to use the Smith Chart
Two-Meter Mobile
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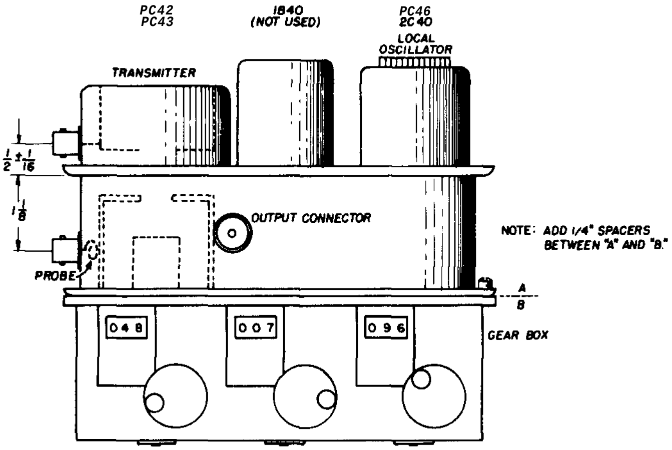
Plus many more!



notes on the APX-6 transponder

The APX-6 still provides the easiest way to get on 1296 MHz. Here are some new conversion techniques, including an FET i-f preamplifier

Les Maurer, W6OSA, 209 Nob Hill Way, Los Gatos, California 95030

There have been a couple of good conversion articles on the APX-6[1,2] and it's not my purpose to repeat all the information contained in these past articles. However, I want to amplify this information with additional knowledge that I gained the hard way—the conversion itself. The ubiquitous APX-6 is still the quickest and most inexpensive way to get on 1296; and, when you graduate to more sophisticated equipment for transmitting and receiving, you can always use it as a signal generator or grid-dip meter.[3]

the receiver

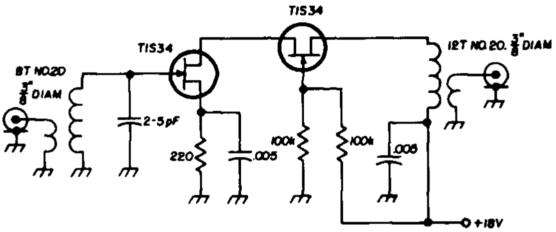
The receiver is the first order of business. There is nothing sacred about the 60-MHz i-f strip that comes with the transceiver. It was designed to amplify pulses—and that's what it does best. If you don't believe me, clip an antenna to the input and listen to the cars go by.

I get much better results with a 50-MHz first i-f and a 15-MHz second i-f. Probably the most important thing in my i-f amplifier is the high-gain low-noise preamp inserted between it and the APX-6 diode mixer.

I started out with a 6BQ7A cascode preamplifier. Later I replaced the 6BQ7A with a 6ES8. Finally, I replaced the 6ES8 with a pair of TIS-34 field-effect transistors, shown in the circuit in **fig. 2.** When there is no rf amplifier ahead of the mixer, the first i-f stage becomes all important. The TIS-34 FET's are about the ultimate at 50 MHz (inexpensive too).

You can use any combination of i-f frequencies from 144 MHz on down, depending on what you have on hand. However, use a cascode circuit, preferably an FET cascode, for the preamplifier. The other point to remember is that a very sharp i-f (such as the 455 kHz i-f in your low-band receiver) will not give satisfactory results with the APX-6. The 2C46 local oscillator has too much drift and frequency moding.

fig. 2. FET i-f preamplifier for use with the APX-6.



The combination of 39 and 4.7 MHz in the surplus Link receivers works out quite well. The 19- and 3.45-MHz i-f's in the guard receiver are also good. You can also use a single-frequency i-f in the 10- to 20-MHz region. The 1N25 diode goes for a pretty stiff price, even surplus. The 1N21 or 1N23 Series work equally well and are much more available.

The 2C46 in the local oscillator can be replaced with a 2C40 or a 446A. All you need to do is wind a strip of copper around the plate cap to make it as fat as the cap on the 2C46. While I'm on the subject of tubes, the 2C43 can be run at higher power than the 2C42 which was originally used in the transmitter, and it is in more plentiful supply.

All three of the APX-6 tuning pistons had to be shortened by 1/4 inch to cover the high end of the 1296 band. I used an Exacto razor saw (sold in hobby shops). As a temporary expedient, you can jack the cavity chamber up 1/4 inch above the gear chassis with washers or metal spacers and accomplish the same result **(fig. 1).**

No one has ever mentioned why you can't use the original 1B40 for a T/R tube instead of all that fuss with a neon bulb. The answer is that originally 450 volts or more was used on the plate of the 2C42 for pulse operation. If you only have a 300-volt power supply, as shown in both conversions, you wouldn't be able to ignite the 1B40.

the transmitter

With regard to cavity modifications, I defy anyone to use the 1/4-inch centers shown for the BNC fitting on the cathode cavity,[1] get the clamping ring under the connector and be able to plug in the male connector too! The original **QST** article shows 9/16 inch from the center of the BNC to the top of the cathode cavity. I recommend 1/2 inch plus or minus 1/16 from the center of the BNC connector down to the lip of the cathode cavity. Take an ordinary BNC connector and saw one

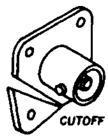
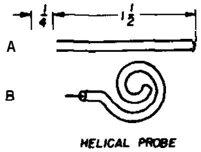


fig. 3. Modifying a BNC connector for use on the APX-6.

fig. 4. Improved results may be obtained by using this helical pickup probe.



corner of the flange off as shown in **fig.** 3; that beats grinding off the flange as one article suggested. Be sure to put the retaining ring on before soldering the BNC connector to the outside of the cathode cavity; it fits under the connector where the flange is cut away.

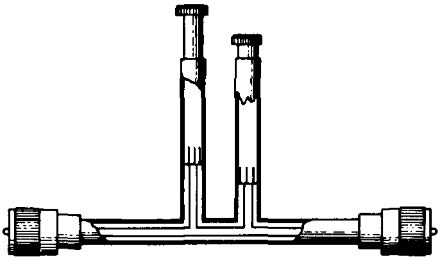
Here is another approach to coupling the BNC connector to the plate cavity of the transmitter. I seem to have better luck with capacitive coupling than with inductive coupling on all my UHF power sources. I tried it in this application and it worked like a charm. The design of the coupling capacitor is a W6OSA original. I don't know *why it* works — but it does.

Take a piece of insulated #18 or #20 wire 1-112-inches long and strip 1/4 inch of the insulation off one end. Solder this bare end to the tip of the inner conductor of a BNC bulkhead-type connector. Starting from the insulated end, coil one inch of the wire into a small helix as shown in fig. 4. The helix now becomes a capacitive probe. The insulation prevents accidental shorts to the plate line.

I use #20 stranded wire for my probes. Stranded wire is not supposed to work—but it does here! Stuff the helix through the 3/8-inch hole you drilled in the plate cavity.[1,2] Adjust the BNC connector in and out until you get maximum output and then clamp or solder it in this position.

I used to wonder why the transmitter oscillator wasn't built like the receiver local oscillator. I tried a 2C43 and a 2C42 in the local oscillator cavity in place of the 2C46 and they oscillated very nicely. I couldn't get much power out through the regular antenna connector, but the coupling to the local oscillator from the antenna cavity is intentionally very loose. By installing a new coupling link,[1] it should be possible to use this cavity for a transmitter. Then, if you had two sets of APX-6 cavities, you could use one for transmitting and one for receiving.

fig. 5. A surplus line stretcher that I use to obtain a greater frequency range in the 1215-MHz bend.



The modified transmitter with the 7.7-inch coax link will only work over a small range of frequencies. This is pretty exasperating if you are trying to hit a specific frequency such as 1296 MHz. The *7.7* inches is derived from multiplying 11-1/2 inches (312 wavelengths at 1220 MHz) times the velocity factor of RG-59/U coaxial cable.

However, 11-112 inches is 3/2 waves long at 1220 only if you add the length of the BNC's and the conductors inside the cavities. Hence the voltage at the plate end is out of phase with the voltage at the cathode end. You could use any odd number of 1/2 waves except one one-half wave which won't reach. To increase or decrease the oscillator frequency very much, you have to change the length of the coax link.

The only way around this is to use a "line stretcher". This device (fig. *5)* will let you move anywhere in the band and still operate at peak efficiency.

The modified transmitter cavity can also be used as a straight-through amplifier on 1296 MHz. A one-watt varactor into the APX-6 2C43 amplifier would make a good combination. Simply run 1296-MHz drive (about 1 watt) straight into the cathode cavity. The output can be taken from the original output connector, but you will get better results with the new capacitive probe.

antenna

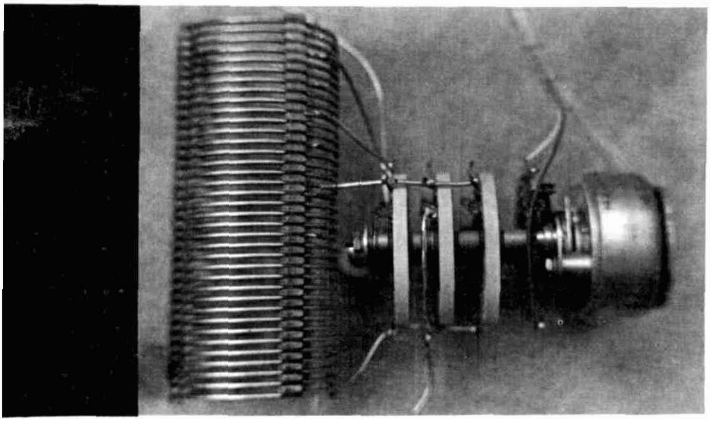
My antenna is a 16-element expanded-extended collinear. The reflector is made of copper window screen spaced one-quarter wavelength behind the eight driven elements. The array is fed with foam-filled tubular twin lead and a Frank Jones 50 to 300-ohm balun down in the shack. The twin lead works fine except when it rains.

There are two DME (Distance Measuring Equipment) stations about 30 miles away between 1204 and 1214 MHz. I use them to check how well the complete receiving system is working before trying to make a contact on 1296 MHz. I believe that every major airport has DME equipment operating somewhere between 1150 and 1215 MHz. I hear radar stations too, but they are so loud they don't afford much of a check. Oh yes, be certain you and the other station are using the same antenna polarity (W6GHV please note). The DME's are apparently vertical.

references

1. B. Brown and A. Katz, "Putting the APX-6 on 1215 Mc," Part I, CQ, October 1965, page 53. Part 2, CQ, November 1965, page 46.
2. E. Tilton, "Communication on 1215 Mc with the APX-6," QST, September 1960, page 31.
3. J. Fisk, "A 1296 Grid Dipper," 73, June 1967.

ham radio



A solenoid rotary switch which is used for selecting the taps on a multiband antenna loading coil.

how to use solenoid rotary switches

John J. Schultz, W2EEY/1, 40 Rossie Street, stic, Connecticut 06355

If you want to remotely control antenna-loading coils or feedlines, why not try a solenoid rotary switch?

Solenoid rotary selectors can be used to simplify a multitude of remote-control functions and are often available on the surplus market. They can be used to advantage in many ham shacks; you can use them to select the proper loading inductance for a vertical antenna from inside the shack or to remotely switch one transmission line to a variety of antennas. Remotely tuning a vertical antenna only saves labor and inconvenience, but when they're used to remotely switch coaxial feedline, the savings are more tangible. This is particularly true if you use many long transmission lines at your station. In more advanced installations, it may even be desirable to remotely select antenna preamplifiers or mast-mounted converters.

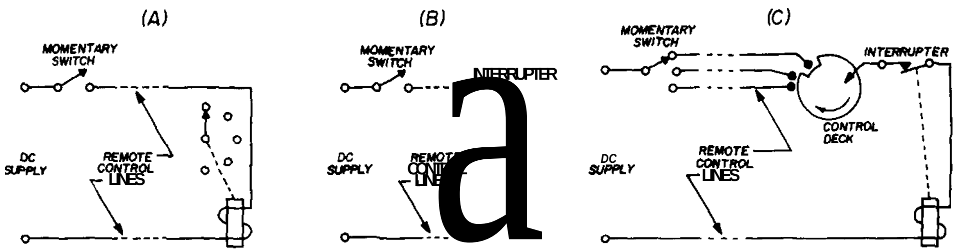
Although many elaborate systems of remote switching have been developed, the easiest and most versatile method entails the use of solenoid-operated rotary-selector switches. They are simple to control, dependable, and will meet any switching function that can be accomplished with a manual rotary switch. This article describes the basic operation of the solenoid switch and some of the control circuits which may be used. I have also given some examples of how I use these devices around my shack—remotely controlling multiband antenna-loading coils and coaxial-cable switching.

basic rotary selectors

Basically, the solenoid switch is nothing more than its name implies—a solenoid-actuated rotary switch. However, a few refinements are necessary to make it operate properly. First of all, the solenoid mechanism is connected to the switch shaft through a ratchet. It's constructed so the shaft advances one position each time the solenoid is energized. The ratchet holds the switch in position when the solenoid is de-energized. You advance the switch to the desired position by simply energizing and de-energizing the solenoid.

If it's desired, a self-advancing or self-stepping action can be easily added. With this modification, the shaft rotates from position to position as long as the solenoid circuit is on. The "interrupter" which accom-

fig. 1. Basic remote-control rotary-switch circuits.



plishes this is mounted on the switch shaft. Normally it's closed. However, it opens up when the switch starts to move. When it reaches the next switch position, it's closed again. Therefore, the interrupter automatically "steps" the switch, position by position, as long as external power is applied.

The solenoid which does the work operates on direct current and is a high-torque, low duty cycle device. Most of them are designed for a 10:1 time-off to time-on operation, so for most amateur applications there is no danger of exceeding their ratings.

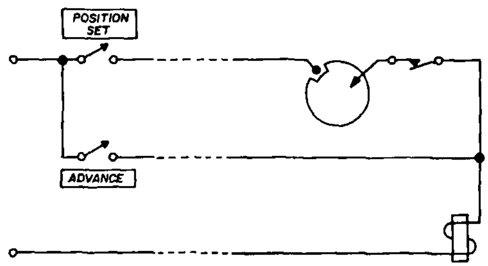
You can often find these switches with ten or more switching decks. In addition to the switching arrangements (number of poles and positions), you should look at the contact current rating and switch insulation. Phenolic switches are suitable for most applications including rf up to 30 MHz or so. Ceramic or epoxy-glass switches are better for VHF or where low-level signals are being switched in high-impedance circuits.

operational circuits

Fig. 1A shows the basic method of wiring up the solenoid with a momentary-contact switch. The use of the interrupter is shown in **fig. 1B.** This circuit is not used in practice because the rotary switch would rotate continuously whenever power was applied. Since each step takes place in a fraction of a second, you would never know what position the switch was in. However, a "control" switch on the rotating shaft as shown in **fig. 1C** may be used to keep things in order.

This control deck has no electrical connection to any of the decks which you are using for remote control. If the selector switch energizes a line to the control deck, the rotary switch will advance until the notch in the control deck corresponds to the ener-

fig. 2. Modified control circuits using a reference or set position.



gized line. Then it will stop. With this arrangement, you can choose which position you want, and the switch automatically steps to that position. The only disadvantage is that a control line is required for each position of the switch.

As you might have guessed, various schemes have been developed to reduce the number of control lines. One arrangement is shown in **fig. 2A.** Here only two control lines

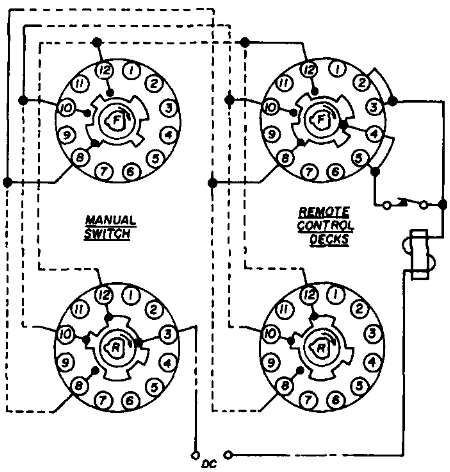


fig. 4. **Methods of arc suppression which are described in the text.**

receiver is turned on. Fortunately, these arcs can be suppressed by placing a resistor-capacitor network across the switch contacts (fig. 3A). This type of suppressor will eliminate the effect of the arc, but not its cause.

A diode or a double-anode zener may be used to eliminate the cause (fig. 3B and C). The diode is installed so that it is reverse biased when the solenoid is energized. However, the back voltage generated by the solenoid when it is de-energized is shorted to ground. Remember though, the back emf generated by the collapsing field around the solenoid can be many times greater than the applied voltage. It is standard practice in military applications to use 500-PIV, 500-mA diodes for 28-Vdc solenoids and relays.

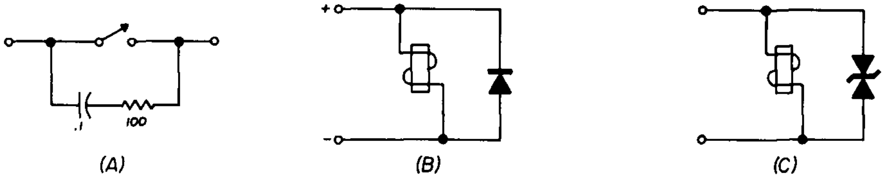


fig. 3. **Modified control circuit using binary coded control.**

are required for any number of rotary positions. When the "position set" switch is closed, the rotary switch will advance to the reference position and stop. It may then be manually advanced by pushing the "advance" switch. This arrangement is real handy if you lose count or don't know what position the switch is in. Just return it to its reference position and start over again.

If a binary-coded control deck is used, up to twelve positions may be controlled with only three control lines (fig. 2B). Both the remote switch and the control deck on the rotary switch have identical contact arrangements. If you trace out the circuit, you'll see that the action is the same as that shown in fig. 1C. This binary idea can be expanded to include more switch positions by changing the layout of the control decks.

arc suppression

Since the solenoid is a large inductor, when it is de-energized a high back voltage is generated. This can cause arcs which may damage the control-switch or interrupter contacts. Also, the rf interference may be very annoying if the switch is operated when your

The zener diode limits the back voltage to a preset level. If its rated voltage is about 10% higher than the applied voltage, it will not affect the applied voltage. Although the diodes tend to slow down the action of the switch, it will still be quicker than your eye.

applications

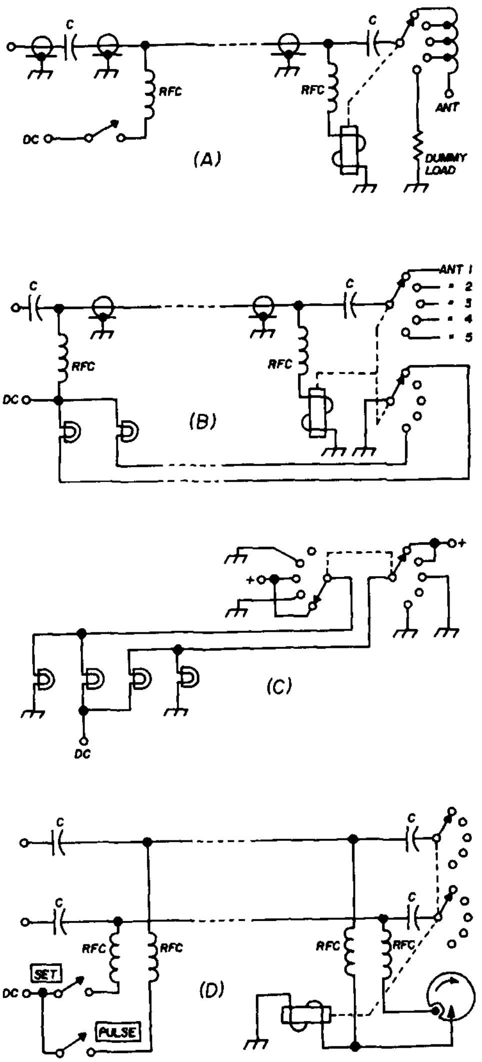
The photograph at the beginning of the article shows a remotely-controlled rotary switch which selects taps on a multiband antenna-loading coil. This is a natural for a base-loaded vertical fed with coaxial cable. The switch in the picture has three switch decks. Since only one switch was required for this job, the contacts of the other two decks were wired in parallel with those on the first deck for greater current-carrying capacity.

As shown in fig. 4A, the coaxial cable which is being switched can be used to carry the control voltage to the solenoid. Blocking capacitors and rf chokes are used to isolate the dc control circuit from the rf path. Since the solenoid is only energized momentarily, 2.5-mH, 250-mA, receiving-type rf chokes are perfectly satisfactory. The capacitors are 0.01 μF mica or disc ceramic.

because of the added resistance of the rf chokes, it is sometimes necessary to use a slightly higher dc control voltage. If the coil is rated at 110 Vdc for example, 130 volts is often used. This is easily supplied by a 110 Vac isolation transformer and solid-state rectifier. No special filtering of the control-voltage supply is necessary.

In the circuit of **fig. 4A,** a momentary-contact toggle switch is used to pulse the control voltage for each step of the rotary switch.

fig. 5. Methods of controlling a single coaxial line (A), selective positive position indication (B and C) and an application to balanced transmission line (C).



The outstanding disadvantage of this circuit is that you don't have positive control of the position the switch is in. That is, you don't always know what position the remote switch is in. The only indication you have is the way your transmitter loads. If you connect a dummy load at one of the switch positions and use an SWR meter in the circuit, this can give you a reference point, but still it isn't the best way to do it.

A better arrangement is shown in **fig. 4B.** This requires an additional two-conductor cable and use of the extra contacts on the rotary switch, but it provides a lamp indication when the switch is in the first or last position. Alternately, the extra cable can be used to wire in the control circuit of **fig. 2A** if the rotary switch is set up with a control deck.

Another interesting position indicator is shown in **fig. 4C.** If there are enough extra contacts on the rotary switch, a two-conductor cable and four indicator lamps will show the switch position. With this circuit, a unique combination of lamps will be energized for each switch position. This diagram only shows the indicator-circuit switch contacts; the rest of the circuit is the same as **fig. 4B.**

If there is a good continuous ground between the remote and control locations, the circuit of **fig. 2A** can be handled by a balanced transmission line as shown in **fig. 4D.** The "set" and "pulse" switches are momentary-contact or push-button types. For tower-mounted antennas, the continuous ground may be provided by the tower. For antennas which are mounted in the attic or near a building, the house wiring conduit may be used.

summary

There are many applications for remotely-controlled solenoid switches in the amateur station. In addition to simple antenna-switching chores, they can be used to control antenna coupler tuning or equipment switching. If you compare the prices of these switches with the cost of separate transmission lines, they often come out far ahead. This is particularly true when you can find a suitable unit on the surplus market.

ham radio

the E-Z impedance bridge

A handy gadget for the active ham–a combination antenna impedance bridge and field-strength meter

Robert Starks, WA9QJP, 1818 Pennsylvania Street, Columbus, Ind. 47201

Here is a little gadget I think you will find very useful.' If you are halfway active, and like to play around with various and sundry items, such as homebrew antennas, it's just the thing for you—a simple, easy-to-build, impedance-measuring device. You can put it together with what you have around the shack in the proverbial junk-box.

construction

I'm sure almost everyone has some sort of a meter lying around that can be used. I used a 3-inch, 200 microammeter that gives plenty of sensitivity. It's easier to read a dip on a large meter than on a small one.

The case is a Bud CMA-1936 standard aluminum meter case; cost about $1.50. You don't have to use this particular case, but I felt that it was the best solution to the problem since you don't have to cut out a hole for the meter and it will hold all of the necessary components very nicely. In addition, this chassis provides two flat sides, front and top, so you can mount the switch and dial to suit yourself.

The lettering on the impedance dial and cabinet are rub-on letters. You can get these

'Based on a design used in the Knight-Kit Bridge marketed several years ago.

in sets for radio test equipment, such as hi-fi or whatever you want. They make a project look very professional. Just refer to the "dry transfer lettering" section in most electronics catalogs. After the job was finished, I sprayed the whole thing with the clear spray. This protects both the lettering and the paint job.

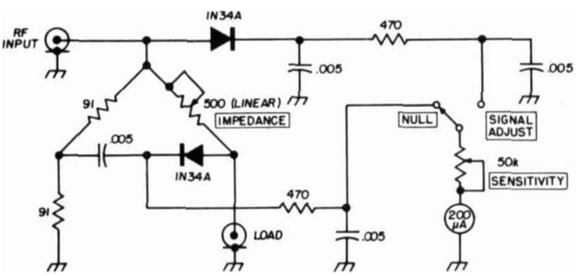
The dial consists of a piece of stiff, glossy, white paper. It is mounted so that the nut which holds the control also holds the dial on. Better still, use double-sided tape or rubber cement. Use your imagination. Put the calibration marks on with India ink, or use strips of the border from the rub-on letters. A piece of plexiglass about 1116-inch thick is rut the same diameter as the dial with a 3/8-inch hole in the center and placed under the nut that retains the pot.

electrical circuit

For a meter-sensitivity control, I used a 50k pot. It gave me plenty of adjustment and at the same time had enough resistance to protect the meter when the control was wide open. The switch is a rocker-type SPDT that I picked up at the local hamfest. It looks nice, but it's more difficult to mount than a slide switch, and doesn't really work any better.

The control should be a 500-ohm linear-taper carbon pot: linear to keep the calibration from bunching up on one end; carbon to minimize reactance in the circuit such as you would get with a wire-wound pot. The same thing applies to the rest of the resistors. All must be carbon. The tolerance of the components doesn't make too much difference, once you get the thing calibrated. However, the two 91-ohm resistors in the bridge circuit (fig. 1) should be closely matched. You are depending on the accuracy of 'the bridge to determine the impedance of the circuit or antenna you are trying to measure.

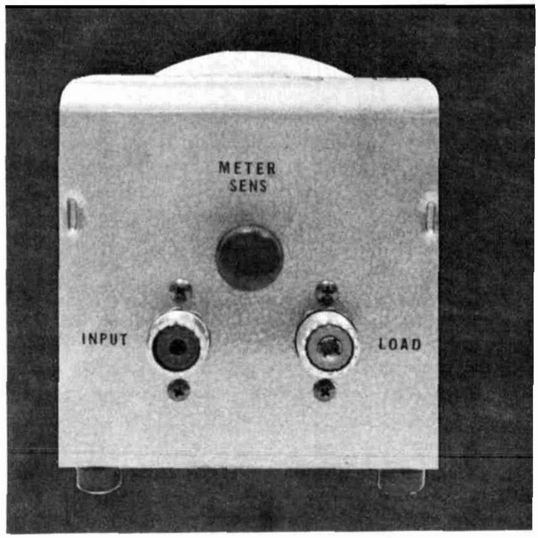
fig. 1. Wiring diagram for the E-Z impedance bridge.



The diodes can be almost anything you have around, such as 1N34A's. The capacitors are .005 microfarads at 600 volts. These are the disc type that you can usually scrounge out of an old chassis.

One thing you'll have to watch: build the circuit as symmetrically as possible in the chassis you decide to use: This is for electrical reasons rather than aesthetic ones. By using terminal strips, I was able to wire in most of the components point to point and still maintain a symmetrical layout You'll probably find it necessary to extend some of the leads, so for this purpose you should use at least #16 solid buss wire. This will keep the components from bouncing around while you are trying to make a measurement. Also, try to keep the leads as far from each other as possible.

Back view of the bridge showing the two coaxial connectors and meter-sensitivity control.



The connectors on my unit are the UHF type. While they are not required, they are fairly universal. Use what you have around, but be sure to keep them pretty close together on the rear of the cabinet. I was able to mount the meter sensitivity control just above the connectors on the rear of the cabinet, but this will depend on the size of the control you have. It's actual placement is not critical.

calibration

The impedance dial should cover from 30 to over 600 ohms. This is enough to cover the normal range of impedances that you are likely to run across. The actual range of measurement will probably be about twice the value of the impedance pot. With a 500-ohm control, you may be able to calibrate the meter as high as 1000 ohms. In order to calibrate the meter, you'll have to come up with some close tolerance resistors. The more accurate the resistors, the more accurate the meter will be. You can buy precision resistors at various ham supply houses, or you can play it cool, and pick up some military-grade jobs at the next hamfest. At any rate, if you have 39- and 56-ohm, 5% resistors, you can pretty well decide where 50 ohms is on your dial. Just use the old noggin. I haven't met a ham yet that couldn't make anything out of almost nothing.

When you have picked up some suitable resistors, fasten one of them across the load jack, starting at the low end, so you will have enough room for all of the calibration points. Couple in a signal with a grid-dip meter or other rf source by means of link coupling and adjust the meter sensitivity control to obtain a suitable reading.

Tune the impedance control through its range until you get a dip on the meter. Mark the dial where the dip occurred and continue with the other resistors until you get the desired calibration. I used six resistors with good results to cover the range I wanted. By the way, I found that, within limits, it doesn't make much difference what frequency you use for calibration. Checks at 4 MHz and 50 MHz showed no appreciable difference.

operation

You can use a grid-dip meter, a signal generator or transmitter for a signal source. However, if you use a transmitter, you'll have to be careful to couple in only enough signal to get a dip on the meter. You should use a non-radiating load with the transmitter, and couple into the unit with a loop close to the final. Use the lowest possible signal level where you can still get a dip.

With the unit in the "null" position and a signal coupled in, attach your antenna or coax to the load jack, and tune the control through its rotation until you get a dip on the meter. There should only be one dip. If you have wired the unit correctly, the dial should read very close to the impedance at the connector. This is the impedance your transmitter will see. For antenna measurements, the bridge should be used at the antenna itself.

To adjust an antenna to resonance, simply set the meter to the impedance you want, couple in the drive signal, and tune the antenna until you get a dip on the meter. To find the resonant frequency of an antenna at the proper impedance, set the impedance dial to the impedance of the antenna and tune the signal source through the range of the antenna until you get a dip. The more resistive the antenna is, the greater the dip will be.

If the antenna is reactive, the dip will not be nearly as pronounced; it will be much broader than it is with a carbon-resistor load. If you can't get a dip, check to see if you have reversed the connections on the rear of the cabinet or have a short or open in the setup somewhere. Also, make sure your frequency source is free of harmonics.

The other position on the switch is "signal adjust". You can use this position for field-strength measurements to peak up your transmitter.

I'm sure you will derive a lot of pleasure out of this little project. I know I did. Especially since it worked the very first time. Many thanks to K9VXL for his very able assistance.

ham radio

Motorola MPS transistors

Plastic transistors carrying MPS numbers below MPS6500 are made by Motorola, and are similar to 2N transistors carrying the same number. MPS stands for Motorola Plastic Silicon, and numbers over 6500 are special transistor types.

W2DXH

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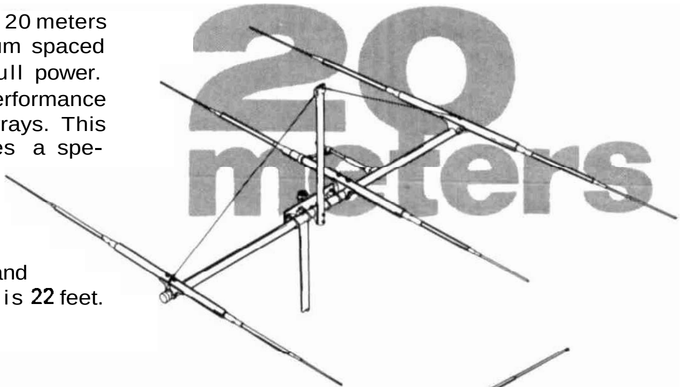

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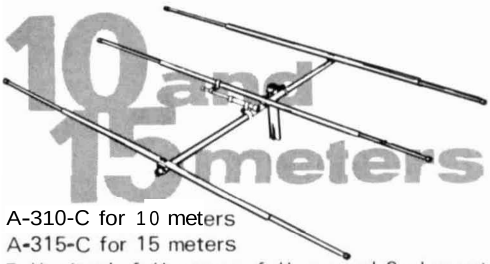

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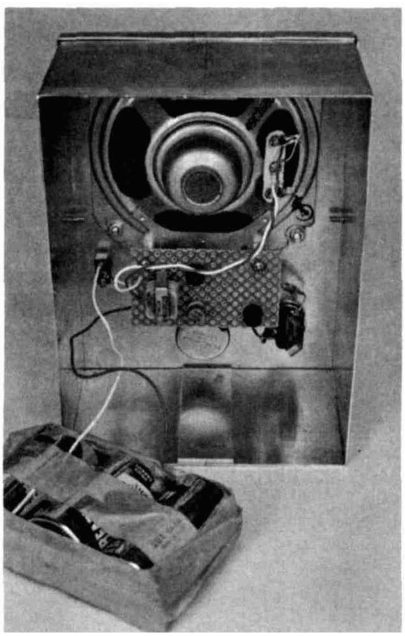
another CW monitor and code-practice oscillator

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

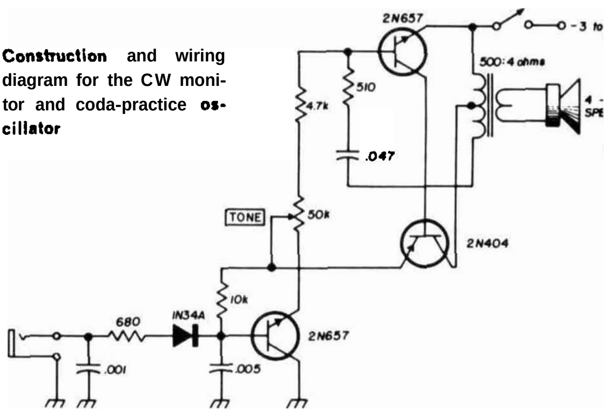
If you ever operate CW, you know the necessity for a good CW monitor. Here is one which can be used with any transmitter that uses grid block keying. It can also be used for CW practice. Although many CW monitors have appeared in print, this one is very practical and simple.

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 2N657's (NPN) 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.



Construction and wiring diagram for the CW monitor and coda-practice oscillator

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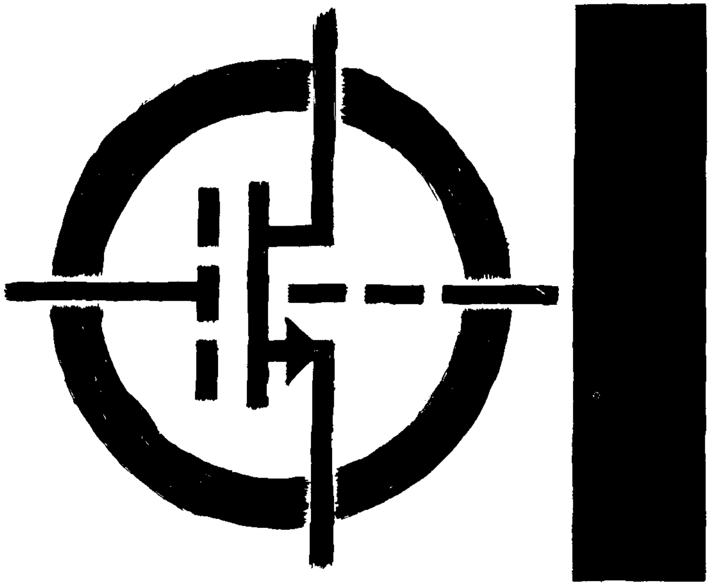


the dynistor

a four-terminal active device

Hank Olson, W6GXN, 3780 Starr Ki g Circle, Palo Alto, California 94306

In recent years the semiconductor field has been pushing the frontiers of technology ahead at an ever-increasing rate. First came the point-contact transistor; then the germanium-junction transistor, the silicon-junction transistor, the junction field-effect transistor, and the insulated-gate field-effect transistor. The monolithic and hybrid combinations of these devices have created our gigantic integrated-circuit industry—and all of this has happened in a short twenty years.

Technological progress has not always chronologically followed the theoretical physics that made semiconductor devices possible. It is well known that the field-effect transistor was completely established theoretically before the bipolar transistor. In another reverse-twist of theory and technology, the **dynistor** has been created; a device that depends on the physics of secondary emission and solid-state surfaces; an exciting active device to be used in circuit design.

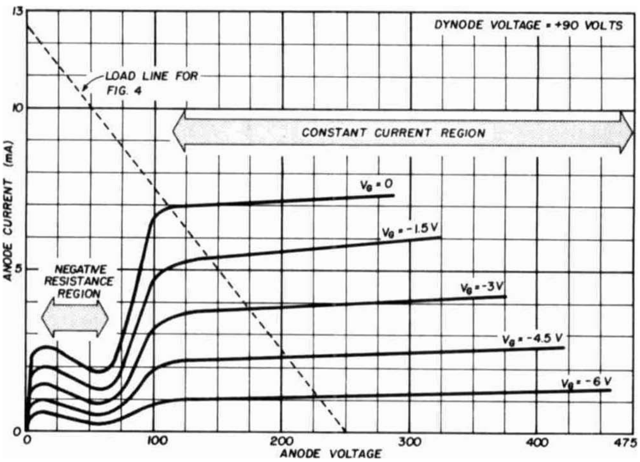
the dynistor

The dynistor, shown schematically in fig. **2,** has four electrodes: cathode, gate, dynode, and anode. The unique characteristics of the dynistor depend upon the dynode. It is **nor-**

mally operated in the depletion mode; that is, with the gate-to-cathode potential negative. However, the gate may be forward-biased slightly (a few volts) without permanent damage, if gate dissipation limits are observed. To take advantage of the "negative resistance" mode, the dynode is operated at a potential higher than the anode.

However, when the dynode is at a lower voltage than the anode, the dynamic characteristics of the dynistor are similar to those of a field effect transistor. This is shown in

fig. 1. Characteristic curves of the UY224 dynistor.



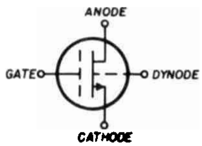


fig. 2. Schematic symbol for the dynistor shows its four electrodes: cathode, gate, dynode and anode.

fig. 1. Note that in this "constant-current" mode, the voltages are higher than most conventional FET's.

One great advantage of the dynistor is its ruggedness. Because of the inherent internal structure of the device, local heating of any one electrode will not destroy it in a few milliseconds. This factor, coupled with the inherent high-voltage tolerance, makes it electrically very durable, unlike bipolar transistors and FET's.

applications

The applications for the dynistor fall into two categories: those where the "negative-resistance" mode is used, and those where the "constant current" mode is used. In the first category are the various types of oscillatory circuits. Fig. 3 shows a simple 1.8-MHz oscillator; note that no cathode-tap is needed (as in a Colpitts or Hartley circuit). It can be recognized that this oscillator is very similar to a negative-resistance oscillator using a tunnel-diode. The details of how a negative-resistance oscillator sustains oscillations is developed at some length in reference 1. Other details on practical construction are contained in reference 2.

If the dynistor is used in the constant-current mode, normal amplifier service is possible. Fig. 4 shows a dynistor microphone preamplifier circuit. Since the UY224 dynistor has a Y_{fs} of 1000 μmho, the voltage gain of this circuit is 20. The high value anode load resistor allows efficient capacitive coupling of the output to the input of another dynistor amplifier stage.

While I have only shown two types of circuits here to illustrate the uses of the two operating modes, you will undoubtedly envision many more. It is hoped that a com-

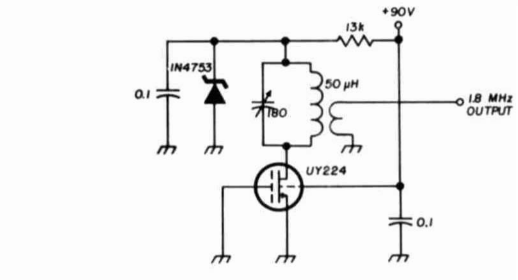


fig. 3. Simple 1.8-MHz oscillator using the dynistor. In this circuit, tho dynistor operates in the negative-resistance region.

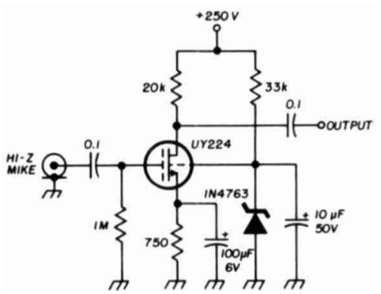


fig. 4. A microphonm preamplifier using the dynistor in the constant current mode.

plete line of dynistors will eventually be developed so we have improved types for higher frequency and power.

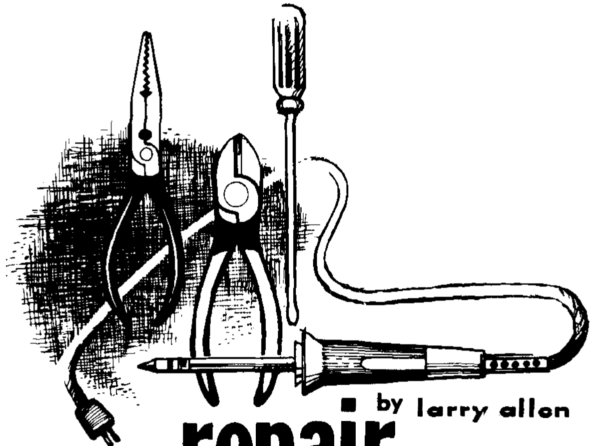
references

1. W. Edson, "Vacuum Tube Oscillators," John Wiley & Sons, 1953, pp. 32-41.
2. A. Chirardi, "Radio Physics Course," Radio Technical Publishing Co., 1931, pp. 904-906.
3. RCA Manual RC-15, Radio Corporation of America, 1947, p 163.

ham radio

As a final note on the dynistor, it should be said that the device does have some mechanically weak points, as shown in the photograph on page 73. These weak points could be corrected, possibly by ceramic encapsulation.

Editor



■ by larry allen

the repair bench

signal tracing in ham receivers

There's probably nothing so aggravating as having your receiver go suddenly quiet in the middle of a long QSO, especially when it's the last one you need to complete your WAS or WAC. Makes you want to drop a hammer into the whole thing.

Instead, drop a probe into it—a test probe. If you've got a signal tracer on the other end of that probe, you can probably find the trouble pretty fast and maybe get back on before the band closes.

With so much ready-made and kit-form commercial equipment, the old familiarity with your home-brew stuff is largely lacking. So what you need is a way to run down faults quickly without being the guy who designed the bloody rig in the first place. Signal-tracing may be it. Once you know how to use signal-tracing, you'll probably agree it's one of the fastest and most convenient ways to track down a trouble.

simple instruments

You can signal-trace with something as uncomplicated as a plain audio amplifier. In a pinch, you could even use a channel of the stereo. One kit-form signal tracer costs less than $25. Inside it is a quiet high-gain audio amplifier, and the probe can be switched to insert a detector diode for trouble-hunting in rf and i-f stages.

One I've used for years—and I see it's still available—is shown in **fig. 1.** This pencil-type is about the handiest thing you can have around for those sudden breakdowns. Its storage box fits easily in a desk drawer. Inside is a transistor audio amplifier, run by a penlight cell (also inside). The clip is the power switch, and you listen through an earphone. Although you can use yourself as the ground, I generally use a jumper lead from the pocket clip to chassis. (Comes from an old habit of never putting two hands in a set at once.) The **Stethotracer*** has a thimbleful of little probes that screw on the tip; for listening in rf and i-f stages, one is a demodulator.

If you decide to use an audio amplifier you already have, it should be very quiet and hum-free. You might want to add a few extra microfarads (40 or so) of electrolytic capacitor across each power-supply filter. A dc-operated transistor amp would be even better, if the noise level is low enough. For rf and i-f tracing, you can build a little demodulator probe like fig. *2.* Or, you can buy the kind that's used with an oscilloscope; a kit model costs under $5 and has its housing and 3 feet of cable.

a tracer at work

Whatever kind of tracer you decide to use, you want to get the most out of it— and **speed** is what it can offer you the most of. Its versatility is something, too. Many who already use signal tracers think they are limited to localizing trouble to one section of a receiver. That's wrong. A signal tracer can also pin down defective circuits and parts in a receiver—and in some parts of a transmitter, too. All you have to do is know how to use it.

Any signal tracer needs one accessory: your brain. Fast troubleshooting with a tracer demands logic, and you have to supply that. I'm going to show you some trouble-hunting in a fairly elaborate double-superhet, to give

* Don Bosco Stethotracer. $34.95 from Allied Radio, 100 N. Western Ave., Chicago, Illinois 60680.

you some idea of the tests you can make with a tracer. However, pay close attention to the **method,** the logic by which the trouble is first localized and then pinpointed. That logic is what'll get you back on the air in a hurry.

Start by looking at the schematic diagram of your receiver or transmitter. Mentally break it up into blocks representing each function or stage in the set. **Fig.** 3 is the functional block diagram of the receiver I'll use as an example. As you see, it's a pretty good ham receiver: double conversion; filters for ssb, cw, or a-m; product detector for RTTY and ssb; and a dial calibrator. For the initial step, group the blocks into four sections; they're marked in **fig.** 3.

First, the rf section. In it, you have to use a demodulator probe with your tracer. What you hear is a mishmash of signals, because the rf stage in this receiver is a broadband one. Once you set the bandswitch, all the stations within shouting distance in that band will be heard if the stages are normal. The plates of the rf amp and the first mixer are the test points for this one. If you get no signal at the mixer, already you know something in the rf section is dead.

The "high" i-f section processes the output of the first mixer. It consists of a bandwidth filter, the second mixer, and the tuning oscillator. If any one of them is at fault, the i-f signal your tracer should pick up at the output (plate) of the second mixer will be missing or fouled up in some way. The normal signal at the bandwidth filter output terminal is a mishmash just as you heard at the output of the rf amplifier. All the stations that are on the air in that hand will be audible through the demodulator probe. In the second mixer, however, is where an individual signal first gets picked out from among all the others. The linear master oscillator heterodynes with one frequency in the passband of the bandwidth filter and creates the "low" i-f — in this receiver, 3.395 MHz.

In the "low" i-f section, you still need the demodulator probe with your signal tracer following the selectivity filter, and for the low i-f amplifiers. The quickest test point for the whole section, though, is after either detector. You should hear a clean, clear audio signal there, without the demod probe.

Finally, the fourth section — the audio stages. The tracer can pick up whatever modulation has made its way this far through the receiver. If the whole set is okay, including the audio amp, you can hear a nice strong signal at the plate of that last stage.

Now, with this broad division of the set in mind, plan your trouble-hunting attack. Remember the secret word: **logic.**

divide and conquer

The first place to touch your tracer probe is at some point about half-way through the set. Use the demodulator probe and make your check at the output of the second mixer. By tuning the master oscillator around the hand, you should be able to find a good station to zero-in on. (If you were cut off in the middle of a QSO, as in my opening example, use the net you've been on.) Remember what an ssb signal will sound like since the demod probe is an am detector. There's a chance, too, that you can tune WWV at 15 MHz, or at one of its other frequencies if you are dealing with a general-coverage receiver.

The output of the second mixer is a good starting test point for two reasons. First, it is the earliest point you can get a single-station

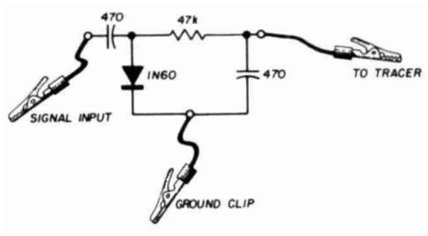
Kit-model signal tracer is high-gain (and quiet) audio amplifier. Probe can be used direct or with switch set for demodulation.

signal, which is easier to evaluate than the many-station mishmash earlier in the set. Second, it divides the set roughly in half, at least by function. If the signal is okay there, you've cleared the whole front half of suspicion. If it isn't, the last half is probably okay.

Suppose you get nothing there. Divide the front part of the set in half, and use the tracer again. You'll still need the demodulator probe, and the output of the first mixer is the place to connect it. Remember that there was no signal at the output of the second mixer. If the signal is okay at the new test point — the mishmash already described — the trouble must lie between the two test points; the second mixer, the master oscillator, or the bandwidth filter has trouble. If there is none, the rf amplifier, the first mixer, or the crystal oscillator must be at fault.

The back half of the receiver lends itself to similar logic. If the signal was okay at the second mixer, the next dividing point is at the am or the product detector output. Omit the demodulator probe, and check directly. A signal in the tracer means everything is okay up to there and the trouble is in the audio section. No signal means it has been blocked between the second mixer and the detector; the filter, one of the i-f amps, or one of the detectors is the offender.

fig. 2. Build your own demodulator probe from this circuit and you can use a plain audio amp.



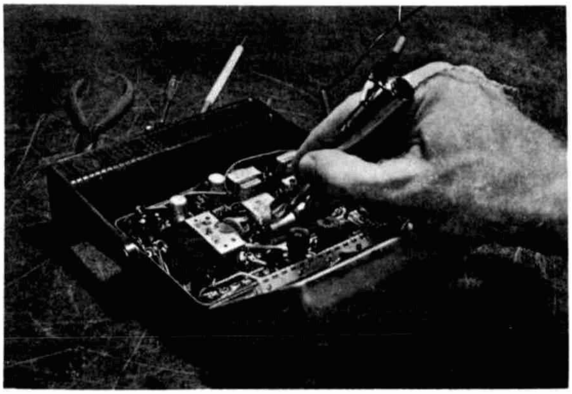
getting closer

You can follow the same logic for each stage. If the signal is okay at the input of a stage and not at the output, it's obvious the trouble is between those two points. The tracer can thus put you right into the stage that's at fault.

This divide-and-conquer technique of stage isolation works just as well for other symptoms as it does for dead sets. You can hunt noise or hum, tracking down the stage where the trouble first shows up. It also works for distortion.

There is one symptom that is probably best signal-traced in a "straight through" method: weak signal. If reception is weak, the fastest way to find out which amplifier isn't doing its job is to check the gain of each one by touching the probe tracer to its input and its output: if there is no increase, the amplifier is weak. Mixers, however, seldom show gain; there may even be a small loss. The filters introduce some loss, too, but you can judge if it's too much, after you get practice.

fig. 1. Servicing a small two-band portable radio with a pencil-type signal tracer.



There are other little tricks of logic that make trouble-hunting easier at this point. As an example: you get am signals okay but can't clear up ssb. The trouble is likely in the BFO or the product detector; they are the only stages that aren't common to am too. Another example: weak signals sound okay, but strong ones distort. A good suspect is the agc stage, which may not be controlling the rf and i-f gain as it should — letting strong signals overload the receiver. This can be serious with ssb, which is especially sensitive to overloading.

Once the faulty stage has been pinned down, the usual procedure is to start measuring dc operating voltages on the tubes. There's nothing wrong with that technique, except that it's a little premature. Your signal tracer can still tell you things you can't find out with a voltmeter. In some cases, you may have to revert to the tracer or even more elaborate equipment after the dc measurements. So, for speed, stick with the tracer a while longer.

And do what with it? Here are some specific things you can check inside a stage, using only your signal tracer. Be sure to use the demodulator probe when rf or i-f signals are involved, and tune in a signal of some sort whenever the front half of the set is working.

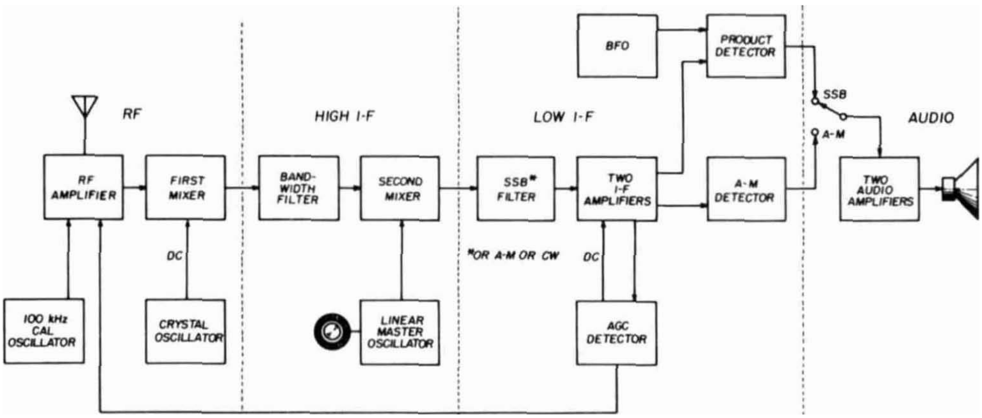
Fig. 4 shows a make-believe stage—a sort of composite to illustrate some of the parts that can be checked with a signal tracer. The different components that are highlighted can be tested right in the circuit, usually without any unsoldering. They are coupling and bypass components, which often can't be checked by voltage measurements.

Coupling capacitor C1 and interstage transformer T1 have one thing in common: they should pass the signal along with very little attenuation. Whether they are large, as in audio stages, or small, between rf or i-f amplifiers, there should be about the same amount of signal on both sides. If there is any attenuation, it should be small. To check, touch the probe to the input side of the part, and listen; then to the output side, and listen again. If the output is much weaker than the input, the part is defective.

The bypass capacitors, C2 and C3, are there to shunt the signal to ground. Their values are chosen to short out almost all signal at the cathode (C2) and at the screen (C3). The tracer should hear very little signal at either point. If there's much, the capacitor is not doing its job. You might find a little signal at the screen of any pentode amplifier, but if it isn't a lot weaker than the input signal (at the grid) something is wrong.

Finally, notice that the B-plus connection is highlighted in fig. 4. You can check for the source of hum with your signal tracer. Just connect it directly to the various B-plus points in the receiver. Power-supply filters are just like any other bypass capacitors; they should shunt all the signal voltages to ground and leave only dc. If one of them is at all weak, you will hear an undue amount of hum in the tracer; you may even hear something (usually a whistling or a hissing sound), that would be rf or i-f signal if you could unscramble it.

fig. 3. Sectionalizing a ham-band receiver by functions. Dividing it up this way makes it easy to track down trouble by a logical method.



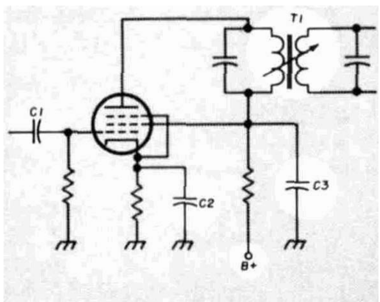


fig. 4. You can test these components with your signal tracer, without unsoldering them.

a few other tricks

Now you know how to pin down a great many faults quickly. If you haven't found the trouble by this time, you're going to miss the rest of the QSO anyway, so don't feel too bad about resorting to the dc voltmeter—that's your next step. You've gone about as far with signal tracing as you can. But you've done it with only a few minutes' work. The whole procedure just outlined takes less than 5 minutes.

There are a few other tests you can make with your tracer, and I'll throw them in for good measure. For one: the accepted way to find out for sure if an oscillator is running is by measuring its dc grid voltage. It that's missing, the oscillator is dead. You can also tell with your tracer, even though there is no modulation to be heard. Look back at **fig.** 3; you can check all those oscillators.

If other tests make you suspect the first oscillator is dead, pull the rf amplifier tube. Touch your probe (without demodulator) to the output of the am detector and listen to the hiss. Pull the first oscillator out of its socket. The hiss will diminish if that oscillator was working.

If you suspect the master oscillator, use a similar technique. Pull the first mixer (to eliminate noises from the front end). Touch the tracer to the output of the am detector. Pull the master oscillator tube while you listen to the circuit noise through the tracer. If the oscillator was working, the slight noise will stop.

The same test will work for the BFO, which is just another oscillator. Pull out the second mixer (again, to kill front-end noise). Listen with a tracer at a convenient point in the audio amps, one that lets you hear the noise in the set (not within the tracer). Switch the receiver from am to ssb or cw, which activates the BFO. You should hear more hiss in the tracer, if the beat oscillator is working. (You may be able to hear this in the speaker, without the extra amplification of the tracer.)

You don't need a trick like this to check the calibration oscillator. Make sure the rest of the receiver is working, and you can hear (or not hear) the calibration whistle readily.

next month

There you have some repair tricks that will save you time whenever your receiver kicks out. You can use the same techniques in some stages of a transmitter, but be careful —your little tracer may not be able to take those power-laden transmitter stages. Confine it mainly to speech amplifiers and low-level exciter circuits.

Though signal tracing, as you can see, is a fast way to hunt down trouble in almost any receiver, there is another method that is similar and almost as useful. Next month, in the **repair bench** I'll show you how you can do many of these same tests with this other technique—called **signal injection.** Then you can choose whichever one you prefer when you hear those signals dropping down into the silence that is so infuriating just when a new joke is coming over the net out of K6-land.

ham radio

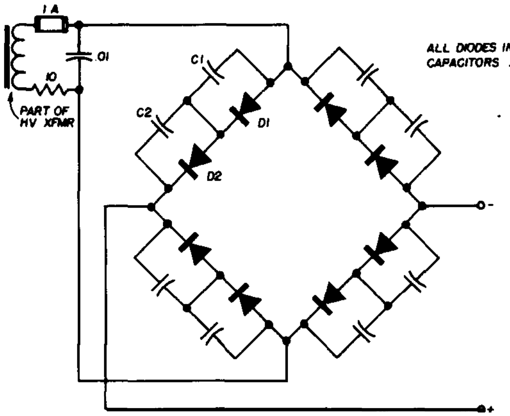
letters

Dear Larry:

I had something interesting happen in my Collins 30S-1 linear and thought you might like to hear about it. In the middle of a sked with a K7 friend of mine, my big jewel shut down. Snooping around, I found the screen-supply fuse blown. Put in a new one, it blew again. I got out the ohmmeter and measured all the diodes. Two of them were shorted— Dl and D2 (see diagram). If one goes, the other is likely to; I replaced them both.

When I fired up, pop went the fuse. Sure enough, the same 1N1492's were shorted again. Another trip downtown, and two more diodes. This time I checked all the others again with the meter. They were all okay, but the D1 I had just put in was already leaky. I took it out, and it checked okay. You guessed it—the leakage was in the capacitor, C1. It was burning out D2, which then burned out D1. Except for my happy accident, measuring the new diodes in the circuit, no telling how many of those 1N1492's I'd have used up. Like popcorn.

Anonymous Friend
Garden City, N. Y.



Okay, thanks, pal. You pinned it down pretty fast. Happy, maybe, but no accident. Sounds to me like you knew the score pretty well; you were in the right circuit, checking the right thing, with the right instrument. That's no accident.

Dear Larry:

My Heathkit HR-10B *receiver keeps cutting out.* It *goes off completely, but the tubes stay lit. It does it whenever the table is jarred a little, and hitting the top makes it snap back on, sometimes. I'm not a ham yet, and don't have any idea how to fix this. I use the* HR-10B *for code practice and I do a lot of* SWLing. I *would appreciate* if *you can help.*

Jeff **Woodroot**
Edison, New jersey

Your best bet, Jeff, is to find a friendly ham in your neighborhood and see if he'll help. Of the several things that might cause the trouble, the only one you can probably do much about is this: there's an octal plug with a black cover that goes in a socket on the rear apron of the chassis. If you built the receiver, you remember that it has a jumper between pins 1 and 6. Make sure that jumper is soldered well in both pins of the plug. Also, make sure the plug isn't loose in the socket. If it is, you can tighten the socket pins with an open safety pin. Run the point down alongside each contact, between the metal and the side of the hole. Do this two or three times to bend the metal contact a little so it grips the pins of the plug better.

Dear Larry:

Maybe you can tell me what to do. I should have had it looked at long ago, but I didn't. I used a Hammarlund SPC-10 (receiver). About a year ago I noticed it getting weaker. At first it was just the sound that seemed weaker, but now I'm sure I'm not getting anybody as well as 1 should. A guy I know across town works 10 meters hot as a firecracker some weekends, yet I can hardly raise a signal out of the noise. I've already retubed it from one end to the other, but that only helped a little. The noise is louder, hi! Got any ideas?

David G. *Montgomery*
Joliet, *Illinois*

Sounds to me, from the few clues you give, like your SPC-10 could use an alignment job, David. I assume you've checked all the voltages in the rf and i-f stages. If you don't have the equipment or know-how to do an alignment, best take it to a qualified service shop. Your first trouble was probably a weak audio tube, which you subsequently replaced in retubing the set. But the long-term alignment drift is apparently still with you. (Incidentally, in a couple of months I'll be writing about alignment in this column, if you can wait that long.)

The repair bench is for you. Tell us about problems you have run into, keeping your rig in peak shape. Questions you ask will be answered only if *accompanied by a copy (not returned) of the full schematic diagram and a stamped, self-addressed* #10 *envelope. We will print some of the most interesting case histories.*

Editor

ham radio

replace one unijunction transistor with two transistors

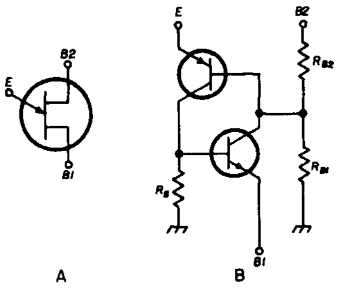
Davisson, K9VXL, RR #3, Columbus, Indiana 47201.

Remember that circuit you recently looked at and wanted to try? You passed on, though, right? The reason was quite simple. It used a unijunction transistor (UJT). Many hams have built up a collection of miscellaneous transistors in the junk box; the right combination to make your own UJT may be there waiting.

the unijunction equivalent

The symbol for the unijunction is shown in **fig. 1A.** The two-transistor version of the UJT is shown in **fig. 1B.** The leads are labeled to correspond with **fig. 1A.** When the NPN and PNP transistors are connected as shown,

fig. 1. The schematic symbol for a UJT is shown in A. The equivalent two-transistor circuit is in B.



they effectively produce the equivalent internal construction of the UJT.

There is only one minor difference—it is necessary to add two resistors externally to the circuit to produce the equivalent resistance found between each base lead and the emitter of the UJT. These two resistors are labeled R_{B1} and R_{B2} in the diagram. Resistor R_S is added for stability; a value around 10k should be sufficient.

As a general rule, the values of the two resistors R_{B1} and R_{B2} may be determined by knowing only two characteristics about the UJT you're replacing; the **intrinsic stand-off ratio** and the **interbase** resistance. These two characteristics are related by the formula:

$$\eta = \frac{R_{B1}}{R_{B1} + R_{B2}}$$

where η = intrinsic stand-off ratio

$R_{B1} + R_{B2}$ = equivalent interbase resistance

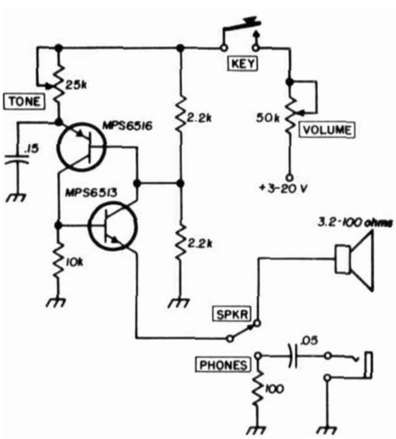
In unijunction transistors, the interbase resistance is typically between 5k and 10k ohms. The intrinsic standoff ratio of a common general-purpose UJT runs about 0.6. Using these values in the formula, you can quickly calculate that R_{B1} would have a value around 3k and R_{B2}, about 2k, to produce the equivalent of a general-purpose UJT.

In most applications, the actual values aren't critical and equal values of resistance could be used—such as 2.2k. This would produce an intrinsic stand-off ratio of 0.5 which is fine. If lower values of resistance are used, the circuit will draw more current and vice versa.

unijunction code-practice oscillator

Now for an actual application. A code-practice oscillator using the two-transistor UJT equivalent is shown in fig. 2. Both tone and volume controls have been provided as well as a choice between speaker or earphones.

fig. 2. Here is a **relatively simple** CPO using the two-transistor equivalent UJT.



In this circuit, a 3-volt supply voltage was sufficient, but you can use as high as 20 volts. The actual voltage will depend upon your transistors and the output level you desire.

In general, a little experimenting with two transistors (one NPN and one PNP) will let you duplicate the function performed by the UJT—generally at lower cost. The transistors shown in fig. 2 are Motorola devices which are inexpensive and usable from audio through six meters: the latest price list shows the MPS 6513 at 57/ and the MPS 6516 at 60C. The price of these two transistors is in the ballpark of general-purpose UJT's, but with this approach, you may already have the UJT in your junk box.

Next time you start to bypass an article using a unijunction transistor, stop and think about how you could substitute two transistors and still have that same useful circuit.

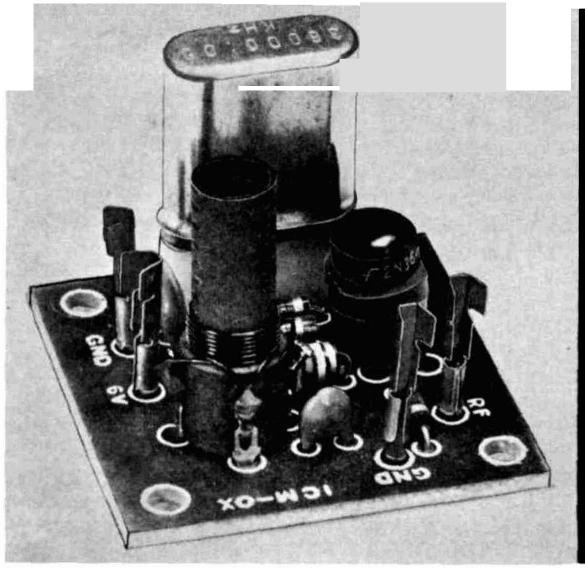
reference

General Electric Transistor Manual, Seventh Edition, 1964, page 300.

ham radio

Interested in saving money?

Turn to page 47



a look at the EX crystal and its oscillator

A review of the new experimenter's crystals and oscillators from International Crystal

Doc N son WB2EGZ, 9 Green Ri Road, shland, New Jersey 08034

A new line of low-cost quartz crystals and crystal oscillators has been offered to amateurs by the International Crystal Manufacturing Company.[*] The crystals, called EX, have a guaranteed tolerance of .02% and are made for operation in non-critical oscillator circuits such as the International Crystal OX oscillator line. In fact, these crystals are only guaranteed to oscillate in that circuit.

Since I'm a gambler at heart, I purchased a 36-MHz unit at $3.75 for use in the nuvistor oscillator circuit of the ARRL Handbook six-meter converter. The excellent results I obtained prompted further investigation of a second EX crystal and the OX oscillator kit.

crystal characteristics

Let's talk about crystals in general to set the stage for my findings in this investigation. The precise frequency of operation of a crystal will be modified by its loading in the oscillator circuit and also by the drive level of the crystal in that circuit. Therefore, the crystal should not be expected to have the same exact frequency from one circuit to another.

Next to frequency, the most important characteristic of a crystal is its equivalent series resistance (ESR) which is related to a crystal's activity. The ESR of a crystal is that

*** International Crystal Manufacturing Company, Inc., 10 North Lee, Oklahoma City, Oklahoma 73102.**

value of resistance which may be substituted for the crystal in an oscillator circuit and maintain the same frequency. Substitution of resistors is basically how the measurement is made; however, I would recommend the use of laboratory equipment over a ham-shack "kluge." In classic theory, the Q of a filter is used to describe its activity. In practice, ESR is more easily measured; and, because of its Inverse relationship to Q, may be used in its place.

Using a standard crystal impedance meter in conjunction with an rf voltmeter and frequency counter to measure two 36-MHz crystals, the following results were obtained:

Frequency (MHz)	Frequency Accuracy	ESR (Ohms)
36.000582	+0.00106%	62
35.999129	−0.00240%	20

Both crystals have much better accuracy than .02%. In the OX oscillator, the crystals are pulled to a slightly higher frequency, but their accuracy was still better than .005%. The equivalent series resistance readings show the first unit to have lower activity than the second. A normal crystal of this frequency would be expected to have a maximum resistance of 40 ohms; therefore, the first crystal may have trouble oscillating in critical circuits.

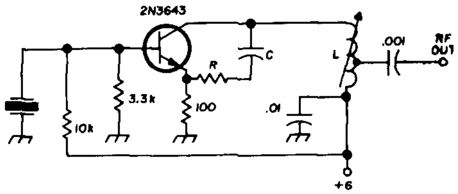
What determines a critical circuit? Usually, the power level at which it drives the crystal. Bipolar transistor oscillators will generally be more demanding than FET or tube oscillators. The Pierce oscillator tends to be more critical than the OX circuit.

I was unable to find what (if any) is the limit to the ESR of an EX crystal. Upon talking to a representative of International Crystal, I was assured that the crystal blanks which are used are the same as those of the higher-priced crystals and most of the manufacturing processes are the same. Stabilization processes are omitted, which means the user may see a greater change in frequency and activity in time than he will when using a more expensive crystal. I should hasten to point out, however, that such changes will not be radical if all crystals are as good as those I tested.

The OX oscillator (fig. 1) is sold as a kit for $2.35 which is less than the cost of the individual components. Although the oscillator is designed to work from a 6-volt source, tests were made with supply voltages ranging from 3.5 volts, where oscillation failed, to 8 volts. I see no reason why higher voltages would not be practical, but the data tabulated was made to show typical performance over a range of unregulated supplies which might be available.

Supply Voltage (Volts)	Frequency Deviation (Hz)	RF Power Output (mW to a 50-ohm load)
4.0	−144	0.29
5.0	−058	0.92
6.0	000	1.8
7.0	+051	2.9
8.0	+102	3.7

fig. 1. The basic International Crystal OX oscillator circuit.



summary

The EX crystal is seen to be a good bet for most amateur work where high accuracy and long term stability are not required. I would not be scared by the conservative advertising, because these are good quality crystals. The units tested will operate in many circuits other than the OX oscillator.

Although the oscillator will operate over a wide voltage range with excellent frequency stability, the user will experience power-output variations which are quite large. This is typical of many transistor oscillators, but is emphasized here as the primary effect of supply-voltage change. In all, the EX crystals and the OX oscillators appear to be a real bargain for the experimenter, and not another low-cost disappointment.

ham radio

Quement Electronics

circular electronics slide rule

Here's an electronic slide rule that's suitable for amateurs, experimenters and engineers

James Ashe, W2DXH, 305 Oak Avenue, Ithaca, New York 14850

Quement Electronics* has been very active in the ham supplier business. I've noticed their large ads several times, so during a recent visit to California I naturally dropped in to say hello. Had a nice talk with Pete Phelps, W6ERP, about the ham market. "Pete," I said at last, "What's new in small things that hams might find useful?" He though for a moment, and looked up. "Jim, we've got these new electronics slide rules, just came in. I think the hams are going to like them. Interested?" I certainly was, and after a careful inspection I bought one.

The slide rule is a resilient plastic disc about 4-3/8 inches in diameter. There is a smaller rotating disc on each side, and a rotating arm that carries the hairline. It's a circular slide rule, with the circular rule's special advantages. The 14 scales are cut into the plastic, and are very sharp and clear.

I sat down with a cup of good coffee to work out the details. Fascinating. Some of the notations seem odd, but it's far better than the archaic multiple zeroes appearing on some rules. The manufacturer uses a comma to indicate a X1000 factor in some places —one or two references to table 1 should clear things up.

The scales work from low audio to VHF and a little beyond. Two dB scales on the front provide for calculations based on power,

* **Quement Electronics, 1000 South Bascom, San Jose, California 95128**

tabla 1.

scale symbol	quantity	exponential
P	picofarad picohenry	10^{-12}
, P	nanofarad nanohenry (thousands of picos)	10^{-9}
u	microfarad microhenry	10^{-6}
m	millihenry (thousands of microfarads, uncommon)	10^{-3}
K	Thousands (kilohms, kilocycles)	10^{3}
M	Millions (megohms, megacycles)	10^{6}
, M	Thousand-millions (gigacycles)	10^{9}

voltage or current measurements. A surge-impedance scale may interest engineers more than hams, but then there's not much difference these days, is there? Another scale, in red, indicates inductive reactance, and the basic L, C and F scales extend around the outside edge. Each scale carries a labeled arrow to indicate the parameter and to remind you of its direction.

On the back, a small scale relates frequency and wavelength from 30 Hz to 3,000 MHz. Note the comma used here, too, as a X1000 multiplier. The capacitive reactance scale is done in red, and the resonant frequency scale in black.

Here are a couple of hints. To become familiar with the Electronics Slide Rule, start with familiar L, C and F values and work gradually into new territory. The rule looks complex at first, but if you're accustomed to working out electronic calculations, it'll become familiar in minutes. If you're a beginner, choose one scale and try a few paper calculations to get oriented.

This is a good investment for any level of work. The rule ought to have a useful life of many years; it's waterproof, and seems to require no special care. It comes with a good plastic case, and a short-form instruction manual. Oh yes, I've thrown away my old paper rule—this one's much better.

ham radio

the secrets of long-tail biasing in transistor circuits

Here's a method for biasing your transistor circuits that offers simplicity with reliability– it offers reduced problems with drift, transistor replacement and silicon-germanium interchangeability

es A. Ai e, W2DXH, 305 Oak Ave e, Ithaca, N York 1 50

At least one-half of the work of designing a new circuit goes into developing appropriate biasing for its active elements. In modern gear, these are almost always transistors, usually with a simple base-bias arrangement. Base biasing is hard to compute, may require selected transistors, and is accompanied by a thermal-runaway hazard. There are other ways to achieve correct bias, and one of them is particularly well suited to amateur design and construction.

This is "long-tail" bias, an arrangement which deprives the transistor of control over its operating current. This reduces unwanted effects of transistor drift and of replacements whose characteristics are different from those of the original transistor. In fact, a long-tail bias circuit can be designed to accept silicon or germanium transistors interchangeably, with both being properly biased. If you use unlabeled surplus and computer transistors, you should find this biasing method particularly interesting.

I'll start by discussing why biasing is an important subject. Then a few words about the competition: factors that upset biasing. I'll finish with some illustrative applications and notes on finding supply-voltage sources.

dual function of biasing components

Before we discuss circuits, let's think about coins. Coins and circuits have something in common. Coins have two faces, and a coin seen heads may appear very different from the same coin seen tails. Yet we recognize coins easily, because we are familiar with them.

Circuits have a double character too, and it becomes very important when we have learned to understand it. Circuit design problems are simplified if we can limit the number of details under consideration. The two pictures of a circuit that we should learn to recognize are the signal view and the bias view.

Of course the signal view is the most interesting: that is why we are working on the circuit. Seen from this perspective, resistors and other components often appear as items that reduce the performance we might obtain without them. This is not a very flattering picture of the components we find in real circuits.

But if we check a real circuit with no applied signal, we find there is something going on anyway. Instruments show constant voltages and currents. Resistors and other components get warm. Energy is consumed, although there is no output. Here is another side to the circuit's life, a quiet and unspectacular heartbeat we may not have noticed before.

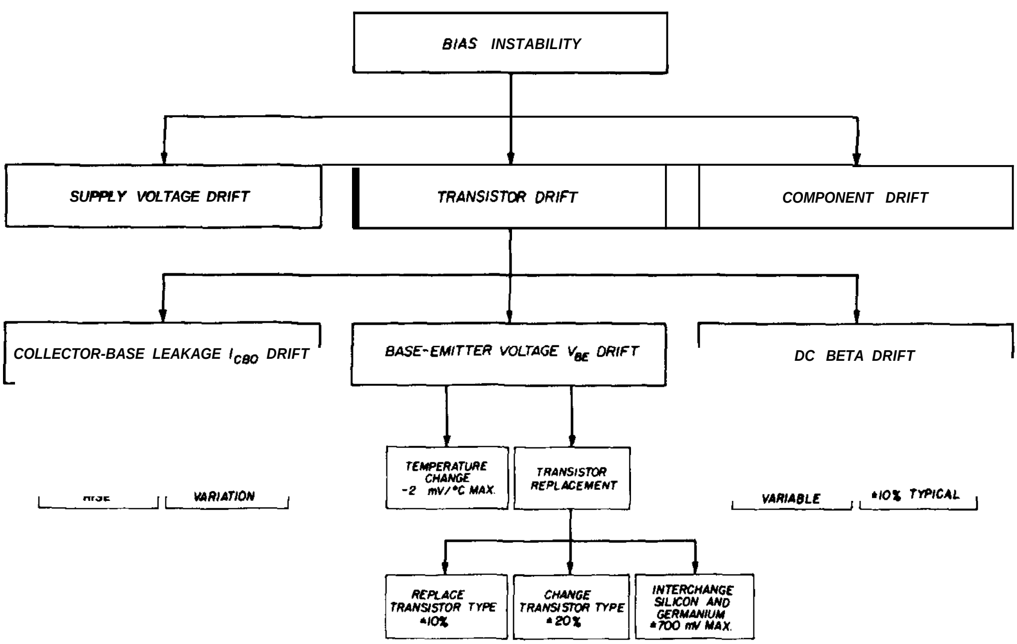
The values of these steady voltages and currents are referred to as "standing" or "static" values, to avoid any confusion with signal voltages and currents. In particular, the collector voltage and current together identify the quiescent point, or Q point, chosen as the clearest indication of bias conditions.

We bias a transistor for two purposes. The transistor is, basically, a piece of crystal. It cannot manufacture power. It can only borrow power, and the output signal is really borrowed from the bias voltage and current. This is why supply lines must be adequately bypassed: so that power borrowed by the transistor is properly returned to the supply circuit.

The other purpose in biasing a transistor is to determine its signal characteristics. All active devices show different input, output, and gain characteristics at different Q points. We might choose an almost-off or a hard-on Q point for pulse work, or some in-between Q point for linear signal amplification.

To a large degree, the Q-point design and stability problem can be separated from the signal design problem. I have summed up the situation in **fig. 1**. The overall problem of designing a circuit breaks down into two simpler, somewhat independent, problems of signal design and bias design. The bias design problem, in turn, breaks down again to ques-

fig. 1. The different parts of the circuit-design problem.



tions of choosing a good Q point, and of designing a circuit that will hold the Q point against upsetting influences. After working out the details, we put them back together again, and we have our completed design.

factors that upset biasing

Suppose you have assembled a circuit and a test shows correct standing voltages and currents. Some time later, maybe minutes later and possibly years later, you make another test and discover that although the circuit looks the same, the Q point has shifted. This is a perfectly natural event, and it will always happen in real circuits. We call it drift. There are five major causes of drift in transistor circuits, and part of the design problem is to take them into account so that drift does not seriously disturb circuit operation.

Here are the disturbing influences:

1. supply-voltage variation
2. component value drift
3. collector-base leakage, I_{CBO}
4. base-emitter voltage variation
5. beta variation

These drift components are not equally important, and their effects show differing emphasis in various circuits. I've arranged them in a convenient chart in fig. 2.

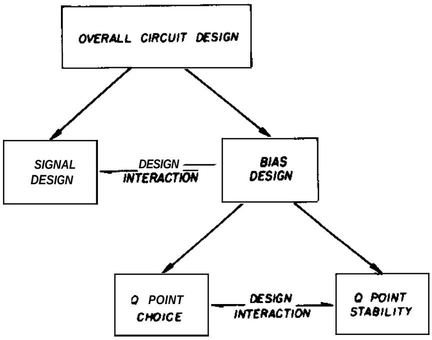
Because the biasing network is connected to the supply voltage source and may be seen by the transistor as a part of it, when the supply voltage drifts, the Q point moves. Very simple supplies consisting of a transformer through a diode rectifier to a capacitor filter, with no regulation, are subject to wild variations in output as load changes, or over a period of time. Plus or minus 30% seems to be in line with real experience. A simple regulating circuit or a zener diode can reduce this to the 1% ballpark and is practically always a good design provision.

Supply stability seems guaranteed when using batteries, but this is subject to hazards. As batteries age, their internal resistance rises, and this may upset a circuit much more than a mere falloff in voltage. Also, old batteries, whether sealed or not, often emit chemicals with a deadly corrosive effect upon electronic components.

As we proceed along the connections from the power source to the active devices, we come upon many resistors. A resistor is a fine, solid, stable-appearing device, but in actual experience only expensive resistors are truly reliable. The common composition resistors show noticeable drift upon exposure to heat. Soldering is likely to produce a permanent change in value. Resistors should be soldered gently, like transistors, and salvaged resistors are not to be trusted. Good design practice is to double resistor tolerances: a 10% resistor is reliable to 20%.

The other three sources of bias instability originate inside the transistor's shell. They are

fig. 2. The major factors causing Q-point drift



results of temperature variations, transistor aging, and transistor replacement. Amateur and experimenter circuits do not normally run very hot, but may at times be exposed to extreme temperatures. Aging is only rarely a noticeable factor. The most abrupt changes occur when a transistor is replaced. Good biasing design will result in a circuit with good tolerance for all anticipated drifts.

The most frustrating factor in base-bias designs is,,I A transistor is basically a pair of diodes, and their close physical proximity does not upset their basic diode nature. The normal diode reverse leakage appears in the base-collector diode as,,I which has considerable nuisance value. Typically, this leakage is in the low nanoampere* range in silicon transistors and the low microampere range in germanium transistors. It may be much larger in a poor signal transistor or a

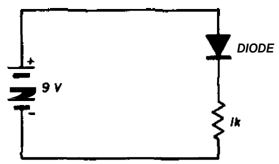
* 1 nanoampere = 0.001 microampere = 10^{-9} ampere.

good power transistor.

I_{CBO} is annoying because its value may range over a 10:1 spectrum for different specimens of the same type transistor. In addition, it shows a very strong increase for moderate temperature rises. It roughly doubles every time the temperature goes up 10 degrees C. A germanium transistor with four microamperes leakage at 25° C (room temperature) will show about 20 microamperes leakage at 45° C, and the next transistor of the same type could have 60 microamps leakage. Small silicon transistors do not have annoying leakage except at very high temperatures.

The second transistor instability problem is rarely noticeable, requiring attention only for some precision circuits and for situations

fig. 3. A basic diode circuit illustrating the principle of long-tail biasing.



where there may be extreme temperature variations. There is normally a small forward voltage from the base to emitter of a transistor biased in its linear range. In addition to the expected small variations depending upon base current, the base-emitter voltage shows a negative temperature dependence that is typically under 2 mV per degree C. A ten degree temperature rise will reduce base-emitter voltage by roughly 20 mV in either a silicon or a germanium transistor.

Replacement of a germanium transistor with a silicon transistor has a stronger effect upon base-emitter voltage. A germanium transistor typically shows near 200 mV, a silicon transistor, near 700 mV emitter-base forward bias. This voltage variation appears in the biasing circuit, and must be considered when designing circuitry that is to accept silicon and germanium transistors interchangeably.

Finally, there is the beta problem. Transistor beta is the factor by which collector current is greater than base current. The transistor has one beta for dc biasing, and a typically lower one for signal computations. Both betas are highly variable, having been brought partly under the manufacturer's control only recently.

Beta variations of 10 or 15:1 are commonly seen between transistors of the same type, and the 2:1 spread of GE's 2N3394 is a recent achievement. Transistor beta also depends weakly upon temperature and somewhat upon collector current. If the transistor is starved for current, its beta falls off.

Temperature drift effects are particularly noticeable in outdoor battery-operated gear. Only I drifts favorably when temperature falls; everything else goes the wrong way. Battery voltage and capacity are decreased and internal resistance increase; transistor base-emitter voltage rises, and beta is reduced. Electrolytic capacitors and other components may be affected too. These combined responses are the reason many amateur-designed circuits fail completely at or slightly below the freezing mark.

Now we see what the competition is. It is impressive. I have broken the overall picture into its component parts in **fig.** 2, which will help you keep the different factors organized. What circuit will maintain a Q point reliably against these influences? It must not be so stable it refuses to respond to the signal! There is one simple arrangement that belongs in the amateur literature, but somehow is not included. In a majority of cases we can delegate the entire biasing job to a single, easily-computed resistor.

long-tail bias design

This resistor is the "long tail" in long-tail biasing. Its length appears as its rather large value, and the substantial voltage that appears across it during normal operation.

To see how this works, let's start with the simple diode-resistor-battery circuit of fig. 3. Imagine that you have an unlabeled diode, but the resistor and battery are I k ohms and 9 V respectively. We want to estimate the highest and least currents we should expect to find in the circuit. The only unknown is the diode: whether silicon or germanium. Its exact characteristics are not very important.

If we have a germanium diode, the voltage across it in forward conduction is unlikely to be less than 200 mV. Or if it is a silicon diode, its forward voltage is unlikely to exceed 900 mV. Subtracting the diode forward voltage from the supply voltage, we find at least 8.1 volts but not over 8.8 volts across the 1k resistor. We would measure a current, if meter resistance is insignificant, of 8.1 to 8.8 mA, an 8.7% variation.

This is the key to our long-tail biasing arrangement. Most of the supply voltage appears across the stable resistor. Very little appears across the diode, which is left with little influence upon overall current. Minor changes in diode characteristics are swamped by the powerful influence of the series resistor.

Now suppose we replace the diode with a transistor, as in **fig. 4.** If S1 is open, the base-emitter junction will appear to the circuit as an ordinary diode. We can plug in germanium and silicon transistors, and the emitter current will show the variations previously observed with the real diode.

What happens when we close S1? The emitter-to-base voltage remains about the same (it will increase by maybe 10%), but the emitter current is stolen inside the transistor by the collector diode before it can get to the base terminal. Its actual value is not appreciably affected. Now the transistor is biased to the 8.1-8.8 mA we chose, less a small base control current.

Let's look that over once more. Knowing supply and base-emitter voltage, we know the voltage across the bias resistor. The effects of drift are easily reckoned by estimating the individual drifts and taking their sum. If bias stability is not good enough we must use a larger resistor and a higher voltage supply. That seems to be the complete answer.. .

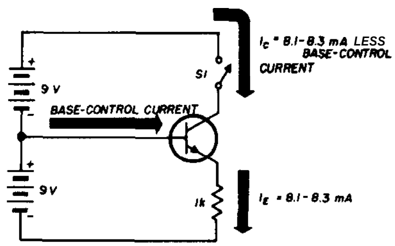
But what current will we choose? A small current is economical and minimizes heating. A large current gives more power gain. In general, the small current wins, if it is not too small. See table 1 for some notes on transistor performance. And there are usually some other factors that will help make a choice.

If we are using a germanium transistor, collector leakage current may be a factor, particularly at high temperatures. The load resistor carries the collector current we intended, and I_{CBO} too. If the leakage current is too great a percentage of normal collector current (maximum maybe 5%), the transistor will be partly paralyzed. I_{CBO} can usually be neglected for silicon transistors.

A second practical limit relates to the size of the transistor's electrical structure. The transistor requires enough current to be well energized. If it is starved, its gain falls off. Modern transistors, particularly some epoxy-packaged ones, have tiny structures and do well at surprisingly low current levels. See the graph which compares the dc betas of a typical CE 2N3394 and an old 2N338.

At the other extreme, we must not overheat the transistor. When transistors overheat, doping atoms inside the crystal structure start jumping around, and they never end up in

fig. 4. A complete transistor bias circuit.



positions as good as those they left. This damage may occur in milliseconds and is permanent. If you don't have the manufacturer's specs, be careful. Also, note manufacturer's derating for high operating temperatures.

Now we can choose a collector load resistor. The collector current will be practically the same whether we use a resistor or not; maybe you will want to use an LC circuit load instead of a resistor. If so, plan to use a small resistor anyway for decoupling. But if the resistor serves as load, choose a value that leaves clearance for collector signal voltage swing. The largest signal should not carry the collector current far into the weak-transistor range or the voltage under a volt or so. A scope check will show distortion or clipping if the signal is too large for the biasing conditions you have chosen.

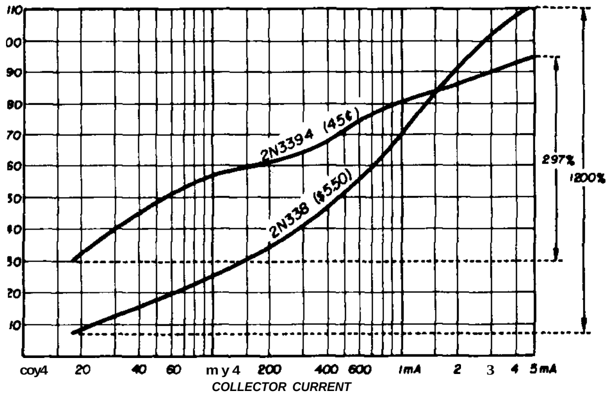
putting the long-tail into a circuit

To make this more real, let's imagine you're looking over my shoulder while I design a simple amplifier stage. The transistor came out of a board; we don't know if it is silicon or germanium: a detail. An ohmmeter check shows it to be an NPN transistor, and in the next two minutes we'll design the circuit. First, we draw a schematic, See **fig. 5.**

Now, our strategy is to estimate or calculate whatever we can, and write it on the schematic. Each voltage, current, or parts value we pin down tells us something about another, and we soon find we have them all.

Starting at R2, whose lower end is at ground, we go to its base end, and let's say the base is at minus 1 volt. Transistor current is determined by R1 and we want to leave lots of supply voltage across it. Proceeding across the base-emitter diode we lose 0.2-0.7 volts depending on what kind of transistor we

Comparison of the operating characteristics of the 2N338 and 3N3394.



have, and that puts the emitter at minus 1.5 volts as a safe average possibility. Each of these values was written on the schematic as we arrived at it.

This unknown transistor should be good for 50 mW collector dissipation. With the base end of the base-collector diode at minus 1 volt and a supply of plus 9 volts, we have about 10 volts for collector swing; half of this reckoned from the base's minus 1 volt puts the collector at plus 4 volts. Assuming five volts available to cause collector dissipation, we wind up with 10 mA as a likely collector current. Write that beside R1 with an arrow indicating direction of conventional current.

Computing R1 from the 7.5 volts and 10 mA figure gives us a resistance we cannot find in the parts box, so we choose 820 ohms.

table 1. Characteristics of GE's 2N3394 silicon NPN epoxy-encapsulated transistor.

V_{CEO}	=	25 V maximum
I_C	=	100 mA maximum
P_C	=	200 mW maximum
I_{CBO}	=	0.1 mA max at 25° C
		10 μA max at 100°C
β	=	58 at −30° C
		100 at room temperature
		130 at +50° C

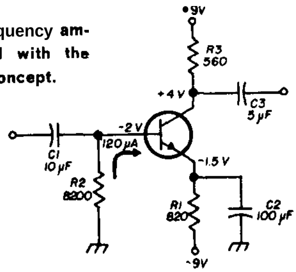
At I_C = 2 mA, common base emitter input resistance ≈ 15 ohms typical

Cross out the 10 mA and write 8.55 mA. Practically the same current flows in the collector circuit, giving 560 ohms for R3. Dividing the collector current by a likely transistor beta value gives an estimated base current of 720 microamps, and knowing base voltage, we find that R2 must be about 8.2 ohms. Only the capacitors remain unknown.

The transistor emitter circuit operates at a tens of ohms impedance level, so we choose 100 μF for C2 as having a comparable reactance at low audio frequencies. C1 is chosen 10 times smaller, since transistor beta is sure to be far over 10 at audio frequencies. This stage will show a low-frequency cutoff determined by C2 alone.

C3 is a little harder to determine, since the load resistance is not given. Source resistance is the transistor's output resistance in paral-

fig. 5. Low-frequency amplifier designed with the long-tail bias concept.



lel with R3; actually, that's another article so just take my word for it. Since C3 should be much smaller than the load resistance at low audio frequencies, a safe value might be 5 μF. A quick breadboard test will tell you if any changes are required, and if the gain is large enough for your purposes.

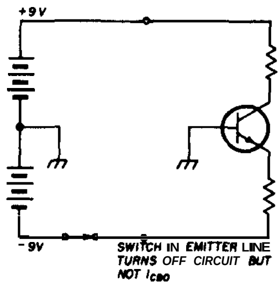
Hints: breadboard rf biasing circuits as audio circuits for scope checks. For small signals, use a large collector resistor for low collector voltage but improved gain.

finding bipolar voltages

There is nothing new about using bipolar voltages (two polarity) in circuit design. It is common practice in good industrial and scientific designs. The advantages are not available in any other way, and perhaps we will discover that bipolar voltages are not so hard to find after all.

The key point is, we don't really need two separate voltages. Rather, the transistor must see two voltages with respect to its base terminal. Since from a signal viewpoint we are free to call any supply terminal the ground terminal, we can find several solutions to the two-voltage problem.

fig. 6. Long-tail biasing from batteries.



It is easy and convenient to call the base terminal chassis ground. If we do this, we will need a two-polarity supply: one positive and one negative with respect to ground. Nine volts is a convenient value, and we achieve the required voltages with a pair of nine-volt batteries as in **fig. 6.**

A second solution involves a pair of zeners across one power supply. See **fig.** 7. The junction between the zeners goes to chassis ground, the transistor base terminal goes to chassis ground, and the transistor sees two voltages although we have only one supply. This approach adds one zener and its shunt capacitor to the collection of components you would have used anyway.

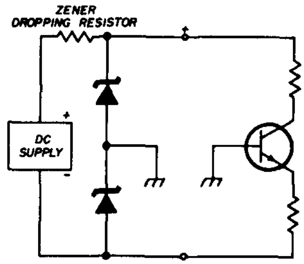
If you are combining long-tail biased circuits with conventionally base-biased circuits, you can use a zener to obtain some intermediate base voltage as in **fig. 8.** The base is at signal ground but zener regulated above bias or supply ground. From the transistor's viewpoint this is the same as having the base attached to the chassis. A true bipolar arrangement is preferable at vhf.

Summing up, there are generally three points you can tie to chassis ground at your convenience. They are supply positive terminal, supply negative terminal, or something in between which is either a tap or a zener terminal. With the assistance of one or two additional zeners you should be able to work out any practical circuit problem.

applications

When you start breadboarding these circuits, you may suspect long-tail bias circuits are particularly prone to oscillation. This false impression arises from the reliability of the biasing arrangement. Hit-and-miss or incompletely worked-out base-bias techniques may not bias the transistor into its really active range, so that such circuits will seem on the average to be more stable. They aren't; they are simply half-dead.

fig. 7. Long-tail biasing from two zeners across a single power supply.



A common-base bias circuit does not compel you to use a common-base signal circuit. There are three types of signal circuit applicable to transistor use, and all of them can be fitted into a common-base bias arrangement. Some careful planning will be required, sometimes. For many useful details beyond those included here, look in the GE Transistor Manual, 7th Edition, Chapters 1, 2, and 4 in particular.

The most familiar arrangement is the common-emitter amplifier. Here, we have the emitter at signal ground, the input signal is applied to the base terminal, and the amplified copy appears at the collector terminal. See **fig.** 9.

A large capacitor is required from emitter to signal ground. This capacitor bypasses the bias circuit at signal frequencies, so that gain is possible. Its reactance should be equal to or less than the emitter resistance, which is typically in the ohms or tens of ohms ballpark. A rough estimate is $26/I_E$ ohms for a germanium transistor or $50/I_E$ ohms for a silicon transistor where I_E is emitter current in milliamperes. The low-frequency rolloff begins where capacitive reactance equals emitter resistance. Choose the smallest capacitor that will do the job.

Base input resistance is approximately beta times emitter resistance. This puts it in the high hundreds and low thousands of ohms. Collector output resistance is typically in the 10k's to 100k's of ohms, so that stage output

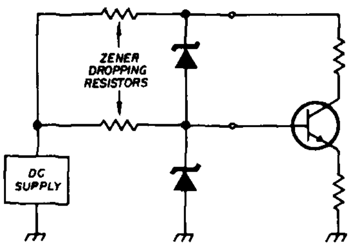


fig. 8. How to combine long-tail biased circuits with conventional base-biased circuits.

resistance is practically equal to the value of the collector resistor.

You have to get the signal in to the base by some arrangement that does not upset the low-resistance dc base-to-ground connection. One approach is link coupling, shown in **fig. 9A.** This is a one-stage i-f amplifier using the tiny i-f transformers available from Lafayette.* The neutralizing capacitor is required. A 470-ohm resistor and .001-μF capacitor are included for decoupling.

If you want to make a high-gain audio amplifier, add another transistor as an emitter follower (more detail on emitter followers) See **fig. 9B.** The roughly 10 mA of current in the amplifier transistor is divided by the transistor beta twice, so that control current through the 100k base resistor produces insufficient voltage across it to upset the common-base biasing. Voltage and power gain are very high.

In the common-base configuration, you apply the signal to the emitter, and take its amplified copy from the collector. The base is fixed firmly at signal ground.

* Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L.I., New York 11791.

Looking into the emitter, you see the same low resistance found in the grounded-emitter circuit, but this time transistor beta is not available as a multiplier. The signal is applied directly to the emitter resistance of a few ohms'or tens of ohms. A large input capacitor is required, having a reactance at the lower rolloff frequency equal to the resistance being fed. Or you can put the signal in by a link or transformer arrangement, but the end opposite the transistor must be well bypassed to signal ground.

The common-base circuit has the highest collector resistance of any transistor configuration, and sometimes you can place it directly across an entire LC circuit. The output resistance is in the 100k's to 1 meg ballpark. Again, if you have a collector load resistor, it will set the apparent output resistance at its own value.

In **fig. 10A** is a very simple audio amplifier At rf you might use link coupling as shown

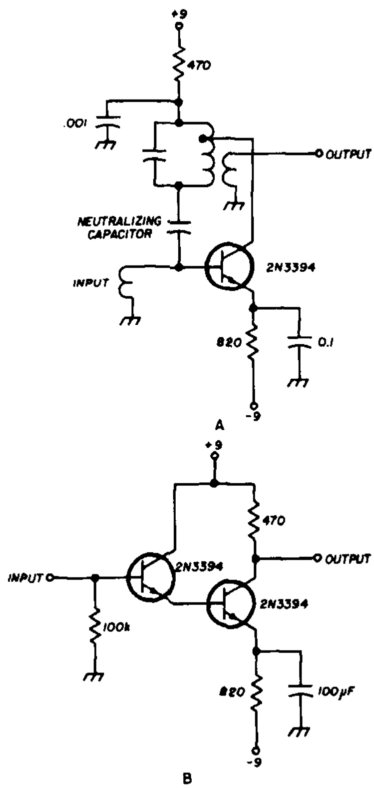


fig. 9. Common-emitter circuits.

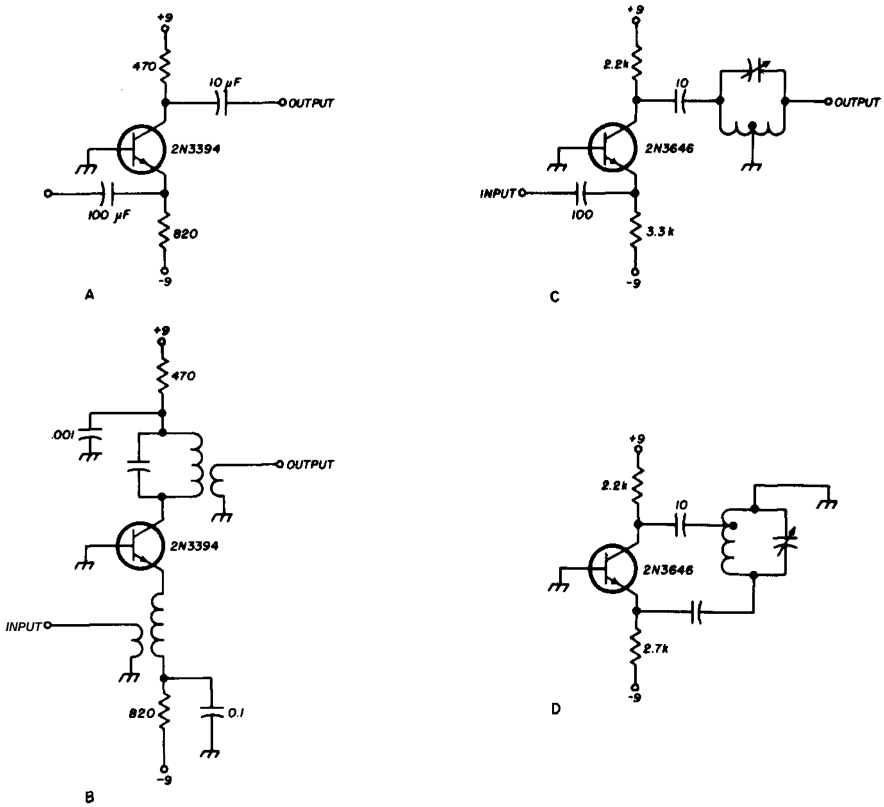


fig. 10. Common-base circuits.

in fig. **10B.** This arrangement is prone to oscillation unless the input circuit is heavily loaded or of low reactance. The feedback is through collector-to-emitter capacitance, and cannot be eliminated.

At vhf, the common-base amplifier becomes **fig. 10C,** which is probably about the simplest vhf amplifier you can build. Heavy input circuit loading is required.

Perhaps the output circuit needs further explanation. This is simply a pi-tuner turned inside out; or you might prefer to look at it as a resonated auto-transformer. Its impedance transformation is adjusted by varying the point at which the coil is grounded. In this case you are transforming about 2k ohms to 75 ohms, so a 5:1 turns ratio is required. The emitter and collector resistors do double duty as decoupling and isolating resistors. Although very good in most respects, this circuit's one shortcoming is a tendency to instability.

If you accept the instability problem, by adding still more feedback you get an oscillator which can be remarkably stable. See fig. **10D.** This is a breadboard VFO assembled for test purposes. It keys well at 145 MHz. The collector is tapped far down on the coil, and for vhf the emitter feedback capacitor is simply a wire close to the hot end of the coil. In some cases, a capacitor from emitter to ground will improve frequency stability.

emitter followers

Finally, there is the common-collector configuration, more generally called the emitter-follower. The signal is applied to the base circuit, and its duplicate appears at the emitter terminal. There is no voltage gain. But there is a power gain which appears as an impedance by a factor which may be as large as the transistor beta. Typical emitter resistances are the $26/I_E$ and $50/I_E$ values that are seen in the common-emitter circuit, but this time

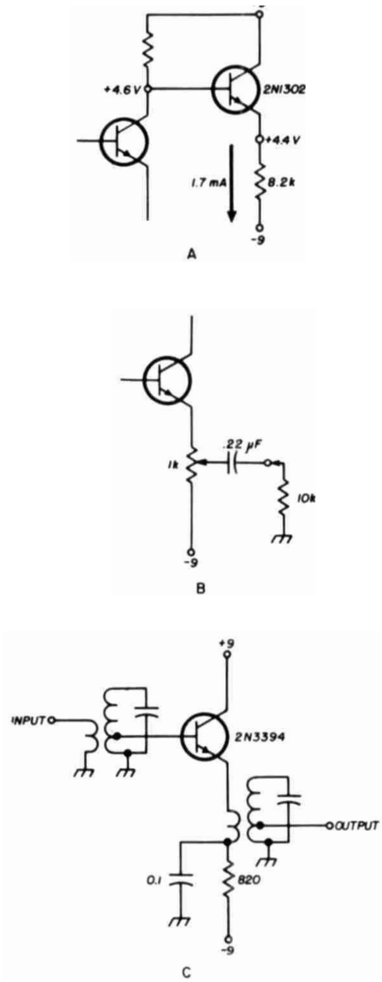


fig. 11. Common-collsctor circuits.

they appear as output resistances. This simple computation is no substitute for finding the manufacturer's specs if they are available. Base input resistance levels are typically beta times larger.

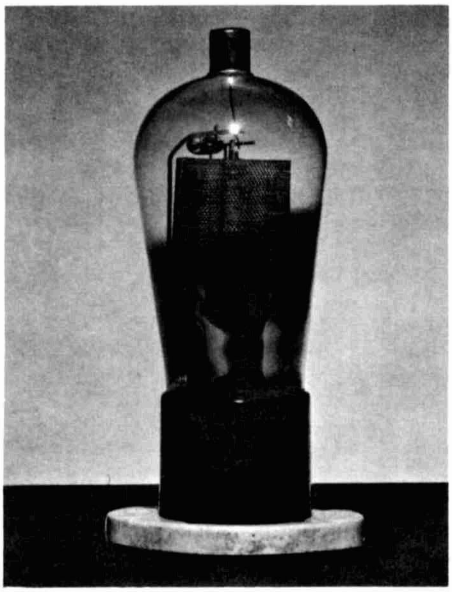
In **fig. 11A** the emitter follower is riding on the preceding transistor's collector. No coupling elements are required, but the follower's base current adds to the preceding stage collector current. Emitter voltage is the base voltage less typical base-emitter voltage, but output resistance is reduced from the thousands-of-ohms level to the tens-or-hundreds of ohms. If we replace the resistor with a potentiometer **(fig. 11B)**, we have a nice low-impedance level attenuator arrangement. This is good for getting adjustable level rf out of a signal generator. The capacitor shown is appropriate for audio applications into a 10k-ohm load. It would be very much smaller at rf.

We can use the emitter follower as an i-f amplifier as shown in **fig. 11C.** If there is no agc circuit, the number of components is reduced to an absolute minimum. If it oscillates, try a transistor with better high-frequency response. Lafayette's miniature i-f transformers are usable in this circuit.

summary

Well, there you are! The long-tail bias circuit requires an extra voltage, but it will save components over an entire project. If you've been frustrated by circuits that don't work at very low or high temperatures, now you know how to design better ones. And you can surprise your friends with good designs that take almost any old transistor and work. . . even silicon and germanium transistors interchangeably. But I've rather neglected the signal view: better do some breadboarding before you carry these ideas into your construction projects.

ham radio



The dynistor—see page 49 for details.

an improved transistor voltmeter and its applications

A transistor voltmeter with excellent linearity, low drift and high sensitivity using two low-cost silicon transistors

R. S. Maddever, Geelong Grammar School, Corio, Victoria, Australia

It is widely known that for accuracy, a voltmeter must draw a minimum of current from the circuit being tested. In recent years we have seen the widespread development of more and more sensitive multimeters at modest prices. However, in applications requiring low loading, the best meter movement is often not as sensitive as you wish. Also, as a rule, fragility, as well as cost, increases with sensitivity.

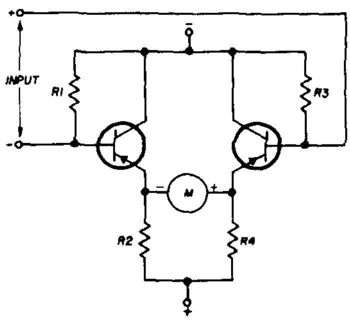
Vacuum tubes have traditionally been used as amplifiers with meter movements of ordinary sensitivity in VTVM circuits to provide low loading. With the appearance of the transistor, it seemed reasonable to assume that it could be pressed into service—to'amplify current directly to a meter connected in its output. It was soon evident that this was a dream; acceptable transistor current amplifiers required elaborate modifications.

problems

The two main problems with transistorized instruments are linearity and temperature-induced drift. Linearity is the ability of the instrument to follow the signal exactly. When the input signal doubles in a perfectly linear device for example, the output signal will also exactly double. The transistor is nonlinear because its gain varies with the current flowing through it. In common general-purpose transistors, the gain often falls by 20% for a 3 to 1 change in collector current. Leakage and gain also vary with temperature. Both can be reduced by the use of negative feedback. Temperature effects can be further reduced by a balanced configuration and the use of temperature compensating elements.

Most of the early articles simply ignored these problems or glossed over them with statements such as, "calibration is good" or "calibration is adequate". A few admitted that simple transistor amplifiers were not linear and needed calibration or calibration curves.[20] One article states simply that, "the

fig. 1. This balanced circuit minimizes the effects of hfe, Vbe and Icbo.



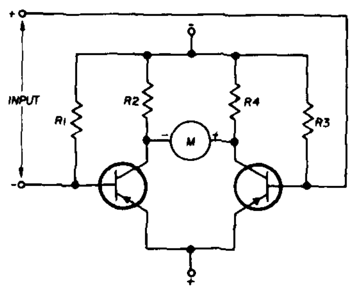
meter shown has a home-made scale which is not difficult to make"."'

If you are prepared to accept the extra trouble and lower performance (as compared with a VTVM), it is possible to use inexpensive germanium transistors in a meter amplifier. A reasonably good circuit along these lines using 0C71 transistors (2N3325, HEP3 or SK3004) appeared in 1958[1]. Since then, many variations have appeared in print, some of which are described in the list of references at the end of this article. Two good germanium transistor circuits appear in references 2 and 6; each of these circuits has its advantages and novelties. A good silicon-transistor design based on the **CE** Transistor Manual was presented in the Equipment **Exchange Bulletin**[4].*

Fortunately, it is now possible to bypass most of these elaborate methods by using a good transistor in a simple circuit. The Fairchild 2N4250, for example, solves many of the problems with gain, noise, linearity and temperature sensitivity because of its high inherent performance. In this article I'll describe the use of this transistor in a practical voltmeter circuit. This circuit is as versatile and sensitive as the ordinary VTVM, but it is smaller, has negligible warmup time, and independent of ac power and line instabilities.

It could be argued that an FET automatically solves all semiconductor voltmeter problems, but this is not altogether true.[13] However, it must be noted that a number of voltmeter circuits using unselected FET's have been published,[14-18] so problems of linearity and reproducibility can be overcome. This article describes a good, but simple circuit using transistors—a circuit with minimum dependence on the characteristics of the individual transistors. Furthermore, in contrast to many FET voltmeters, it uses a low-voltage power supply (with an excellent lack of voltage dependence) and may be used with a 1-mA meter movement.

fig. 2. This circuit also minimizes the effects of hfe, Vbe and Icbo, but has a lower input resistance than fig. 1.



On the other hand, I can't deny that you can get higher input impedance with an FET. Although ordinary VTVM and FET voltmeters limit the input resistance to the 10- to 20-megohm range—typical transistor voltmeter performance—it is possible to increase the input resistance to several hundred megohms! In some cases this level of input resistance may introduce leakage problems. For some ohmmeter circuits, or where very high input resistances are required, an FET voltmeter is unbeatable—if the static charge problems don't plague you with other headaches.

* This is now known as The *Australian EEB* and is a fine informal experimenters' magazine.

basic design

To minimize the effects caused by variations in h, V_{BE} and I_{CBO}, a balanced circuit may be used as shown in **fig. 1** and **2.** The circuit in fig. **1** has a higher inherent input resistance, because of the large series resistance used to obtain low input current. Although its high input resistance is not necessarily a good reason for choosing **fig. 1,** it does have one major advantage over **fig. 2:** R2 and R4 provide negative feedback which improves the already excellent linearity of the 2N4250.

The high linearity of the 2N4250—down to collector currents as low as 1 μA—permits operation at unusually low collector currents.[12] This improves the low-noise performance of the device. Also, the leakage of the 2N4250 is only 10 nanoamperes at 40 volts, and appreciably less at normal operating voltages.

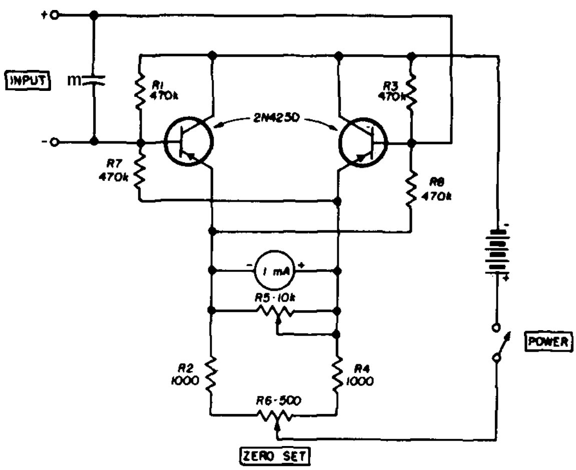
You have to make some provision for transistor gain variations and temperature effects. This is most easily accomplished by adding the potentiometer R6 shown in **fig.** 3. However, with matched transistors and slight adjustments to either R1 or R4, this potentiometer can be omitted or preset. Zeroing is then done with the meter's mechanical zero adjustment.

You can calibrate the unit for a given input current with the resistor across the meter (R5). For a 200 μA meter, the input current will be less than 1 μA for full-scale deflection. One-mil movements require from 3 to 5 μA; the exact value depends on transistor gain. The capacitor across the input leads reduces ac pickup which could give misleading readings.

If you find that the sensitivity is too great and want to reduce it to a more convenient value, such as exactly 1 μA for 200 μA full-scale deflection, the negative feedback should be increased. This is better than reducing the value of R5 which effectively reduces meter sensitivity. The added feedback further reduces temperature drifts and improves linearity. This is done by adding R7 and R8 as shown in **fig. 4.**

If 470k-ohm resistors are used for added negative feedback, they will reduce the sensitivity from 10 to 20 percent. Smaller values will reduce sensitivity still further, but will improve drift performance proportionally. Therefore, during initial calibration, R5 should be maximum and R7 and R8 adjusted until the overall sensitivity is nearly correct. Final adjustments are then made with R5.

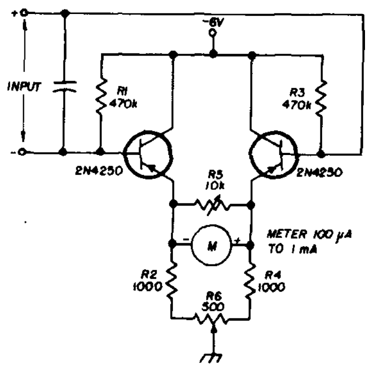
fig. 4. To reduce sensitivity and temperature effects negative feedback is introduced through resistor R7 and R8.



With the resistor values shown in **fig. 4,** the static collector current through the transistors will be 2 to 3 mA. If you're using a 200-μA meter movement, this is needlessly high. You can gain some improvement in noise level and overload characteristics by reducing it to about 1 mA. This can be done by increasing R2 and R4 to 2.7 kilohms, and R1 and R3 to about 1.2 megohms.

To minimize the effect of any sudden temperature change, the two transistors should

fig. 3. Transistor gain variations and temperature effects are nulled out of the basic balanced circuit by resistor R6.



be in contact—taped together or coupled with a narrow copper strip.

meter protection

In the event of a catastrophe, about 5 mA could flow through the meter in the circuit of **fig.** 4—500 μA when the resistors have been changed to reduce the static collector current. Therefore, the meter movement should be protected with diodes, particularly if it is 100 μA or less. This can be done most easily by putting two back-to-back **silicon** diodes across the meter.*

Although this protection technique is very useful, it will not permit indefinite liberties with the instrument. If an excessive input voltage is applied, it is quite likely that the transistors themselves will be damaged. The circuit could be made completely foolproof by putting several forward-biased series-connected silicon diodes across the input. Although a similar scheme has been used[19], it produces slight nonlinearity and hardly seems worth the trouble.

multimeter applications

This instrument is suitable for a wide variety of measurements. As it stands, it is a very rugged and stable current meter with an input sensitivity in the microampere range and input impedance of two hundred-thousand ohms. Therefore, it is very useful for circuits using photo-electric cells at very low light levels, thermistors where small temperature changes are encountered, or as a sensitive galvanometer in bridge circuits.

You can also use this basic circuit as the basis for a very good multimeter. A few of the principles in multimeter design are discussed below, but I will leave it up to you to devise suitable switching schemes. It seems pointless to present a complete switching system for a multimeter, because individual tastes vary so much. Also, it's a lot of fun to design a system to your own requirements.

use as a voltmeter

Since the input resistance of the amplifier is about 200k ohms, we can represent the system as the "black box" in fig. ***5.*** For purposes of this discussion let's assume that the amplifier has been adjusted so that exactly 5 μA is required for full-scale deflection of the meter. The circuit's function as a voltmeter is easily illustrated by Ohm's law.

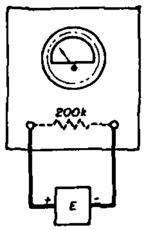


fig. 5. The transistor voltmeter may be represented by a "black box" with an internal resistance of 200 kilohms.

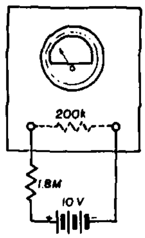


fig. 6. Addition of a 1.8 megohm multiplier for a full-scale reading of ten volts.

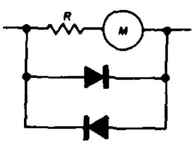
In this case, the circuit is 5 μA, so for a full-scale deflection of, say, 10 volts, the required resistance is:

$$R = \frac{E}{I} = \frac{10 \text{ volts}}{5\ \mu A} = 2 \text{ megohms}$$

Since the internal circuit already has 200k ohms or 0.2 megohms, we'll have to add 1.8 megohms externally as shown in **fig. 6.** If we want the meter to read 20 volts full scale, then we need a total of **4** megohms (3.8 megohms external).

The main difficulty here is to obtain the proper value of fixed resistance. Standard

* This will limit maximum meter current to three or four times the full-scale deflection value. An overload of this sort could bend the meter needle, and for better protection, a resistor should be included as shown here. The value of R is increased until the meter reads about 0.5% low with full-scale current. Then, with an overload, panic current will be limited to about 1-1/2 times the full-scale value, which can be comforting!



values are only available in increments of 5 or 10 percent, but virtually any resistance may be obtained by various series and parallel combinations. With a variable voltage source and a reasonably accurate auxiliary voltmeter, you can quickly determine what resistances are required for full-scale deflection of your transistor voltmeter at various voltages. This is much quicker, cheaper and easier than trying to buy special resistors.

On high-voltage scales, it's possible to run into some problems. For example, 1000 V full scale would require 1000 megohms of series resistance if the input sensitivity is 1 μA. Resistors this large are available at a reasonable price*, but can lead to difficulties with insulation leakage unless your instrument is built and kept scrupulously clean. Methods for constructing practical systems with lower input sensitivity (the 100- to 200-megohm region) are discussed in the **Equipment Exchange Bulletin*.** I might note that the series-parallel method will work just as well for large-value, high-stability resistors. For example, two 1000-megohm units in parallel with 125 megohms give 100 megohms, but the final result should always be compared with the performance of a known voltmeter.

Typical voltmeter ranges you might use are:

1 V Useful for emitter- or base-bias readings, though it may be necessary to take the input current into account for the latter.

5 V 1.5- and 3.5-V circuits

10 V 9-V circuits

50 V 20- to 50-V power circuits

And the usual scales for 100-, 500- and 1000-volts.

Use several resistors in a series for the 1000-V range, coming from a separate plug, not through the switch!

If you are lazy, you can manage adequately with ranges every decade—1-V, 10-V, 100-V, and 1000-V, plus the extra 5,000-V range.

*** Proops Brothers, 52 Tottenharn Court Road, London, W1, England. Welwyn glass-encapsulated types come in 125, 1000 and 10,000 megohrns at 25c each plus postage.**

When you put these resistors around the usual multipole rotary switch, you can arrange them in two ways as shown in **fig.** 7.

For the values shown in **fig. 7,** an input sensitivity of 1 μA is assumed, using a 200-μA meter movement. Although less sensitive meters will work well in the circuit, if you're looking for operation comparable to a VTVM, the 200-μA meter is necessary. A 100-μA meter would be even better. In **fig. 7** an input resistance of 200k is assumed, though this will depend on transistors, and on the feedback introduced by R6 and R7 **(fig. 4).** R5 is adjusted for 1 V full scale at the first switch position. If this cannot be achieved, reduce R_y slightly.

Note here the existance of R_x—this is an extra resistor placed directly in the probe lead to isolate the voltmeter from sensitive circuits. This considerably increases the versatility of the instrument, and allows voltage measurement on high-impedance rf circuits with low-capacitance loading. It's possible to build a small resistor into the test probe by the exercise of nominal ingenuity; it's not difficult. With the probe value shown, the actual voltage applied between points "A" and "B" is about 0.5 V.

For a given current sensitivity, the multiplying resistors can be calculated by Ohm's law, and then adjusted to give exact results by experimentally-determined series-parallel combinations.

The systems of **fig. 7** are "constant current" —the same maximum current is required from the source for full-scale deflection on any range. With 1 μA sensitivity, a 5000-V scale would require the addition of a 4000-megohm resistor to drop another 4000 V if it were connected in series with the 1000-V position (fig. **7B**). This could be made up of four 1000-megohm Welwyn resistors in series, but the resistors should be arranged carefully to minimize leakage paths or voltage breakdown; 5000-V is high voltage!

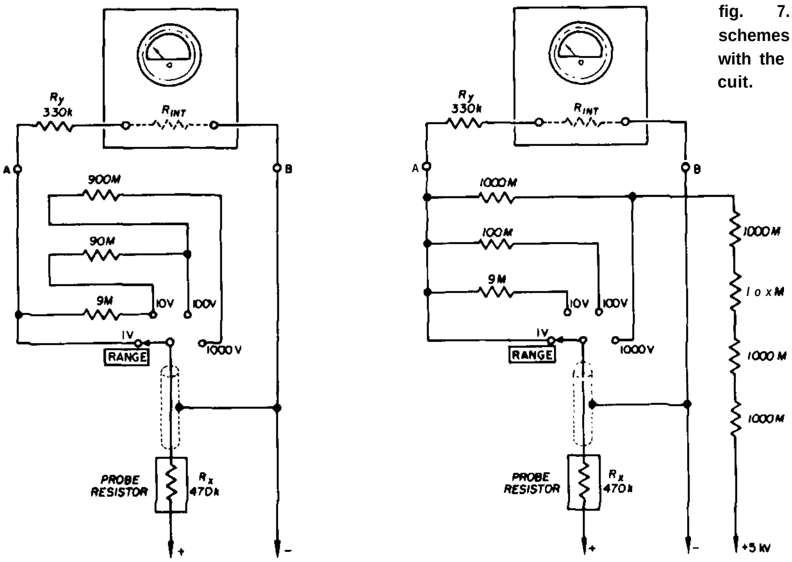
It must be noted that very high resistances used in the voltage-multiplying system shown in **fig. 7** will only give satisfactory service if all the relevant insulation points are perfect. This requires the use of ceramic insulation, including switches, and careful soldering to avoid flux bridges. For ordinary construction, you may find it more suitable to add an extra gang to the rotary switch to place a shunting resistor between points A and B on the 100-, 1000-, and 5000-volt ranges to reduce sensitivity. If you used a 25k-ohm shunt, the series-multiplying resistors would be approximately 5, 50 and 250 megohms respectively on the 100-, 1000- and 5000-volt ranges.

Except for the probe resistor, the multiplying resistors should be "deposited carbon", not the "molded-in-case" type. The former are more exact, and will be considerably more stable with respect to heat and aging. To reduce ac pickup with this high-impedance instrument, you should use shielded wire for the probe lead just as you would with a VTVM.

For selecting the various ranges, you can use a multi-pole switch as shown, or small plugs and sockets. For reliability and safety, the switch is better, but the plug and socket arrangement is better for maintaining low leakage with large multiplying resistors. The exact layout of the complete instrument will not be described here; it is best arranged to suit the requirements of the individual amateur. Similarly, I assume that everyone knows what a voltage divider is and how to use it with a known meter to obtain standardizing voltages.

If your requirements call for a millivoltmeter, the input voltage can be applied directly to the basic circuit. For 1-μA sensitivity, this will provide a full-scale reading with a few hundred mV input.

The use of rectifier probes to convert the instrument to ac measurements is conventional and will not be discussed here. If you're only interested in ac measurements, it may be worthwhile to build an instrument specifically for that purpose as described in Radio-Electronics[X]. To obtain different current scales it is only necessary to put shunting resistors across the input to obtain the desired full-scale reading. For shunting circuits, it is essential that all connections are well soldered, and that switch contacts are clean and have low resistance. Also, it is impractical to use ordinary switch contacts for shunting current in excess of one ampere.



A

B

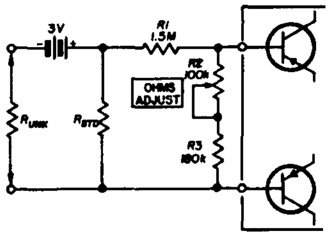
fig. 7. Two multiplier schemes which may be used with the basic voltmeter circuit.

use as an ohmmeter— ordinary ranges

Standard ohmmeter circuits can be found in any reference book or from the operating manual of a good instrument, such as the Simpson 269. An interesting variation of the usual VOM circuit is shown in **fig.** 8. In this circuit, adjustment of the "Ohms Adjust" control has negligible effect on calibration compared with circuits where the control is in series with the battery.

The higher the battery voltage, the larger the maximum resistance that can be measured, and unfortunately, the higher the external current on the low-ohms scale. A simple way to solve this is to use a larger battery voltage for the higher resistance ranges. With a 3-V battery, the lowest practical value of

fig. 8. A variation of the usual VOM ohmmeter circuit.



R_{std} is about 10 ohms, which allows reasonable measurement down to about 0.1 ohm*. For the highest practical value of R_{std}, the largest unknown resistance is about 100 megohms, which is not bad. Increasing the battery voltage to 45 V permits measurement up to 1000 megohms, but only if all the switching and contact terminals have very low leakage.

**Calibration of the low-ohms scale is an unavoidable burden, but it can be made easier by using the following formula:

$$R_{unk} = R\left(\frac{10-1}{V}\right)$$

where:

R_{unk} = Unknown resistance

R_s = Value of R_{std} in parallel with the rest

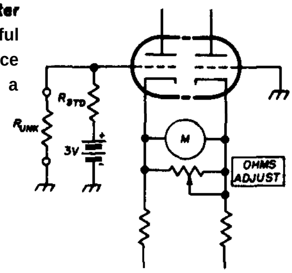
R_s is the same as that value of R_{unk} which gives $V = 5$ when full-scale deflection is $V = 10$; in other words, half-scale deflection. V is the reading of the voltmeter for a given value of R_{unk} with the full-scale deflection taken as 10.0 when R_{unk} is shorted out.

If R_s is known accurately, you can calibrate your entire ohms scale with nothing more elaborate than a pen and a slide rule.

In general, the design center of R2 plus R3 should have about the same resistance as the internal resistance of the voltmeter (without R_y). This can be easily determined directly from **fig.** 8: with R2/R3 disconnected and the input leads shorted, adjust R1 for full-scale deflection (R1 should be approximately 3 megohms.) Now connect a potentiometer in place of R2/R3, and reduce resistance until you get a half-scale reading. The value of this auxiliary resistor will then be equal to the internal resistance of the instrument. For a 200-k input resistance, R2 and R3 should have the approximate values shown in **fig.** 8, and you can proceed from there.

In **fig.** 8 the actual standard resistance seen by the battery will be R_{std} in parallel with the combination of R1 in series with R2/R3 paralleled with $R_{internal}$—whooh! Thus, for R_{std} less than about 0.01 R1, the central-scale resistance will equal R_{std}. This is because at center scale the same voltage is developed

fig. 9. An ohmmeter circuit which is useful with high-impedance circuits such a a VTVM.



across both R_{unk} and R_{std}. When R_{std} becomes comparable to R1 and the rest, the actual standard resistance seen by the battery will be less than R_{std}.

In practice this is nothing to be concerned about because you merely adjust R_{std} to give the high-ohms scale calibration desired when a known value of test resistance is put across the input. This assumes, of course, that you have calibrated the ohms scale with reasonable accuracy for a lower ohms range**.

The maximum ohms scale will be the one where R_{std} is infinity—absent. Since this gives an awkward scale value when compared to the lower ranges, R_{std} maximum is simply adjusted to give the highest nominal center-scale value. For the values shown in **fig.** 8, with 2.7 megohms for R_{std}, the center-scale reading is 1 megohm.

If the standard resistor is, say, 10 ohms, and the battery voltage is 3 V, there will be 300 mA flowing in the probes when they are shorted, or about 20 mA through a forward-biased diode placed across the probe. This is the reason for avoiding the lowest ohmmeter range when measuring the forward-conduction characteristics of transistors. Similarly, the use of a large battery in an ohmmeter circuit will imperil the breakdown rating of some transistors on the highest resistance scales (usually the top two). Therefore, if you must use an ohmmeter to measure the characteristics of semiconductors, be sure to choose an intermediate range.

Nearly every ohmmeter circuit works in essentially the same manner; a voltage is applied to a standard and an unknown resistance in series, and the voltage across the standard is interpreted in terms of the resistance of the unknown. This means that the center scale will read the value of the calibrating resistance (R_{std}). useful measurements may be made on unknowns over a factor of 100 higher or lower than this. In general, you will find it more convenient to assign the value of R_{std} to the scale (e.g. 10 ohms, 1k, 10k, 1M) than the traditional "X1, X100, X1000, etc.". If your standard resistor is 100 ohms, and you call that scale "100 ohms", you will be able to read it more rapidly than if you had to multiply the scale by some constant figure.

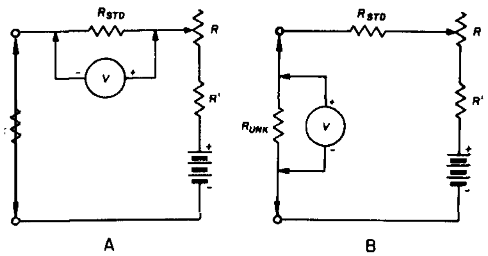


fig. 10. Proportional voltage method of measuring resistance.

There is one other type of ohmmeter worthy of mention: The voltage is measured across the unknown, rather than across the standard resistance (**fig.** 9). It has the advantage that the ohms scale is forward-reading rather than the reverse. However, consistent scale calibration is only possible when the resistance of the voltmeter is appreciably higher than the highest value of the unknown resistance to be measured. It is, therefore, only well suited for VTVM circuitry. Unfortunately, even the best transistor voltmeter draws an order-of-magnitude more input current than a VTVM grid. Maybe not for an FET, but an FET can have problems[13] or can require elaborate circuitry for best results.

The circuit of **fig.** 9 has another interesting advantage in addition to the forward reading scale: the internal resistance of the battery may be taken into account by reducing the value of R_{std} by an equivalent amount. If R_{std} has been reduced by exactly the internal resistance of the battery, this will give an accurate reading over the whole scale.

This system could be used to good advantage on the lowest ohms range if extra switching were provided, but note that the zero adjust must now be done directly at the meter. The system of **fig.** 9 would be practical with the transistorized voltmeter on the low resistance ranges because of the high internal resistance of the TVM compared to R_{std} in this case. Unfortunately, it would require an extra forward-reading scale, which hardly seems worth the relatively small improvement in accuracy.

use as an ohmmeter—proportional voltage method

This method is somewhat cumbersome, but is capable of considerable accuracy down to very low values of the unknown resistance and is discussed in detail in **Radio-Electronics.**[21] We shall assume that the internal resistance of the voltmeter, V, is much higher than that of R_{unk} or R_{std}. The voltmeter is placed across R_{std} as in **fig. 10A,** and R is adjusted for full-scale deflection. Then V is placed across R_{unk}, and resistance is read directly on the ordinary linear scale. If the unknown is, for example, 0.5 ohm, the scale with maximum reading "5" can be considered a "0.5-ohm" range, with all readings in proportion, as you would expect for an ordinary voltage reading. This is because the current through the standard and unknown is the same because they are in series; therefore, the voltage across them is proportional to their resistances—in accordance with Ohm's law.

If the maximum full-scale deflection is 0.5 ohm, it is possible to measure down to 0.01 ohm, which is very convenient. A limitation of this method lies in the fact that the voltmeter leads must be placed directly across the standard or unknown to obtain full accuracy for low resistances. Since it is as cumbersome to build this capability into a multimeter as to do it outboard, it is advisable to set up a special breadboard with the standard resistances and leads for this purpose. Although only about half of the total range can be covered compared to the methods of **fig. 8** or **9**, accuracy is constant over the entire range, and no additional scales are required.

To keep the current through the unknown to the lowest value (and thereby reduce battery drain), it is desirable to use a voltmeter with the maximum possible sensitivity. This is best accomplished by using a given meter without any additional series-multiplying resistors. For an ordinary meter movement at V in **fig. 10,** the sensitivity will be about 100 mV full scale; for the transistor voltmeter, it will be a few hundred mV, depending on the meter used. If we assume 200 mV across a 0.5-ohm standard, the current will be 400 mA, and R plus R' will be about 7 ohms for a 3-V battery. For 200 mV across a 5-ohm standard, current will be 40 mA, with ten times R plus R'. Therefore, a 100-ohm pot at R and a fixed 5-ohm resistor at R' should suffice. If still higher ranges are to be covered, a separate R should be used for each range; this is a good idea for the outboard system in any event.

If an ordinary meter movement is used at V, the internal resistance will be several hundred ohms. This limits the maximum practical value of R_{std} to about 10 ohms if scale calibration is not to be affected appreciably. With a transistor voltmeter the internal resistance is very high and R_{std} could be used to about 5 k. For a vacuum-tube voltmeter there would be **no** great advantage in increasing R_{std} above 10 megohms because the circuit in **fig.** 9 would cover an appreciably larger range for a given standard resistance.

Although it is impractical to use the method of fig. 10 for more than a few ranges because of the many components required and the limited range covered by each standard, it should be kept in mind as a relatively simple circuit for obtaining high-accuracy resistance measurements—particularly at low resistances—without the extra complexity of a balanced bridge.

performance

In a unit I built from fig. 4, one transistor had a gain of 250, the other 450, but the results were very good. When a 200-PA meter was used, the input required about 0.6 μA for full-scale deflection before feedback was applied. After feedback and slight adjustments to R5 to give full-scale deflection for 1 μA, the zero drift appeared to be less than 2% of full scale between zero and 30°C. The gain for a given input current changed by less than 3% from zero to 30°C and less than 8% from zero to 45°C. The long term stability is within 2% and these variations have always been within the mechanical zero set for thr meter. With the feedback resistances shown in fig. 4 linearity is excellent—better than I can discern with an ordinary meter.

Less than 1% (of full scale) zero shift is observed when the leads are shorted. Therefore, the same zero adjustment can be considered satisfactory for all resistances across the input. Increase in battery voltage has negligible effect on sensitivity and a 25% decrease reduced the sensitivity by a mere 3%. In this respect, at least, the unit appears to be far better than several differential integrated microcircuits where the gain is often very dependent on supply voltage. The inherent linearity of the 2N4250 system is also appreciably better—and the 2N4250's are cheaper!

Later units built with more closely matched transistors seem to give even better performance, and a hit of effort spent in matching transistor gains could prove rewarding. This circuit seems to he the simplest and best yet for transistorized voltmeters. It owes its success to the high performance of the 2N4250 transistors.

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ham radio

more 144-MHz moonbounce to Australia

Just as we were about to go to press, I received the news that the EME path to Australia had been conquered again. This time by Henry Theobalt, KØIJN, of Minneapolis. The amazing thing about this contact is that this was Henry's first moonbounce schedule, and he had heard his own echoes for the first time just fifteen minutes before schedule time.

The 160-element collinear that he had put up this winter in sub-zero temperatures was really working. In addition to working Ray Naughton, VK3ATN, he was hearing Mike, K6MYC, Bill, W6YK and Ross, WB6DEX, although he couldn't identify WB6DEX's call.

Just before schedule time, K6MYC reports that he turned everything on, sighted the antenna on the moon and immediately heard two signals. Tuning to the strongest one, he copied KØIJN calling K6MYC—a thrill because there had been no schedule setup. After the VK3ATN schedule, Mike and Henry couldn't copy each other very well, but they were still hearing each other's echos.

After Henry's EME QSO with VK3ATN, K6MYC called Ray on ssb. Ray was experiencing difficulty with local line noise, and didn't copy. Mike was hearing his own ssb echos, but they were not readable. All in all, it sounds like a very exciting evening.

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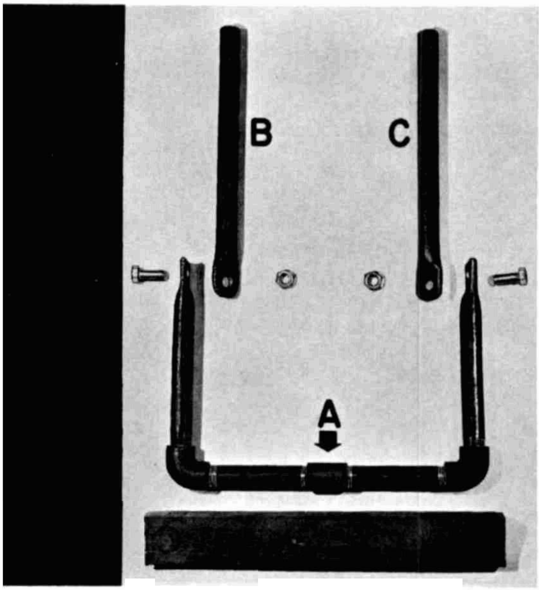

The tiltover tower base. Tha U-Shaped piece is buried in cement; the two uprights. B end C. are slipped into the tower legs. The hinge is formed by the 3/8-inch bolts.

a low-cost tiltover tower base

Ted Wo lw BP, Ce Road, Shrewsbury, Massachusets 01545.

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!

ham radio

EQUIPMENT FROM THE SATELLITE JUST ARRIVED NEWEST MODULE PEG TYPE BOARDS

All parts are useable as leads are wrapped around terminals. (not soldered) Therefore, no damage by heating. All parts can be removed simply by unwrapping leads from around terminal.

Boards contain:

Switching transistors, zener diodes, (signal level and power), signal level diodes, precision trim pots, resistors, adjustable pots, precision capacitors, pulse transformers.

Identification listed below:

This is a sample on various boards.

Some have as many as 25 of one type of listed below

Number on component

LN575	6.2 Volt zenier diode 400 Millowatt cap.
2N1051	NPN Transitor Hi gain 40 Volt P.I.V.
40044	NPN Silicon transitor General purpose Hi gain — small signal 50 Volt P.I.V. MAX. Power — 350 Millowatts
40045	NPN Silicon transitor Hi gain — small signal 50 Volt P.I.V. 2 amp. cont.
40056	Switching Diode P.I.V. 62 volt 400 Millowatts

Also boards are peg type construction. Terminals can be removed and placed anywhere in boards to form your own circuits.

PRICE: $1.50 ea. or 5/$6.00 f.o.b.

VARIABLE VOLTAGE DUAL PLATE POWER SUPPLY

Input: 117V 60 cy.
Output: #1 0-600 V DC @ 0.25 amps
Output: #2 0-600 V DC @ 0.25 amps
Output: #3 12.6 V DC @ 2A fixed

These power supplies consist of two separate supplies on one 19" wide x 10½" high for rack mounting. Chassis is 13" behind the panel. There are two separate variacs used for voltage control and one circuit breaker for protection. Both outputs are well filtered with a Pi section filter. Uses two type 5R4 rectifiers.

PRICE: $49.95 f.o.b. shpt. wt. 80#

ARC-5 TRANSMITTER LIKE NEW 3-4 mHz.

Complete with all tubes. One of the most versatile pieces of surplus eqpt. ever. Get yours now — not many of these left. Excellent condition.

PRICE: $12.95 ea. f.o.b.

NEW CRYSTALS

100 crystals in a case originally for type MAR eqpt. These crystals cover the range of 4844.44 kHz to 7778.78 kHz. There are 16 units in the 40 meter amateur band, and an additional 7 units are useable in the 6 meter band. Also 17 are useable in the 10 meter band. Complete with case.

PRICE: $12.95 or 2/$25.00

ZERO CENTER 4" METER

1 ma. amp. movement. Calibrated 100-0-100, good for TT distortion measurements.

PRICE: $8.50 ea.

AUDIO CHANNEL FILTERS

Centered on following freq:

595 cy.	1105 cy.	1765 cy.
765 cy.	1445 cy.	1955 cy.
935 cy.	1615 cy.	

Impedance: 600 ohms

PRICE: $10.00 ea. or 2/$18.00

TT 63A/FGC

TT regenerative repeater accepts on-off signals whether tone or UC. See Feb. 1967 issue of RTTY far conversion information for frequency shift. NEW.

PRICE: $34.95 ea. f.o.b.

FLANGELESS SILICON DIODES

All rated at 1 5 amps

50-200 volts	6c ea	20/$1.00
200-400 volts	9c ea	15/$1.00
400-600 volts	12c ea	10/$1 00
600 plus	15c ea	7/$1.00

TRANSFORMERS

1000-0-1000 or
1200-0-1200
200 ma. CCS
110 or 220 Volts

PRICE: $9.95 ea.

DIRECT PLUG-IN REPLACEMENTS

ratings 1 amp. 1800 P.I.V.

Replace the following tube types — #5Y3, 5U4, 5Y3G, 5Y3GT, 5V4, 5V4GT, 5AU4, 5T4, 5W4, 5Z4, 5AW4, 5V3, 5AS4, 5AX4, 5AZ4.

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5R4, 5R4WGY replacement ratings 4000 P.I.V. @ 1 amp.

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All above merchandise subject to prior sale. All merchandise guaranteed.
All prices f.o.b. our warehouse, Philadelphia, Pa. Minimum order $5.00

SELECTRONICS

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Phone: Area Code 215 HO 8-7891 · HO 8-4645

Galaxy Solid-State Receiver

A solid-state general-coverage receiver of advanced professional design has been announced by Galaxy Electronics. It will cover from 0.5 to 30 MHz in 500-kHz segments and will have 1-kHz dial accuracy over this range. Among the many interesting features offered by this receiver are an adjustable noise blanker, a variable rf attenuator at the antenna input and an adjustable BFO control for RTTY. Stability is assured with the use of a phase-locked fundamental oscillator and permeability-tuned VFO. Crystal-lattice filters are used in the high-frequency i-f strip for optimum selectivity characteristics. Rear outputs are provided for the PTO, high-frequency i-f, avc, rf gain control and audio to permit dual and space diversity operation with a minimum of additional equipment.

Although this is a professional piece of equipment, its price is not out of range of other equipment offered in the amateur field.

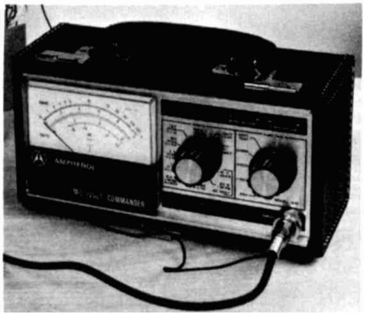
This receiver should be available as you receive this magazine at a price in the range of $700. For further information on the Galaxy R-530 receiver, write Galaxy Electronics, 10 South 34th Street, Council Bluff, Iowa 51501.

EACO Coaxial Switches

A new coax switch has been introduced by the Electronic Applications Company. EACO surveyed the market and found that a four-position switch would answer 90% of amateur requirements. This new switch features the concept of not paying for unused positions. In-line connectors are offered to facilitate installation behind a panel. The escutcheon features a surface suitable for writing. A separate escutcheon is also available for front-panel use when the switch is mounted behind the panel.

These switches are available in two types: the four-way switch and an in-out model for use with a linear amplifier or other accessories. The silver-plated contacts arc designed to handle up to 1000 watts of a-m or 2000 watts of sideband. Negligible insertion loss is claimed up to 160 MHz with a maximum SWR of 1.2 at that frequency. Various types of connectors can be supplied. These switches are priced at $7.65 each from Electronic Applications Company, Route 46, Pine Brook, New jersey 07058.

Amphenol FET VOM



Here is a FET VOM which offers several features of interest to the amateur. The Model 870 **Millivolt Commander,** introduced by Amphenol, can measure voltages as low as 0.1 volt dc full scale or .01 volt ac full scale. These scales can be of tremendous value in servicing and debugging solid-state equipment. By comparison, a standard voltohmmeter might have a maximum sensitivity of 1.2 volt full scale. This instrument features a single probe for both ac and dc use and is rated for 2% dc accuracy full scale; 3% ac accuracy. It is battery powered, and in normal use, the shelf life of the battery will equal battery life. The unit weighs slightly over 4 pounds and is supplied with a lid for full protection when not in use. A pouch in the lid provides room for the probes.

The suggested retail price is $99.95. Further information may be obtained by writing Amphenol Distributor Division, Amphenol Corporation, 2875 South 25th Avenue, Broadview, Illinois 60153.

1968 Radio Amateur's Handbook

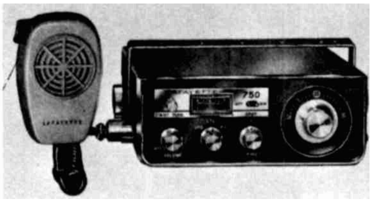
In case you haven't noticed, the new edition of the **Radio Amateur's** Handbook is now available. Doug DeMaw, W1CER, the new handbook editor, has added a lot of new information in the latest volume. Obviously, you can't completely change a handbook of this magnitude in one year, but overall, the editor has done a tremendous job.

The semiconductor chapter has been enlarged to include some typical transistor circuits plus text on FET's and integrated circuits. In addition, transistors are used in many of the construction projects in the rest of the book. There are a few projects carried over from the last edition, but there are lots of new projects, including a FET converter for 40 and *80,* a 75-meter ssb transceiver, a transistor five watter for the novice, a stable FET VFO, and new amplifiers for 432 and 1296.

The VHF and UHF chapters have been completely overhauled with lots of interesting ideas for receivers, converters, transmitters and antennas. Even the appendix has been changed! Most of the low-cost transistors that are suitable for amateur work have been Included in the data section. Although the list is not too long, Doug has chosen types that will satisfy most amateur requirements. This simplifies the task when you are trying to choose a transistor from the several thousand types that are currently available.

If you haven't seen this new volume yet, you owe it to yourself to take a look at it the next time you're in the local electronics emporium. A best buy at $4 from your local distributor, or you can order directly from the American Radio Relay League, 225 Main Street, Newington, Connecticut 06111.

Lafayette 6-meter Transceiver



Lafayette Radio Electronics has announced a new solid-state 6-meter transceiver. This looks like a perfect low-power rig for keeping in touch with the local gang from your car or shack.

The transceiver includes a VFO and is designed to cover the busiest half of the six-meter band (50-52 MHz). The receiver includes one stage of rf and three stages of i-f amplification, while the seven-stage transmitter has a power input of 5 watts to the final rf stage. The transmitter can be either VFO or crystal controlled. Other features include an S-meter, spotting switch, noise limiter and a huilt-in speaker. It can also he used as a 2.5-watt PA system with an external speaker. The Model HE-750 is designed for 12 volt dc (negative ground) operation and includes a power cable, mobile-mounting bracket and microphone. Both a 115-ac supply (HE-502) and a portable battery supply (HE-506) are available as accessories. The price is $119.95 from Lafayette Radio Electronics, 111 Jericho Turnpike, Svosset, L. I., New York 11791.

Estes Noise Suppression Kits

Estes Engineering Company has announced new "bubble-pack" kits of noise suppression accessories for your mobile installation. Kits available include ignition suppression kits, alternator filtering and shielding kits, dc power-line filter kits and universal suppression kits. These packs will give you all the parts you need in one package thus simplifying the organization of your shielding problem. For further information write Estes Engineering Company, 543 West 184th Street, Gardena, California 90247.

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Advertisers iNdex

For Sale--More Usable Sideband Talk Power

A new design of distortion free audio clipper called the Comdel CSP 11 has been made available to the trade. Frankly, we are enthused about its performance from actual on-the-air tests. Two NCX-5's were hooked up to a SPDT coax antenna switch. One was barefoot with the processor in series with its mike. The other was using a BTI loaded to one kilowatt input and the same mike without processer. **Reports indicate almost comparable results. I** am not saying that the Comdel replaces the linear, hut I do say that the greatly increased talk power is most obvious and therefore advantageous in pile ups or when the going gets rough.

The Comdel speech processer is in itself a complete miniature sideband transmitter and receiver with a common oscillator. Its circuitry includes filters and limiters designed so as to keep out the "crud" and at the same time, raise the average level of the spoken voice by a factor of 10 db. Each human voice is different. Various qualities of inflection, euphonics, and amplitude are characteristic of each of us individually. Our human voice has a notoriously low mean-to-peak signal ratio. Hence, the average signal, which determines the loudness at the receiving end, is only a small fraction of the total available peak power output. Conventionally, clippers are effective for increasing the mean-to-peak power ratio at the expense of severe and often objectionable harmonic distortion. Normally this distortion limits the usefulness of these clipping devices. In the Comdel speech processor, the objectionable harmonic distortion is absent and the intelligibility is enhanced by the unique circuit shown in the block diagram.

Conventional sideband transceivers or transmitters have power supplies which are designed for a duty cycle of about 15 to 25%. Application of the Comdel speech processor will make it necessary for the power supply to bear a substantially greater burden, since the average power is now approximately 60%. Thus, the average transceiver of and by itself cannot be utilized advantageously by the Comdel. But if you have a transceiver which drives a linear with lots of room to spare, or if the linear that you have has a real bruiser of a power supply, such as may be found with the Henry II K or the BTI, or some Collins linears, or most home-brew linears, then the Comdel will positively amaze you with its effectiveness. The unit is completely transistorized and requires but 9 volts of DC at 18 miliamperes, with the negative side grounded. This power may be supplied by dry cells or by means of a dropping resistor from a higher voltage supply. The front panel provides an in and out switch which connects the microphone straight through the equipment, or shunts the microphone around the equipment, depending upon your own wish. The volume control provides a means for setting the peak level when the device is turned on. The Comdel is priced at $120.00, postage paid, in the continental limits of the United States. We have this very advantageous tool in stock, for immediate shipment, and we are heartily endorsing this product for use by radio amateurs or even commercial sideband stations. Literature is available for those seeking it.

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- Optimum frequency response for voice
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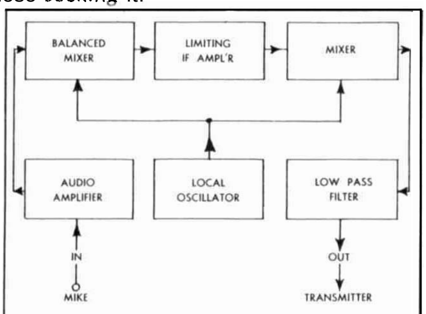

BLOCK DIAGRAM CSP 11

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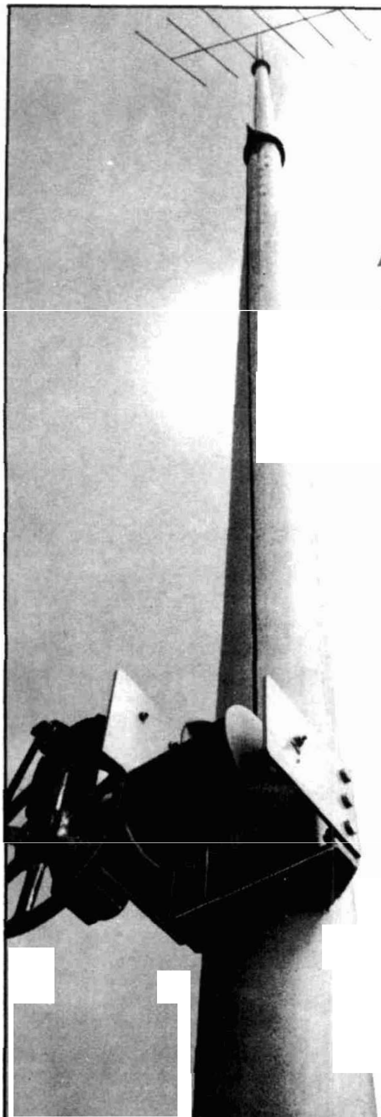
Woodchuck Hill, Harvard, Mass., 01451

Telephone 617-456-3548



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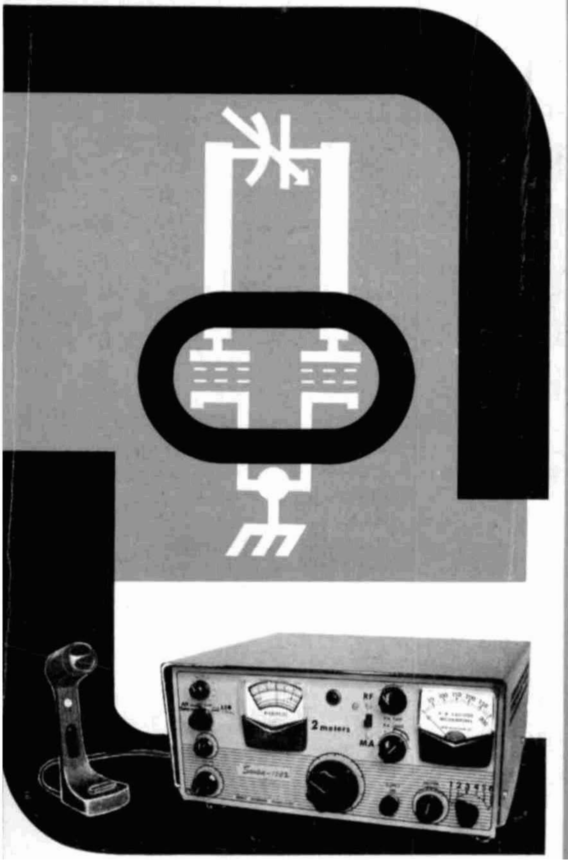

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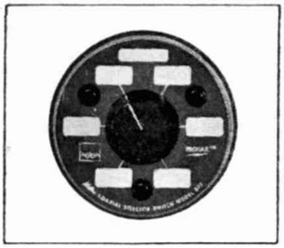

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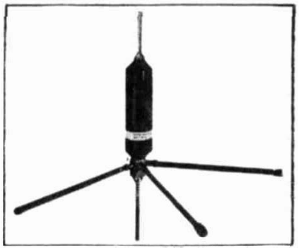

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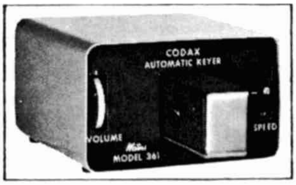

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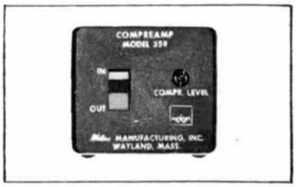

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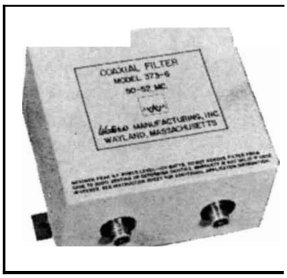

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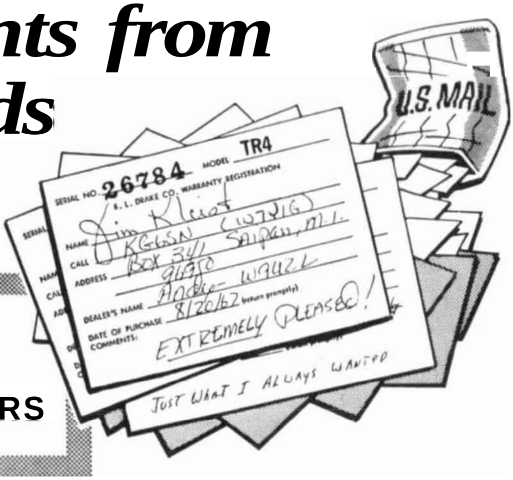
Some comments from warranty cards
by owners of



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. . . or write for details. . .

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may 1968

volume 1, number 3

ham radio

contents

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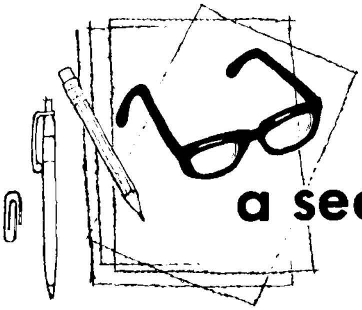

a second look

by jim fisk

How do you sell ham radio? How do you bag it and tag it and interest a customer in buying it? How do we get new hams? I wish somebody would come up with the secret; I've been looking for it for a long time and really haven't found a satisfactory solution. One thing I do know, before you sell it, you have to advertise it; and advertising means publicity.

The biggest thing that ham radio needs today is publicity. Good publicity and lots of it. There are lots of easy ways to do this and some take little effort.

As an example—have you noticed the many signs along the highways saying, "We monitor channel 9"? Amateur radio clubs should do the same type of thing. This is a small matter, but if you had a sign saying: "Attention amateur radio operators—we monitor 7255 kHz," it would help visiting harns. At the same time, it would publicize amateur radio.

I think it's pretty obvious that citizens band has stolen a lot of the glory of ham radio over the past four or five years. Many of the youngsters who would normally be interested in hamming have gone the easier route to CB Most Americans have heard of citizens band and know something about it, hut I wonder how many know of amateur radio? I think you would be quite surprised if you ran a poll.

Amateur radio gets its biggest publicity boost when hams furnish communications during times of disaster and catastrophe. But there are lots of other occasions when amateur radio is equally useful The occasions don t reach the national import of the Alaskan earthquake or the large Southeastern and Caribbean hurricanes, but nevertheless, amateurs can and do furnish a very good communications service in times of public need

Perhaps the best example of this is the West Coast Amateur Radio Service (WCARS) This group, started about five years ago, monitors 7255 kHz daily with a roll call at noontime WCARS enjoys a large membership and through the work of its publicity chairman, Ed Gribi, WB6IZF, news releases appear in the California, Oregon and Washington papers quite frequently

For some reason amateurs in other parts of the country have not picked up the idea However, Dave Flinn W2CFP, has volunteered to start an East Coast net He will monitor 7255 kHz each noontime and asks that anybody on the East Coast who can hear his signal check in with him Perhaps the East Coast group will grow to the extent that WCARS has

In a recent letter to me, Dave indicated that he would like to start the ball rolling along these lines, but that he doesn t want to duplicate the efforts of any other amateur If you have started a public-service net, or know of anyone who has, please let Dave know so he isn t duplicating someone else's efforts Perhaps with an East Coast net, as well as WCARS, amateurs in other parts of the country who are out of the range of WCARS and W2CFP will start their own nets

WB6IZF told me recently that the WCARS net often gets into the inland states during the evening hours on 7255 kHz. If we can get amateur public service nets started in other parts of the country, we can develop a liaison between these nets on some other frequency hand.

What better way to obtain publicity than to provide a public service? It's just the kind of thing that newspaper reporters and radio announcers are looking for. Interestingly enough, on New Years Day, net control of WCARS took roll call just to see how many amateurs would check in; almost 100 amateur stations were monitoring the frequency. During working hours, of course, there are not that many stations listening to 7255, but, nevertheless, there are always a few mobilers or retired persons at home who are able to pass traffic.

One member of WCARS is WA6PCY, a member of the California Highway Patrol. John has a mobile station mounted in his patrol car for immediate liaison with other WCARS members.

If you're interested in forming a public-service net and are not within range of W2CFP or WCARS, why not write to WB6IZF. I'm sure Ed would be more than happy to give you the necessary details for forming an amateur-radio service group. Ed has been busy for the past several months sending out WCARS publicity releases with the result that it is quite widely known in the western part of the United States.

If you're like me, you'll have the receiver turned on while reading the mail and not actually operating. This is ideal when you're doing a home construction project, cleaning out the shack, or making out QSL cards. As long as you're reading the mail, you might as well do it for a worthwhile purpose.

With amateur radio service nets throughout the country, there would be thousands of amateurs tuned to 7255 kHz any time of the day or night, 365 days a year. In addition to helping the general public in times of emergency, it would be very helpful for mobile hams who run into trouble on the road. It's always nice to know that you have a friend that you can rely on just by pushing the mike button.

In addition to the frequency-monitoring signs and public-service nets, there are other ways to get in the public eye. How about a weekly amateur-radio column in the local paper? Most newspapers have space that needs to be filled. If you can't write the material yourself, there must be a member of the local club who can.

There are a lot of interesting things going on in amateur radio that the general public would he interested in. There have been a number of news stories concerning phone patches to Viet Nam, slow-scan television transmissions to Antarctica and amateur radio on board the Queen Mary's last voyage, just to name a few.

Once the stage is set with a human-interest story, the skillful writer can throw in some sales pitches for amateur radio. What other hobby can you think of allows the bed-ridden and handicapped to stay in touch with the world? What other hobby can provide a training ground for electronics engineers and technicians? What other hobby is as many-faceted as amateur radio? What other hobby can promote as much international friendship and understanding as amateur radio?

There are a lot of potential amateurs out there that just need to be sold on the idea. After they're sold, it's up to us to help them along. What's simple for the old timer is complex to the newcomer. When a newcomer asks you a question, try to give him a straightforward answer. Help him find those hard-to-get parts. Open up your junk box and give him some of that stuff you've been saving for twenty years. You can't make him into a ham, but you can certainly help him along the right path. It's time we got off our haunches and made amateur radio grow.

Jim Fisk, W1DTY
Editor

RELIABILITY QUALITY VALUE

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The Model 410C Frequency Control Unit is designed for full coverage of 80, 40, 20, 15 and 10 meters. It is intended for fixed station operation and plugs directly into Model 500C. It may also be used with Model 350C. Eight ranges, 500 kc each, 5 kc calibration.

$115

DUAL VFO ADAPTOR

Provides for the addition of second VFO for separate control of transmit and receive frequencies. Plugs directly into Model 500C and may also be used with Model 350C and other Swan transceivers.

MODEL 22 . . . . . $25

MARS OSCILLATOR

Five crystal controlled channels with vernier frequency control. Plugs directly into Model 500C and may also be used with Model 350C and other Swan transceivers.

MODEL 405X (less crystals) . . $45

SWAN 500C SSB-AM-CW TRANSCEIVER

Five band, 520 watts for home station, mobile and portable operation.

The new model 500C is the latest evolutionary development of a basic well proven design philosophy. It offers greater power and additional features for even more operator enjoyment. Using a pair of the new heavy duty RCA 6LQ6 tetrodes, the final amplifier operates with increased efficiency and power output on all bands. PEP input rating of the 500C is conservatively 520 watts. Actually an average pair of 6LQ6's reach a peak input of over 570 watts before flattopping!

The 500C retains the same superior selectivity for which Swan transceivers are noted. The filter is made especially for us by C-F Networks, and with a shape factor of 1.7 and ultimate rejection of more than 100 db, it is the finest filter being offered in any transceiver today.

For the CW operator the 500C includes a built-in sidetone monitor, and by installing the Swan VOX Accessory (VX-2) you will have break in CW operation.

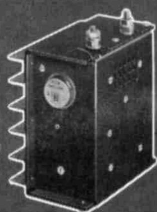
Voice quality, performance and reliability are in the Swan tradition of being second to none.

$520

SWAN 117XC MATCHING AC POWER SUPPLY

Complete A.C. supply for 117 volts, 50-60 cycles, in a matching cabinet with speaker, phone jack, and indicator light. Includes power cable with plug for transceiver, and A.C. line cord. Ready to plug in and operate.

$105



SWAN 14C DC CONVERTER

Converts the above 117XC A.C. power supply to 12 volt D.C. input for mobile, portable, or emergency operation.

$65

SWAN SPEAKS YOUR LANGUAGE . . . ASK THE HAM WHO OWNS ONE

POWER

VERSATILITY

SWAN MARK II LINEAR AMPLIFIER

Two Eimac 3-400Z Triodes provide the legal power input: 2000 Watts P.E.P. in SSB mode or 1000 Watts AM or CW input. Planetary vernier drives on both plate and loading controls provide precise and velvet smooth tuning of the amplifier. Greatly reduced blower noise is provided by a low RPM, high volume fan. Provides full frequency coverage of the amateur bands from 10 through 80 meters and may be driven by any transceiver or exciter having between 100 and 300 watts output.

$395

PLUG-IN VOX UNIT

Plugs directly into Model 500C, and may also be used with Model 350C and other Swan transceivers.

MODEL VX-2 . . . . $35

SWAN 350C SSB-AM-CW TRANSCEIVER

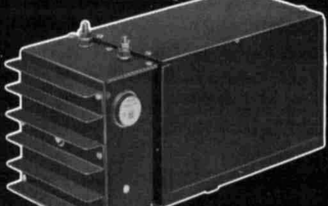
Our improved standard 5 band model, now in production and still only . . .

$420

MARK II POWER SUPPLY

May be placed beside the Mark II, or with its 4½ foot connecting cable, may be placed on the floor. Silicon rectifiers deliver 2500 volts D.C. in excess of 1 ampere. Computer grade electrolytic filters provide 40 mfd capacity for excellent dynamic regulation. A quiet cooling fan allows continuous operating with minimum temperature rise, thus extending the life and reliability of all components. Input voltage may be either 117 or 230 volts A.C.

$235



SWAN 14-117 12 VOLT DC SUPPLY

Complete D.C. supply for 12 volt mobile or portable operation. Includes cables, plugs, and fuses. Will also operate from 117 volt A.C. by detaching the D.C. module & plugging in 117 volt line cord. Negative ground standard. Positive ground available on special order.

$130

Illustrated on these pages is a complete Swan amateur radio station, one of the finest money can buy. Starting with the powerful 500C and an AC power supply, you are immediately on the air with a big, high-quality signal. Thanks to the excellence of the high-frequency crystal lattice filter, made especially for Swan by CF networks, you will have one of the cleanest and most readable signals on the air, as well as outstanding receiver selectivity and sensitivity. The various accessories from the Swan line may be added at any time, providing greater operating pleasure and performance. The tremendous acceptance of Swan products by radio amateurs throughout the world is most gratifying to all of the people at Swan. It is our continuing policy to offer the finest communications equipment we know how to design and manufacture, with quality control, craftsmanship, and service that is second to none.
73

SWAN
ELECTRONICS
OCEANSIDE, CALIFORNIA
A Subsidiary of Cubic Corporation

converting the Swan 120 to two meters

With increased ssb activity on two meters, this transceiver should be a popular unit for mobile or fixed operation

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

My first vhf conversion of a Swan 120 was a unit that I put together for six meters.[1] It worked so well I bought another single-band Swan transceiver for this conversion to two meters. This conversion results in a complete ssb/a-m or CW transceiver with good sensitivity and selectivity. The transmitter puts out about 40 watts on ssb or CW using the original power supply.

receiver

The modifications to the receiving section are quite similar to the six-meter model. I used the same front-end design, changing the tank circuits to tune 144-147 MHz. The cascode rf amplifier works as well at 2 meters as it did on six; it exhibits low noise figure and more than adequate gain. The mixer also provides very good results on two.

The incoming rf signal is picked off the final plate circuit and fed into the cascode preamplifier through a 2.2 pF capacitor. This simplifies the problem of what to do about transferring the antenna from transmit to receive. The original 6BA6 rf amplifier is completely disconnected; the new front end feeds directly into the 12BE6 mixer.

An agc circuit is incorporated with separate rf and audio gain controls. With this circuit, the audio level is held constant with very little popping and clistortion. During round-table QSO's, the rf gain can be turned up with good results with both weak and strong signals.

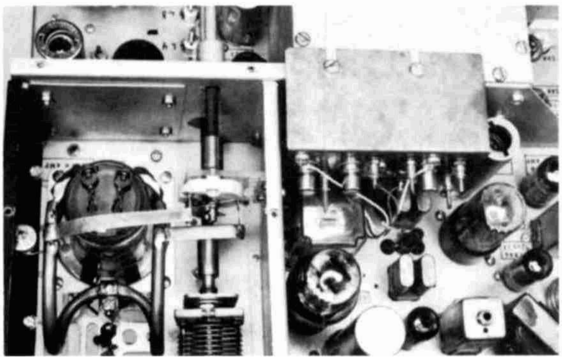
transmitter

The new two-meter transmitter circuit uses the original 12BY7 driver stage as the second mixer. L3 is relocated to form a 14-MHz bandpass coupler with L4. Spacing the two coils one inch apart provides the correct amount of coupling for tuning between 14.20 and 14.35 MHz.

Local oscillator injection is fed into the cathode of the 12BY7 across the original 100-ohm resistor. I removed the original bypass capacitor and connected a length of miniature coax between the local-oscillator output and the 12BY7 cathode. A renter-tapped coil in the 12BY7 plate circuit provides two-meter tuning and permits matching to the grid of the new 12HG7 two-meter drivrr.

A 130-MHz trap reduces the amount of local-oscillator signal feeding into the driver. The 2.2-pF capacitor is connected to one side of the final tank and is fed to the rf stage through the final compartment wall. No antenna switching is necessary; all switching from transmit to receive is accomplished hv the original bias method.

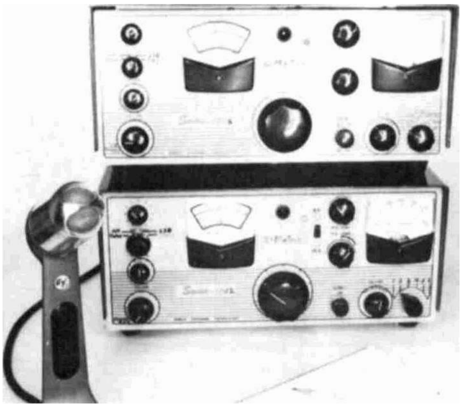
Top view of the **modified** Swan **120** showing the final-amplifier **compartment** and **new** convertor chassis. The plate-balancing capacitor **(C44)** is to the far **left** nert to the **plate** tine.



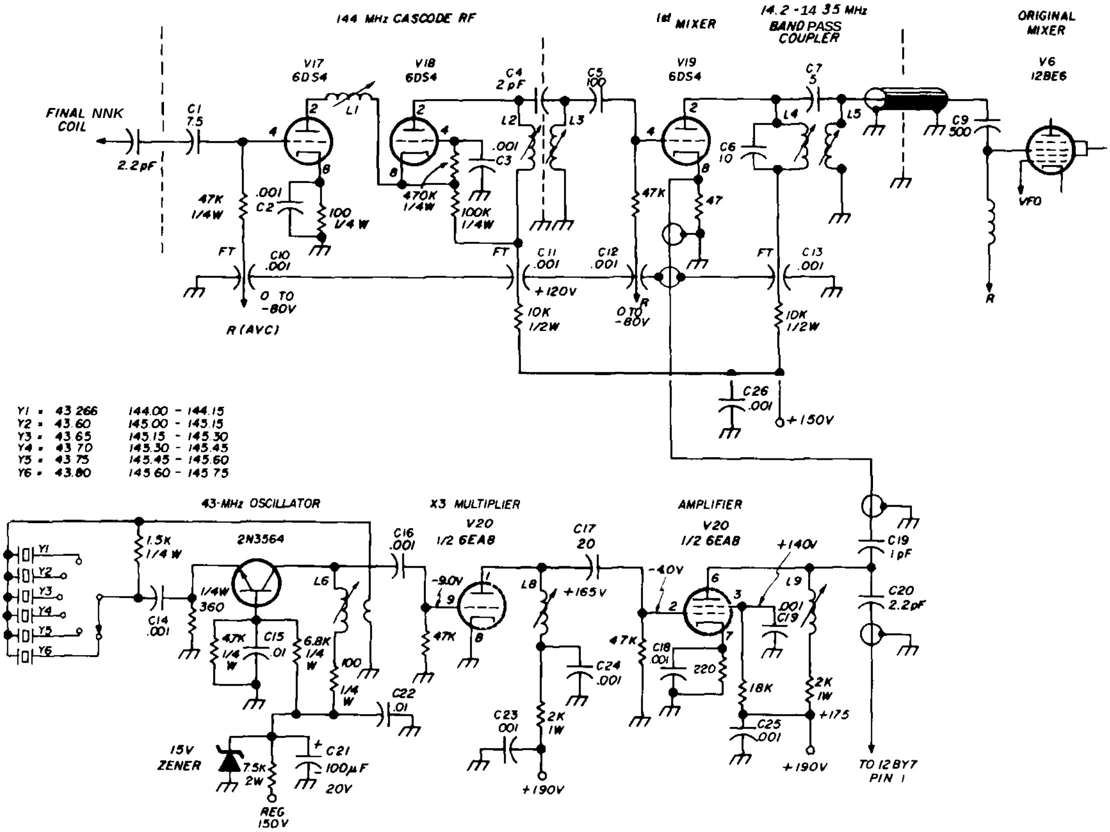
The 12HG7 is a rugged version of the 12BY7 with a higher plate-dissipation rating. When operated in Class AB_1, it provides plenty of output to drive the final. A series-tuned coil in the 12HG7 plate circuit is used with an adjustable capacitor to provide balancing. The 12HG7 tank circuit is tuned with a variable capacitor ganged to the mixer plate capacitor. This arrangement provides a single control for retuning when you change frequencies.

Thr two-meter output stage uses a 5894 operating in class AB_2. Static cathode current is adjusted to 50 mA. Ninetv watts PEP input produces over 40 watts peak output. Link coupling is used to couple the driving power from the 12HG7 into the 5894 grids. This is a simple and effective scheme for matching into the 5834 at 144 MHz.

Two converted single-band Swans: one for the sir. the other for two.



Although I didn't neutralize the 5894 at the beginning, I found that the final could be made to take off with a mismatched load. In addition, maximum output power didn't peak simultaneously with the plate current clip (a good indication of unstable conditions). Cross neutralization–a pair of insulated wires from the grids, crossed and extended up through the socket beside the 5894 plates–stabilizes everything very simplv.

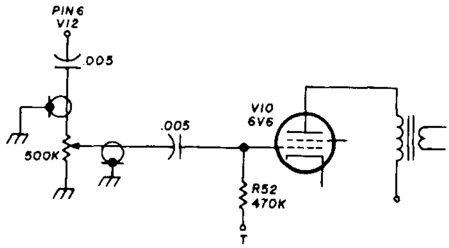


L1	**6** turns number **20.** White slug.	**L6**	14 turns number **26** enameled. White slug.
L2, L3	**4** turns number **20** spaced diameter of wire. White slug.	L7	**1-1/2** turns number **26** enameled. Wound on bottom of LG.
L4, **L5**	30 turns number 30 enameled. Red slug.	**L8, L9**	5 turns number **16** enameled. White slug.

fig. **1.** The receiver section of the two-meter Swan **120.** All coils are wound on **1/4"** diameter, slug-tuned ceramic forms. FT indicates feed-through capacitors. The voltages indicated are those found during normal operation.

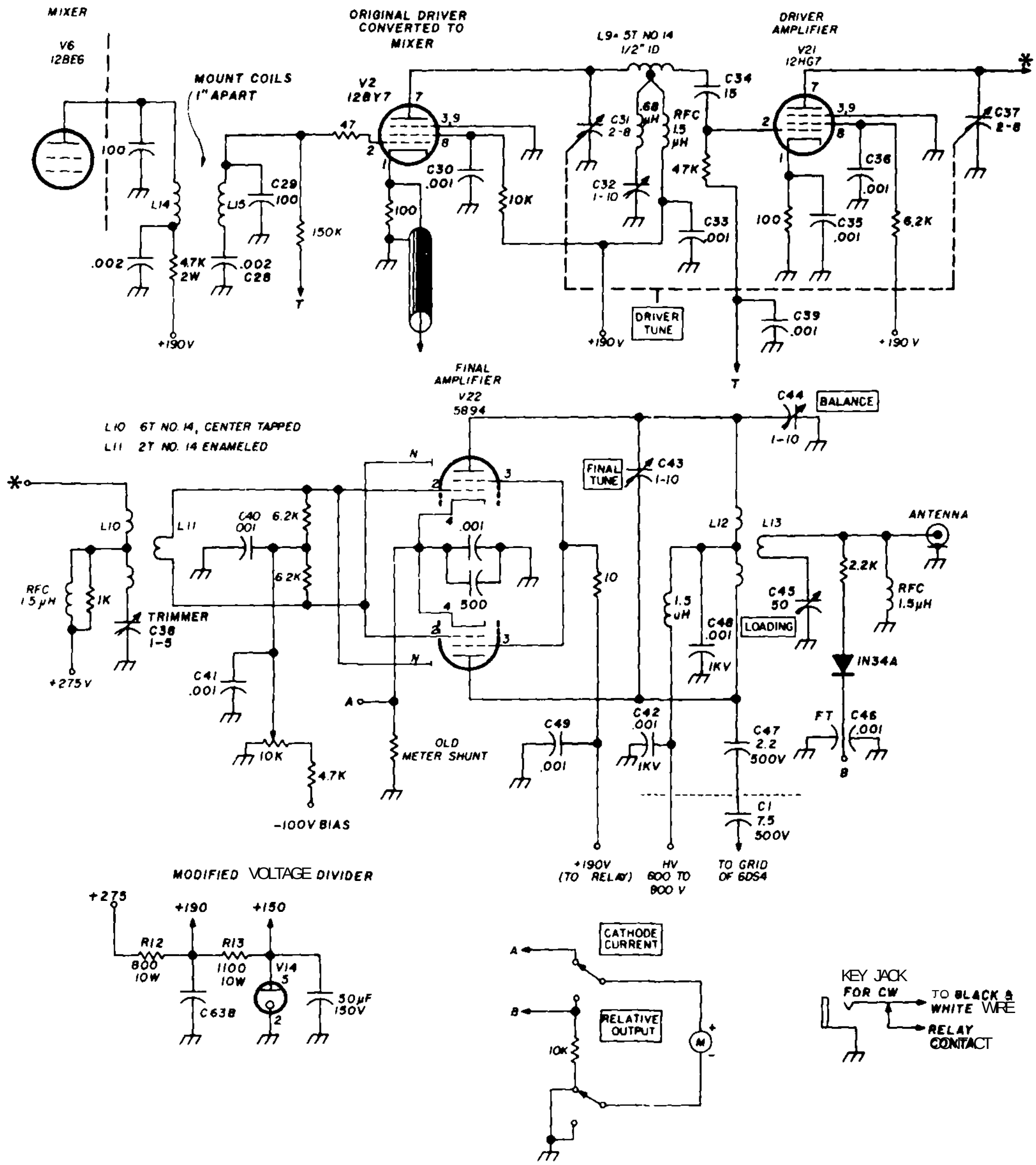
The 5894 plate tank uses a line made from 114-inch copper tubing. It is forrncd to fit into the original final compartment as shown in the photographs. A capacitor mounted on a plastic insulator tunes the tank circuit. This

fig. 2. New volume-control circuit for the modified Swan **120.**



"floating" capacitor is positioned so that an extension shaft comes out through the hole in the tront panel that held the original Swan plate-tuning control.

The final tank is balanced with the small piston variable on the opposite side of the final tank compartment. Power is coupled out by means of a rectangular loop of heavy wire connected to the original output jack The reactance of the loop is tuned out with a 50-pF variable (original Swan plate-tuning capacitor). An rf choke connected from the antenna jack lo ground provides a dc return tor the output monitor anti safety protection in case one of the coupling capacitors shorts out



C31, C37 1.5- to 10 **pF** air variable. Johnson **9M11** or 160-104.

C32, C44 **10-pF** piston capactor.

C43 **10-pF** air variable. Johnson **7J12.** Spread rotor plates for wider spacing.

fig. 3. Transmitter schematic. The rf chokes are **Ohmite Z-144's** (1.5-1.8 μ**H).** The voltages shown were measured at full drive on transmit with a **Simpson** 260 (through an rf choke). L14 and **L15** ara **the** original L4 and L3 respectively.

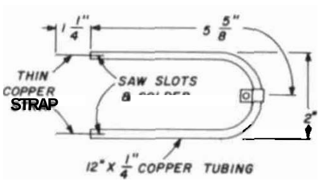
The original Swan cathode meter was modified by removing the 300-mA shunt from inside the case. Use care! The shunt is connected across a DPDT slide switch mounted on the front panel near the meter. With this arrangement, relative rf output or cathode current can be selected as desired.

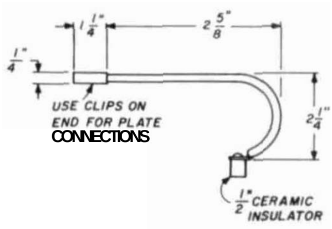
local-oscillator chain

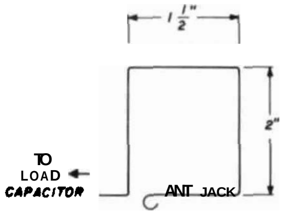
Local-oscillator injection is provided by a transistor crystal oscillator operating in the 43-MHz range. In this circuit, the crystal, operate in the series mode, but if other types of crystals are used, the VFO calibration can be readjusted slightly to correct for any frequency differences. A six-position switch provides six 150 kHz sections in the 144- to 148-MHz range. The dc power for the oscillator is supplied from the +150-volt supply by a dropping resistor and a 'IS-volt zener diode.

The triode section of a 6EA8 is used as a times-three multiplier, and the tetrode section, a 130-MHz buffer-amplifier. This tube provides plenty of injection for both transmit and receive; the screen resistor is selected for the desired injection level. Coupling is provided by a 1.0-pF capacitor to the receiving mixer and a 2.2-pF capacitor to the transmitting mixer. Small coaxial cables are used to feed the injection signal to each mixer cathode.

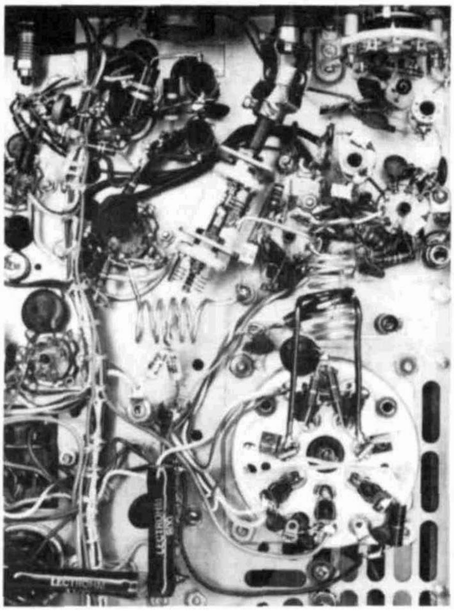
fig. 4. Construction of the final-amplifier inductor and output coupling link. First bend the tank lines as shown in A. and then bend as shown in B. Thin. 1-1/2" long strips are soldered on the ends for connection to the 5894 plate pins. The output link is made from number 12 bus wire.







Under the chassis. The new crystal oscillator and switch am in the upper right-hand comer. Tubes. from top to bottom. are 6EA8, 12HG7 and 5894.



chassis modifications

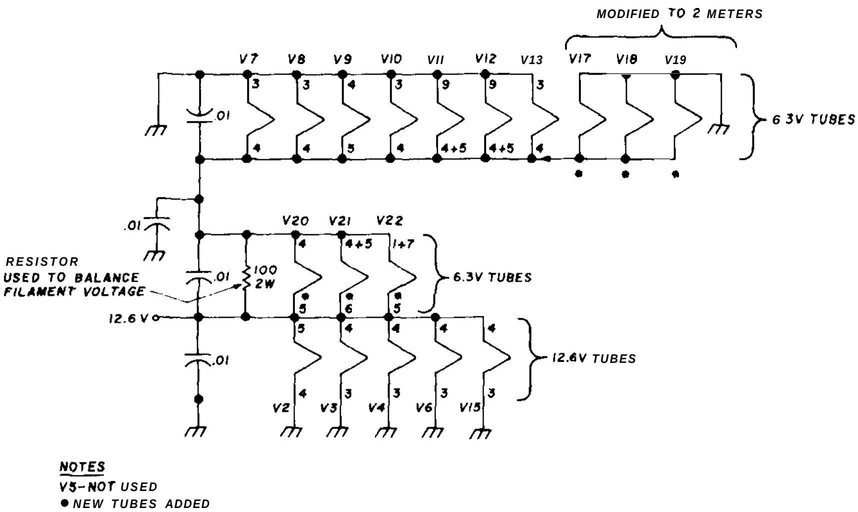
Before you can start the two-meter conversion, you have to take a few things out of the Swan 120:

1. Remove the front panel by removing knobs, control nuts, etc., and disconnect the meter and dial light.
2. Disconnect the wires going to the final amplifier tube socket and identify each one as you remove it. Pull the wires back through under the chassis.
3. Remove the final-compartment cover and shield, including the hack panel of the transceiver.
4. Disconnecl and remove all the parts in the final compartment.
5. Disconnect and remove the parts associated with the driver plate and final grid circuits.
6. Disconnect the parts from the 6BA6 rf amplifier (V5), and remove the tube.
7. Remove the switches on each side of the driver control; the on-off switch position is

used for the new rf gain control, and the tune switch is replaced by the new crystal switch. A new volume control with a built-in switch provides power-supply switching

is drilled so it fits snugly around the capacitor shaft; a 1/4-inch panel bearing is installed in the original driver tuning control hole on the front panel to prevent any shaft wobble. You

fig. 5. New filament circuit for the two-meter Swan.



8. Check the chassis to make sure that the areas to be modified are ready for drilling

Refer to the layout drawing and the photographs to locate all the holes that have to be drilled and punched. Careful layout and drilling is required to make sure all the parts fit properly, particularly the crystal sockets. After the holes are drilled, clean out the chips and filings before installing any new parts. The holes for mounting the terminal strips may be left until the major parts are in place.

assembly and wiring

Refer to the photos and drawings to determine the correct position for the major parts and check them temporarily before you wire them into place. First, install the crystal sockets next to the 6EA8 and 12HG7 sockets. The double capacitor used for mixer and driver tuning should be assembled to the capacitor bracket and installed temporarily while you're wiring the new circuits.

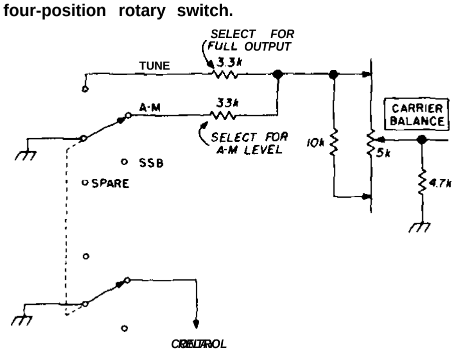
A flexible coupling and hollow shaft are connected to the driver capacitor. The shaft

can locate the various terminal strips by referring to the photos; use the smallest ones that will do the job.

The components associated with the transistor crystal oscillator are wired to a terminal strip right next to the crystal sockets and coil. Voltage for the crystal oscillator is provided by a resistor, zener diode and filter capacitor mounted on spare terminals on the voltage-regulator socket.

Use a shielded-type nine-pin socket for

fig. 6. The function switch for the two-meter Swan uses a double-pole, four-position rotary switch.



the 6EA8. The 12HG7 socket is a flush-mount tvpe installed from underneath the chassis; a shield can't be used because of the 12HG7's large envelope. A Johnson 122-101-100 socket is used for the 5894 power amplifier. Because of the size of the 5893, this submount-type socket must be used. It works very well and leaves clearance room at both the plate and grid ends of the tube. It also provides good ventilation.

The capacitors I used for coupling between stages were surplus ceramic units, but any good quality discs or micas will work quite well. All bypass capacitors are disc ceramics. Air-wound coils are made from number 12 bare wire formed around a 112-inch dowel.

The 1.5-pH chokes can be Ohmite Z144's or any small-size 1.2- to 2.0-pH chokes. The choke used in the 130-MHz trap has to be very close to the inductance indicated on the schematic. The best approach here is to wind a coil, temporarily connect it across the variable capacitor and grid dip it to 130 MHz before installing it permanently.

receiver section

Since the original 6BA6 rf amplifier is not used, it is removed along with the rest of the stage. A new front end is built on a brass subchassis and mounted on the VFO housing as shown in the photographs. A small coax cable is used to connect the output from the bandpass coupler to the grid of the 12BE6 mixer. The capacitance of this cable is used to tune coil L7.

The incoming 144-MHz signal is picked off the final plate tank. By using the original feed-through terminal, I eliminated any antenna relay and simplified the transmit1 receive switching operation.

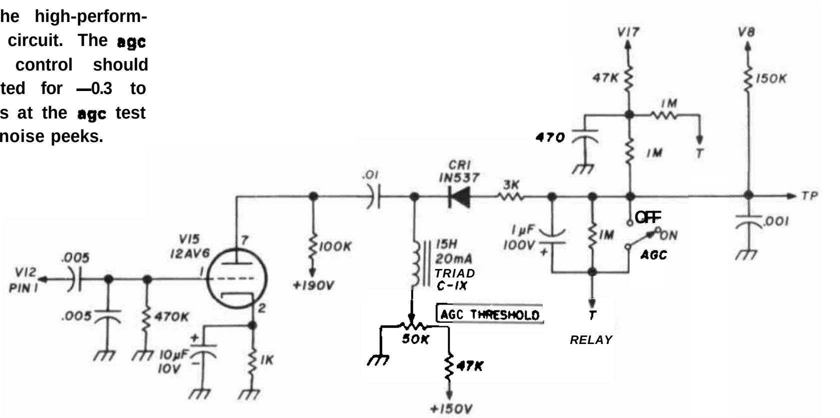
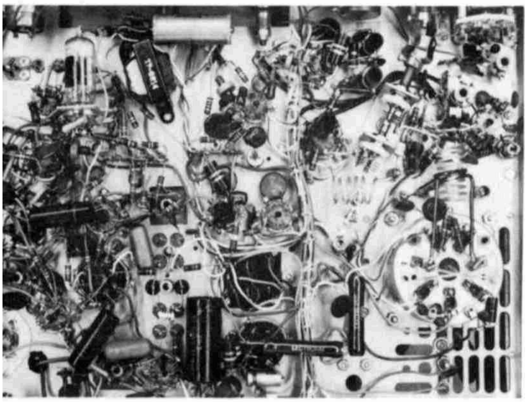


fig. 7. The high-performance agc circuit. The agc threshold control should be adjusted for —0.3 to —0.4 volts at the agc test point on noise peeks.



Below-chassis wiring. The 5894 final and 12HG7 driver are to the right; the tube in the upper left is the voltage regulator.

other changes

An agc circuit, similar to the six-meter model, was also used in this unit. A new audio gain control is also used. Several .001 capacitors are used to bypass points in the transceiver to prevent audio feedback caused by two-meter rf.

Bypass capacitors were connected between

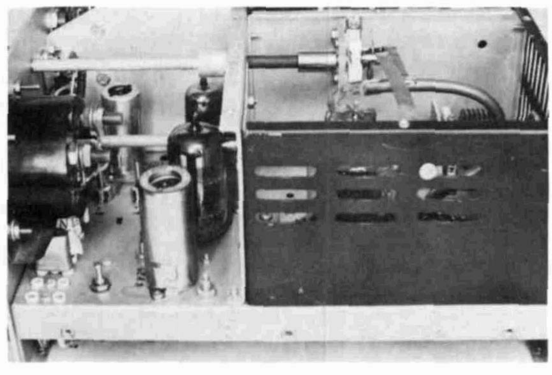
V11, pins 3 and 8, and the chassis. They were also used across several B+ points. The bypasses used on V11 are a must to prevent rf energy from being picked up by the long cathode leads; this causes rf feedback into the microphone amplifier.

adjustment and operation

Check all the receiver circuits to make sure thcy are properly wired. In addition, a quick check with an ohmmeter should he made before applying the power. Check all the coils to be sure thcy are adjusted close to their operating frequency. After power is applied, check the test points for correct voltages. Adjust the crystal oscillator, tripler and local-oscillator-chain amplifier for maximum output with the grid dipper in the diode position.

Adjust the new two-meter front end before mounting the chassis on the VFO housing. Adjust the bandpass coupler for flat response over a 150-kHz range. The two-meter coils have to cover a larger portion of two meters.

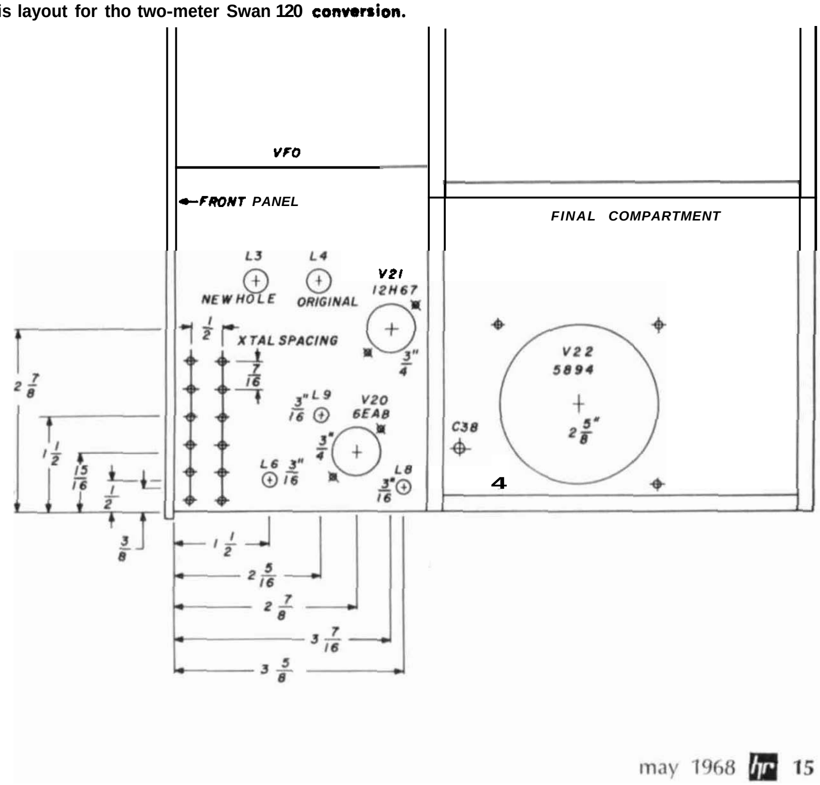
Side view showing the local-oscillator tube and crystal sockets. The 12HG7 driver, is just behlnd the 6EA8 oscillator.



transmitter adjustment

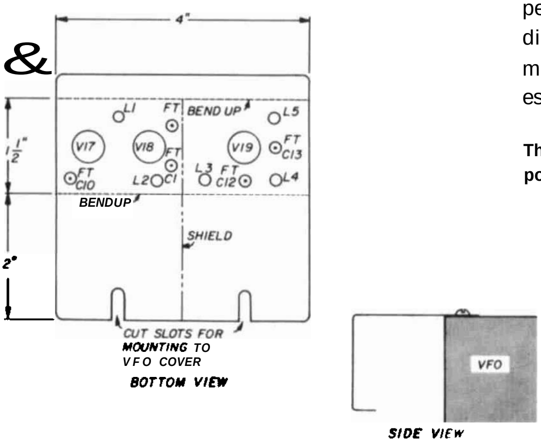
Preset all the coils to the desired frequency with a grid-dip meter, apply power, and after warmup, turn the transmitter on. Connect a 50-ohm load to the antenna connector and adjust the carrier-balance control for maximum injection. Peak all the controls for maximum output power. Next, with the VFO dial set in the center of its range, peak the 14-MHz bandpass coupler (L3 and L4) for maximum signal level. Adjust the driver-balancing capacitor for maximum output while simultaneously adjusting the driver control; this sets the tracking of the mixer- and driver-plate circuits.

fig. 8. Chassis layout for tho two-meter Swan 120 conversion.



Neutralization of the final stage is done after disconnecting the B+ and screen voltage from the 5894. With full drive applied, adjust the crossed wires next to the plates for lowest signal feedthrough at the antenna jack. These stubs may have to be cut and formed for best results (he sure to insulate them with spaghetti). You may have to re-neutralize after making the rest of the PA adjustments.

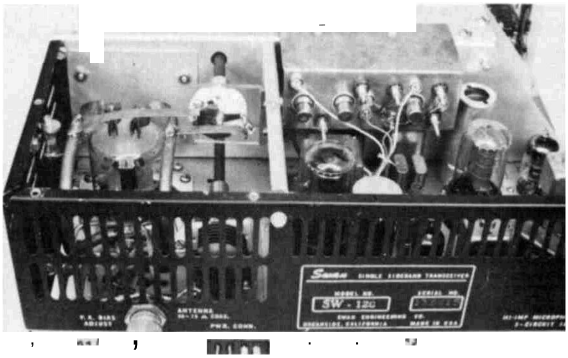
fig. 9. Construction of the chassis for the 144-MHz converter.



After the screen voltage and B+ have been re-connected, install the cover, insert carrier, and bend the antenna link for maximum output power. The plate-balancing capacitor must also be adjusted for highest output level. To obtain the correct drive levels, you may have to repeak all the driver circuits. The voltage readings marked on the schematic will help to determine if these circuits are working properly.

Adjust the 130-MHz trap with a grid dipper set in the cliode position; couple the dipper to the driver coil and tune for maximum, then adjust the trap capacitor for lowest output at 130 MHz.

The egc bias control and voltage test point are located on the rear deck.



parts list

1	7-pin socket (Johnson 122-101-100)
1	9-pin ceramic tube socket with shield
1	9-pin ceramic tube socket, flush mount
12"	1/4" copper tubing
15"	Miniature 50-ohm coax (RG-174/U)
8	1/4" slug-tuned ceramic coil forms
6	ceramic crystal sockets
1	DPDT slide switch
1	small single-pole, 6-position rotary switch
1	500k potentiometer with switch
1	12HG7 tube
3	6DS4 tubes
1	6EA8 tube
1	5894 tube
1	1N34A diode
1	2N3564 transistor
1	12- to 17-volt zener diode
1	100-μF, 20-V electrolytic capacitor

If you're interested in a particular section of the two-meter band, final adjustments can be made for optimum performance on that section. Be sure you recheck all the adjustments after the covers are installed.

The output meter shunt resistor is adjusted for center scale at 40 watts into a 50-ohm dummy load–10k ohms is about right.

operation

If you change your operating frequency by more than 150 kHz, you must repeak the transmitter controls. When you switch from 144 to 145 MHz, the power output will be low but you can still make contacts. Since the first tuned circuit in the receiver is also the final plate circuit, you'll notice a slight reduction in sensitivity unless the 5894 plate circuit is peaked up, but for local reception it isn't required.

After the driver and final circuits are tuned for maximum output, the microphone gain control should be adjusted so that voice peaks result in 35 to 50% of the maximum 5894 cathode current.

final comments

I spent a lot of time trying different circuits and layouts, and after a lot of experimentation, the conversion described here is the most simple and best performing I could obtain. The transistor crystal oscillator is very stable, even during mobile operation.

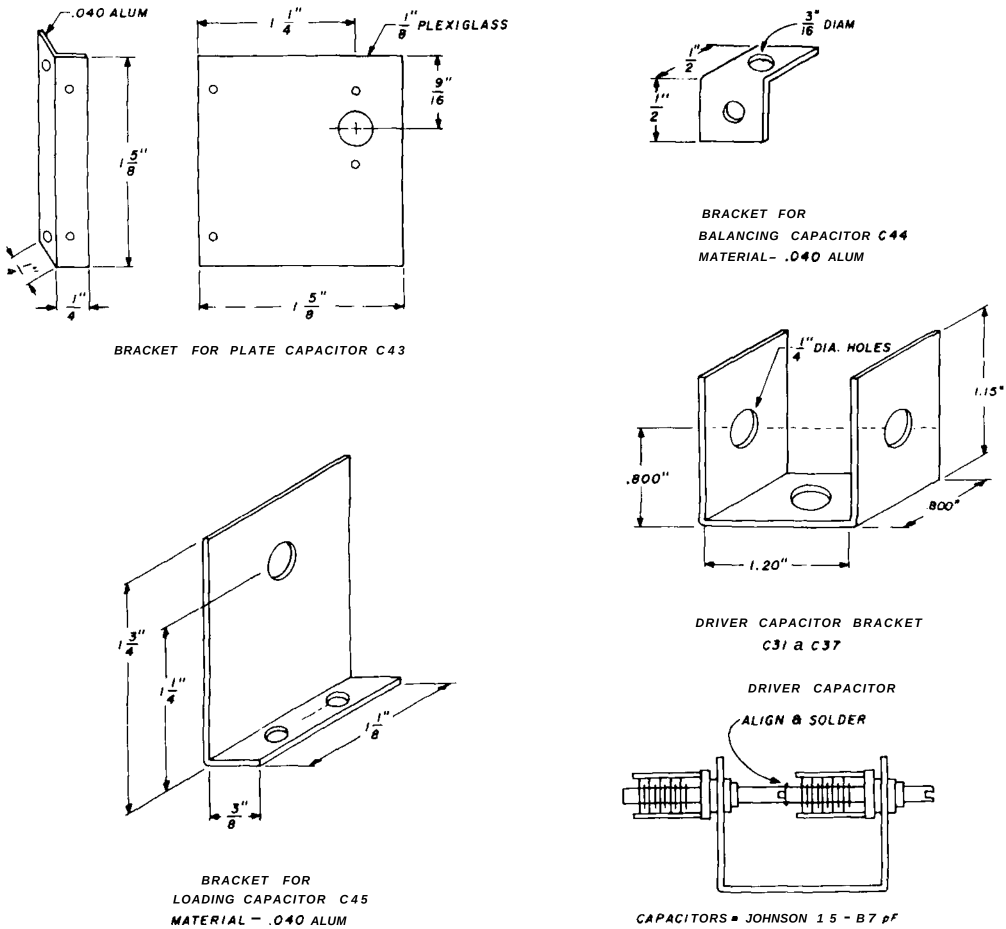


fig. 10. Construction of the various capacitor-mounting brackets and assembly of the driver capacitor from two Johnson 8.7 pF air variables.

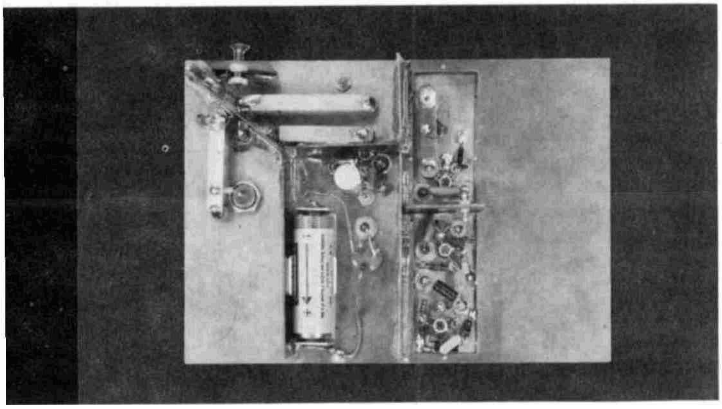
This conversion is only recommended for the more experienced home builder. It can be reproduced quite easily, but don't take any short cuts or make any parts substitutions. And, although this particular conversion was made to a Swan 120, Swan 175 and 140 transceivers could also be used if you make the necessary changes in the mixing circuits.

Since I completed this conversion, I have worked stations up to 200 miles away on two meters. I can change frequencies from one section of two to another, tune up and be ready to transmit in a matter ot seconds. We need more ssb on our vhf bands, and this conversion is a good way to do it for 144.

references

1. D Crowell, K6RIL, "Converting the Swan 120 to 6 Meters," 73, June 1967, pp 28-34.

ham radio



Inside the low-noise 432-MHz converter.

low-noise
432-MHz fet converter

A high-performance converter for 432 MHz with excellent cross-modulation and gain characteristics

Bob Kolb, WA6SXC, 1300 W. Oak Street, Fullerton, California 92633

The availability of moderately-priced field-effect transistors with low noise figures in the vhf region has prompted a number of amateurs to design preamps for existing converters. The complete 432-MHz converter described here is a high-performance unit consisting of a single-stage rf amplifier, a multiplier chain and a mixer with a 28-MHz i-f.

The neutralized rf amplifier has 15-dB gain as shown by the scope trace. The 3-dB bandwidth is 10 MHz and the 1-dB bandwidth, 5 MHz. The rf amplifier uses the flat-pack 2N4417. The mixer, a 2N4416, operates at pinch-off, so it doesn't provide any conversion gain. However, this mode of operation offers excellent cross-modulation characteristics which are sorely needed in southern California where radar interference is a problem.

The 2N4416 is the same device as the 2N4417 rf amplifier, but it's mounted in the familiar TO-18 can. The multiplier uses a 101-MHz CR56/A crystal available from JAN Crystals* followed by two doublers to pro-

* JAN Crystals, 2400 Crystal Drive, Fort Myers, Florida 33901

duce 5 mW at 404 MHz. All bipolar transistors in the oscillator/multiplier circuits are 2N918's.

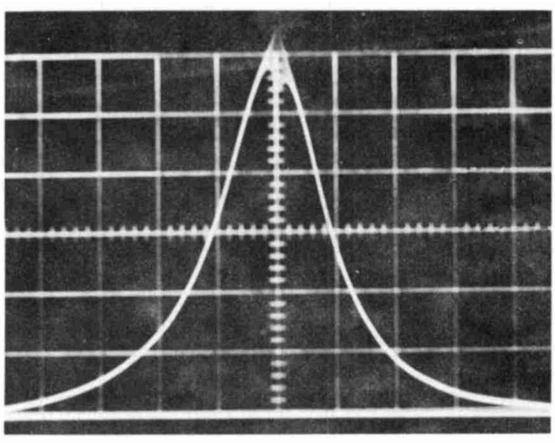
The net gain of the converter is only 12 dB. However, this is sufficient to overcome the noise figure of most high-quality amateur-band receivers. The signal-to-noise ratio is significantly better than my old two-stage 2N2857 preamp which measured 4.5-dB noise figure the last time I checked.

construction details

The circuit is quite simple and easy to construct. The chassis is made from 1116-inch epoxy-fiberglass printed-circuit board. The dividers are 0.9-inch high and the entire converter can be mounted in a l-inch deep aluminum chassis. The board size shown in the drawings is 5 by 7 inches with a 2-1/4 by 5-inch section left for an internal power supply. The local-oscillator multiplier chain was constructed on a 1-1/4 by 5-inch board and tested independently.

This modular construction has a number of advantages, some of which will be noted in the tune-up section. For one thing, each

Gain-bandwidth characteristic of the rf amplifier. Tha horizontal scale is 5 MHz per division: vertical scale is approximately 1 dB per division above the center graticule, which represents 12-dB gain.



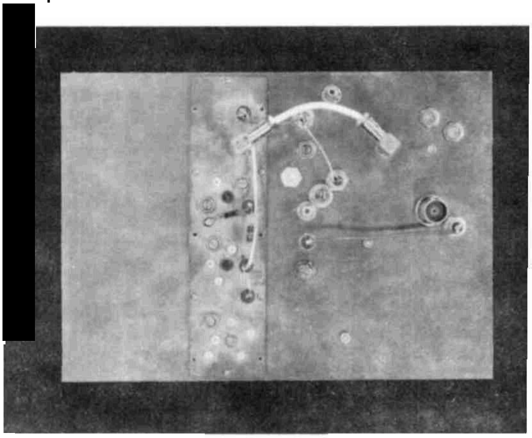
portion of the converter can be individually checked for performance. Furthermore, the local oscillator can be easily replaced by a different frequency unit if you want to change i-f's.

Use a large alligator clip to heat sink the Johanson capacitors when soldering to them. They're constructed with 570°-F solder and are easily damaged by medium-power soldering irons.

L1, L3, L4 and L5 are made from 0.020-inch thick by .250-inch wide copper shim stock. Brass or beryllium copper may also be used. Each of these inductors is mounted 0.3 inches above the chassis and supported by the low-pass filters (LPF's) and capacitors as shown in fig. 3. The mounting of L5 is shown in fig. 3B.

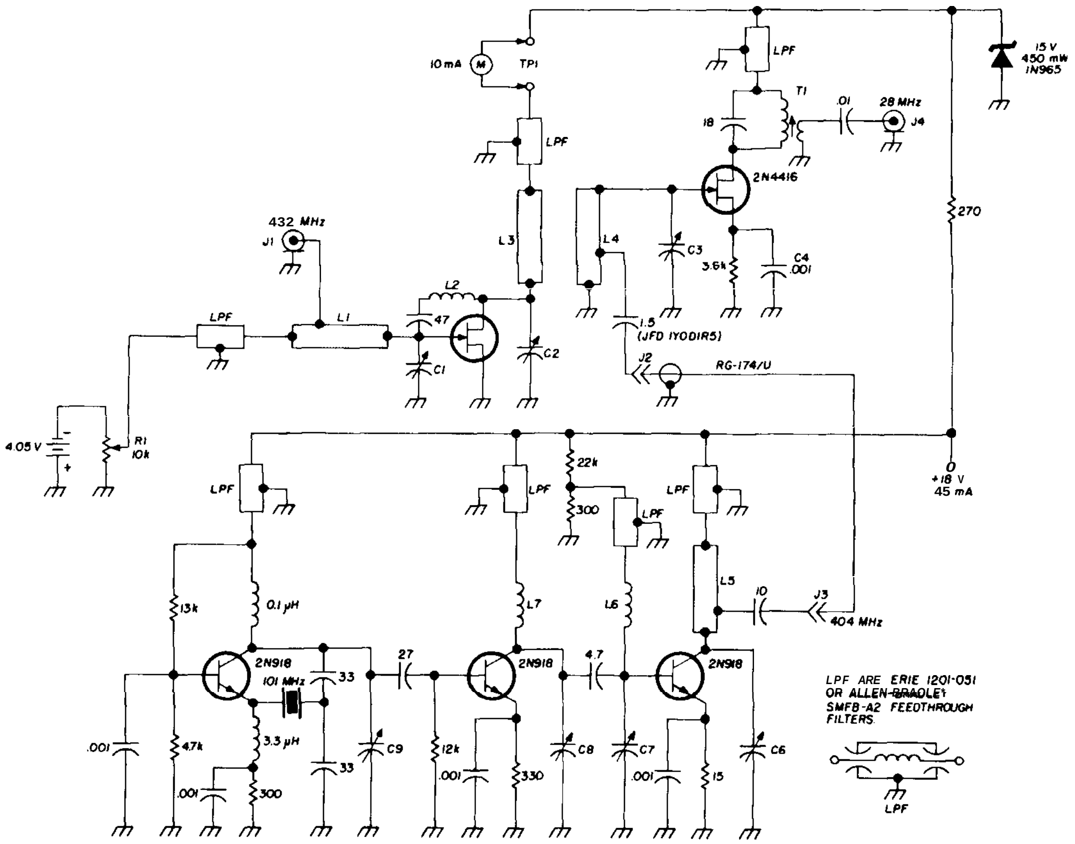
The 2N4417 rf amplifier is mounted with

Top view of the 432-MHz converter.



the gate lead soldered to the top of C1, the drain lead to the top of C2 and the source lead to the grounded standoff terminal 0.3-inch high. Don't bend the source lead and solder it to the shield. The ceramic-glass seal won't take the thermo-mechanical strain; this advice comes from costly experience. **Fig.** 4A shows this mounting from the source side looking along L1. **Fig.** 5 shows the FET and transistor connections.

The manufacturer of the FET recommends the use of a Miller 4403 0.99- to 1.5-pH variable inductor for L2, but don't mount the form on a grounded surface. The slug must not be grounded; if it is, neutralization is next to impossible! Furthermore, L2 should be mounted on the drain side of the shield and the neutralizing capacitor on the gate side. Drill a small hole through the shield and remove a 118-inch circle of copper around it to



BT1	Mallory T-133R
All variable capacitors are 0.8-10 **pF** piston (Johanson JMC 2954)	
C4	1000 **pF** (Allen Bradley **SB4A** 102W **stand-off**)
J1, J2, J3	Miniature coaxial connectors (Amphenol Subminax 27-800 or **Selectro Conhex** 3102)
LPF	Feedthrough filters (Allen Bradley **SMFB-**A2 or Erie 1201-051)
L1	**1.25"** x **.25"** x **.020"** copper strap tapped **1/4"** from cold end
L2	6 turns number 28 close wound on a **1/4"** **teflon** core with **1/2"-long number-6** brass screw
L3	**2"** x **.250"** x **.020"** copper strap
L4	1.3" x **.250"** x **.020"** copper strap tapped at **1/2"** from cold end (see fig. **3B)**
L5	**1"** x **.250"** x **.020"** copper strap tapped **.375"** from cold end
L6, L7	3 turns number 24 wound on **100k, 1/4W** resistor
T1	**.51-.85** μ**H** (Vanguard **8923-1)**

fig. 1. Schematic of the low-noise 432-MHz converter. This converter **uses** a low-noise **2N4417** rf amplifier and **2N4416** mixer.

feed the lead from L2 to the neutralizing capacitor.

L2 is mounted on a 0.9- by 0.75-inch piece of printed-circuit board 1116-inch thick. The copper is stripped from both sides of the board 1/4-inch up from bottom as shown in **fig. 4B.** The bottom of the board is then soldered to the main chassis. The easiest way to remove the copper from the board is to score an outline of the area to be removed with an

X-acto knife, flow solder over that area, and use the knife to lift the copper while it's still hot.

Note that the base divider is mounted on top of the local-oscillator board. Feedthrough capacitors may be substituted for the low-pass or feedthrough filters called out, but look out for self-resonances at 432 MHz. If feedthrough capacitors are used, connections in the dc power lines should be made through 10-pH rf chokes. BNC connectors can be used for all coaxial jacks, but J2 must be repositioned midway between L3 and L4, and J3 moved to the right of L5.

noticed that most solid-state amateur converters suffer from insufficient local-oscillator injection; on the other hand, too much local-oscillator power degrades noise figure. Note that minimum sensitivity can vary as much as two S-units with a 20-dB change in local-oscillator power.

The generator I used had only +10 dBm available. I would expect the curve to rise slightly as the local-oscillator power increases. At this point noise figure would he severely impaired. Local-oscillator injection into the source circuit usually requires even more power. By tapping local-oscillator pow-

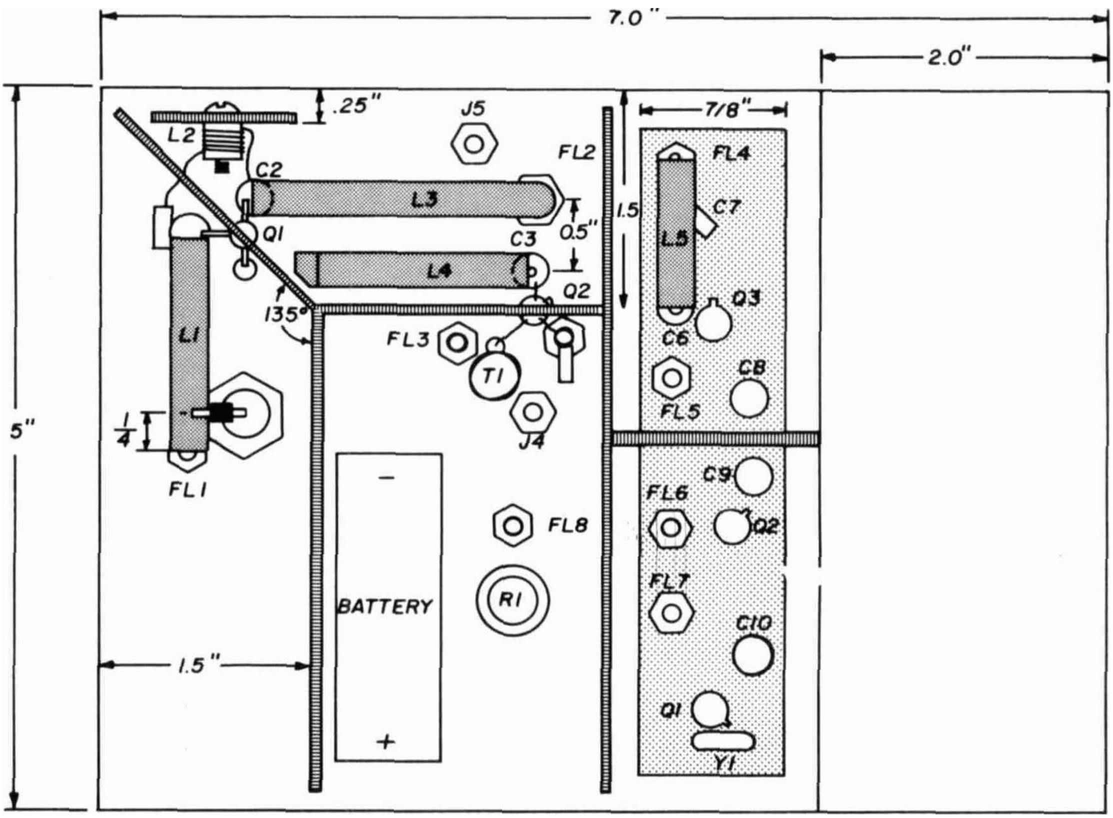


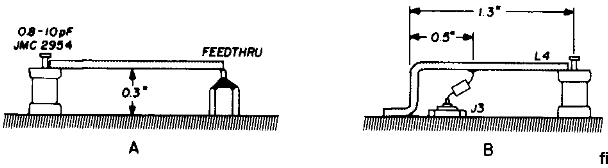
fig. 2. Layout of tho low-noise 432-MHz converter. Double-sided copper-clad board is used for the chassis as well as the shields.

The tangential sensitivity (minimum discernible signal) versus local-oscillator injection power is shown in fig. 6. This data was taken with a 50-ohm source applied to J2 and the power level varied in I-dB steps. I have er into L4 you get a good match to the local-oscillator circuit while isolation between the local-oscillator and i-f is enhanced by the portion of L4 that is in series between the local-oscillator input and the i-f output.

tune-up procedure

You can tune up the local oscillator with any rf-sensing device such as a detector probe for a VTVM or grid-dip meter in the diode mode. The choice of a 101 MHz crystal eliminates the possibility of tuning the local oscillator to the wrong harmonic or the develop-

fig. 3. Inductors L1, L3 and L5 are shown in A. L4 construction is illustrated in B.



ment of birdies due to nearby harmonics. It can be tuned up on the bench using a 51-ohm resistor as a load.

the rf amplifier

To adjust the rf amplifier, connect an antenna or 50-ohm load to J1. With plus 15 V applied, adjust R1 for 5 mA of drain current (I_D) measured at the dc input to L3. The voltage on L1 should be approximately —0.5 V. Remove the connection at J1 and note any change in current. If the drain current changes, L2 must be adjusted. Apply a signal to J1 and monitor the output of the rf amplifier at J2 (local oscillator removed). Alternately adjust C1, C2, C3 and L2 for maximum signal with no change in current (I_{DQ1}) while J1 is open circuited. When ,I remains steady with J1 open, and the amplifier is tuned for maximum signal, it is properly neutralized.

Do not attempt to neutralize the rf amplifier by adjusting L2 for minimum signal feedthrough without B+ applied. The B+ changes the junction capacitance of the FET and detunes the circuit. Small changes in B+ don't shift the bandpass significantly, but in areas where commercial power voltages vary considerably, it's a good idea to regulate the B+ to the rf stage with a 15-V zener diode.

Fig. 7 shows the method I used to determine converter sensitivity. Noise figure was calculated at 3.1 dB based on a 250 kHz i-f bandwidth for the HP417. **Minimum Discernible Signal** with 50% modulation was —125 dBm or approximately 0.1 µV. I attempted to measure MDS with the HP608C at 0 dBm with precision step attenuators in series with the input. However, this approach was abandoned because the unshielded converter picked up radiation from the generator when the attenuators were increased beyond —65 dBm.

fig. 4. Method of installing the 2N4417 rf amplifier (A). The support for L2 is shown in B.

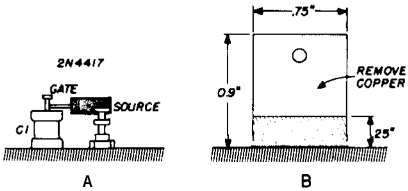


fig. 5. Transistor base diagrams.

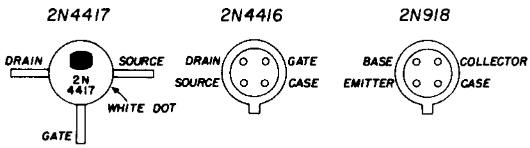
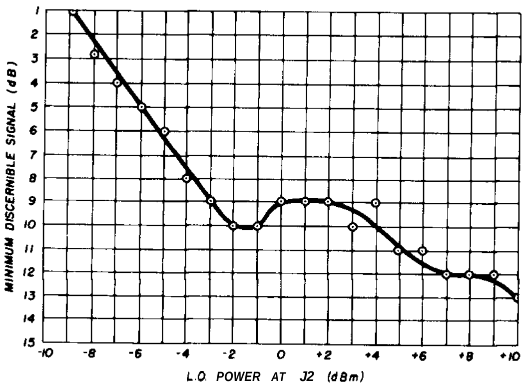
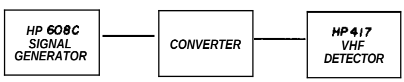
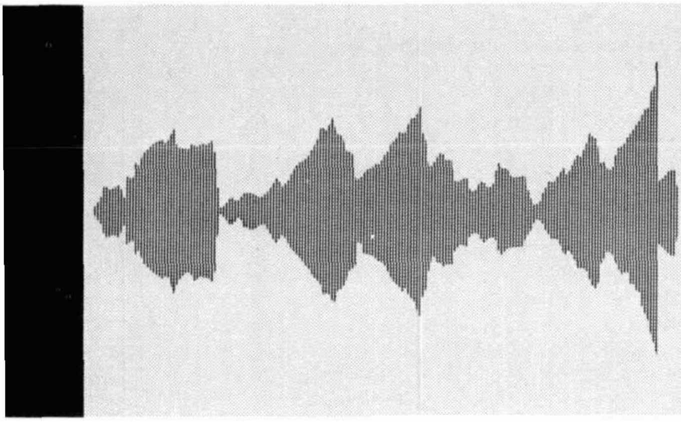


fig. 6. Minimum discernible signal vs. local-oscillator power for the 432-MHz converter.





ham radio



generating ssb signals with suppressed carriers

The inside story of the balanced modulator

Forest H. Belt, 119 Baker Avenue, South Plainfield, New Jersey 07080

The first step in forming a single-sideband signal is the generation of what is called a double-sideband suppressed-carrier or **dsbsc** signal. Some form of filter then removes one sideband completely. It takes a special kind of modulator to create the sidebands and eliminate the carrier, and the one used most is called a **balanced modulator.**

balanced modulators

A modulator is nothing more than a special mixer for mixing the voice signal with the carrier. With ordinary amplitude modulation, the mixing creates sideband frequencies equal to the sums of and differences between the voice modulation and the carrier–and, of course, both original signals remain in the output, too. The circuit that produces a suppressed-carrier signal must form the sidebands exactly the same as an ordinary a-m modulator and yet eliminate, as completely as possible, the carrier against which the voice modulation beats to create the sidebands.

First, then, to make it easy to understand the principles of carrier-suppressed modulation, let me explain a way to feed a carrier signal into a modulator circuit in such a way that the circuit is controlled by it and yet the carrier itself does not appear in the output. **Fig. 1** shows how this works.

Look at **fig. 1A** first. In this arrangement, the rf signal is fed to V1 and V2 in parallel. In other words, when the grid of one is on its positive rf half-cycle, so is the grid of the other. As is usual in amplifiers, each tube inverts the signal.

Look what happens when the outputs of the two tubes are connected in push-pull. The output of V1 is a negative half-cycle, and is applied in one direction through the transformer. The output of V2 is also negative-going, but it is applied in the opposite direction through the transformer.

The result is cancellation in the transformer of the effects of either signal. If the amplification of the two tubes is exactly equal, and the transformer itself is well balanced, there is total cancellation of whatever signal is applied to the grids of the two tubes.

Next, look at **fig. 1B.** If a signal is fed to this stage in push-pull, the halt-cycles of rf sine wave drive one tube in one direction and the other tube in another. However, if the outputs of both tubes are connected in parallel, the positive excursion of one always cancels the negative excursion of the other.

For example, suppose the signal at the grid of VI is on its positive excursron; the signal at the grid of V2, then, is on its negative excursion. As usual, each tube inverts the signal. The output of V1 is a negative half-cycle, and the output of V2 is a positive half-cycle. Since these are mixed in the same load, they cancel each other. Thus, push-pull in and parallel out also means cancellation of the input signal.

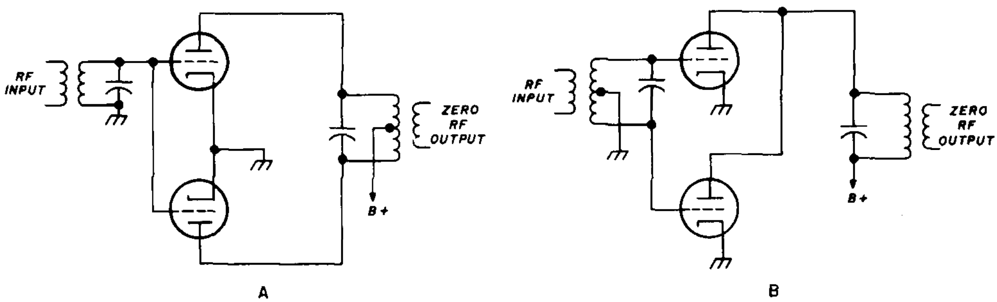
If you're wondering why any purpose is served by feeding an rf signal into the stage only to have it canceled in the output, think about what happens whenever the tubes become unbalanced. Imagine that V1 in **fig. 1B** amplifies less while V2 amplifies more. One of the signals that appears in the parallel output will dominate the other, because it is stronger.

If the situation were reversed, with V2 amplifying more than V1, the unbalance woulcl create a dominance of the opposite polarity. If something switches the amplification of the two tubes alternately up and down quite rapidly, the output signal varies back and forth at the same rate. The result is an output that is a rapidly fluctuating rf signal of first one polarity and then another

Consider the same action in **fig. 1A.** With V1 conducting more than V2, the opposite signal components in the transformer are no longer equal, and a certain amount of rf output is coupled to the secondary If V2 conducts more than V1, the unbalance is in the opposite direction Again, if something switches this unbalance back and forth between the two tubes at a rapid rate, the output varies at that same rate

Fig 2A shows a convenient method of varying the gain of the two tubes What you see is the same circuit you saw in **fig. 1A,** but with a speech input *transformer added in* **push-pull.** It is eacy to see that the push-pull speech signal can swing the amplification of the two tubes hark and forth at an audio rate

fig. 1. Cancelling the carrier signal: input in parallel, output push-pull (A); input push-pull, output parallel (B).



The tubes become alternately unbalanced at an audio rate, and rf shows up in the output—swinging back and forth from positive-going to negative-going output at the same rate. The effect is that the rf and speech signals are "mixed" and sidebands are created, yet the rf carrier signal itself does not appear in the output. The instant the speech modulation is removed, there is zero output from this balanced modulator.

Fig. **2B** shows the principle applied to the circuit of fig. **1B.** In this one, as you can guess if you now see the underlying principle of the balanced modulator, the speech signal is sent to both tubes *in* parallel. Since the tubes conduct alternately as far as rf is concerned, varying the gain of both tubes with the speech signal results in the same kind of amplification unbalance at an audio rate that was described in fig. **2A.**

The result in the output is exactly the same. The carrier itself does not appear in the output; in fact, with no speech input, there is no output. When there is speech modulation, however, the output consists solely of sidebands created by mixing the speech signal with the rf carrier.

The pattern of operation here should be clear. In the balanced modulator, the rf signal is applied in one mode and coupled out in the other. That is, if rf is fed into the stage in parallel, it is taken out in push-pull. If fed in in push-pull, it is taken out in parallel. This is true of all balanced modulators; that's why they suppress the rf carrier. The speech signal, on the other hand, is always applied in the same mode as the output is removed.

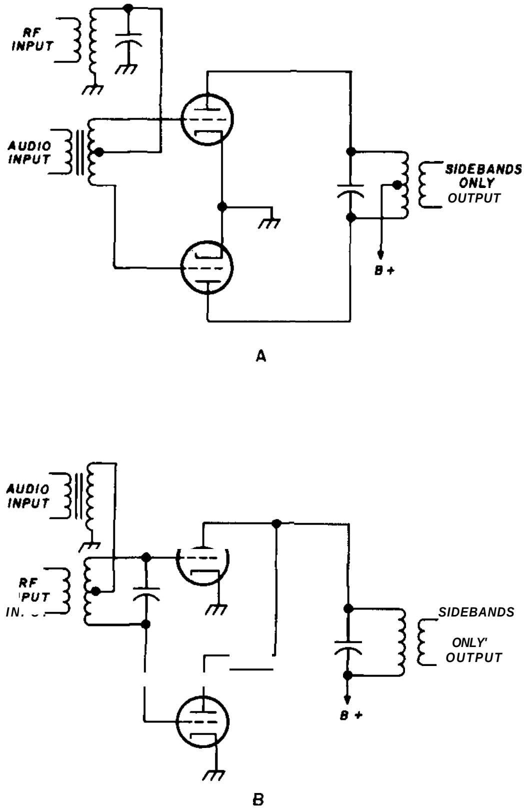
Balance is important. If either tube becomes slightly unbalanced, the carrier is then amplified constantly, even though slightly, by that tube. Many balanced modulators include a balancing adjustment which is set for minimum rf output with zero audio modulation.

Now that you are aware of the principles involved, let's examine some actual circuits. With your new understanding, you'll find the circuits themselves are easy to figure out.

diode balanced modulators

The simplest and least expensive balanced modulators generally use semiconductor diodes. They seem to be more stable than tube-type balanced modulators, and are not prone to change characteristics over periods of time. Well designed diode balanced modulators provide about 40 dB of carrier suppression–more than tube types do (with the exception of the special beam-deflection-tube balanced modulator, which will be explained

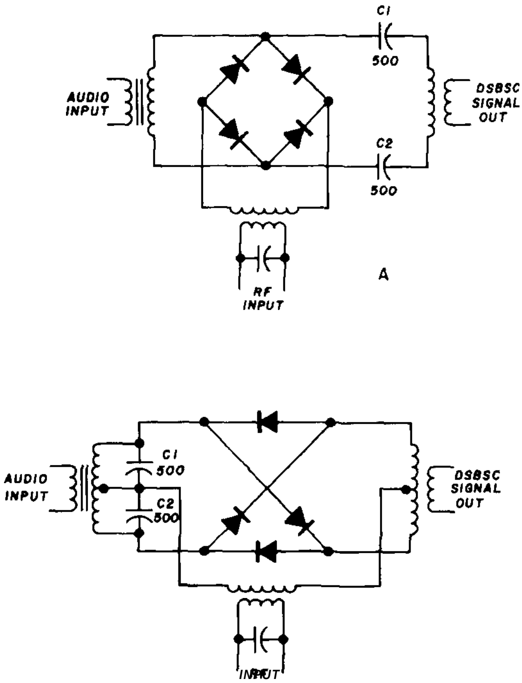
fig. 2. Adding audio input for moduation: rf parallel, audio push-pull (A), rf push-pull, audio parallel (B).



later). That means the power in the sidebands, at 100% modulation, will be at least 40 dB stronger than whatever carrier power slips through.

Two easy-to-understand diode balanced modulators are shown in fig. 3. At **3A,** you see the bridge type. In it, the rf and the speech signals are mixed in a four-diode bridge. Notice that, effectively, the principle of balanced modulators is adhered to. The speech signal is applied to the stage in the same mode in which the output signal is taken out.

fig. 3. Basic diode balanced modulators: bridge (A), ring (B).



The rf signal, on the other hand, is applied to the "balanced" corners of the bridge. One end of the rf input transformer is connected to the cathode of two of the diodes, and the other end is connected to the anodes of the other two. The result, of course, is that the rf signal is "shorted" to ground by the diodes except when there is speech modulation to unbalance their conduction. When that happens, the output becomes a double-sideband suppressed-carrier signal.

ring modulators

An improved version is shown in **fig. 3B.** It, too, use four diodes–in a circuit called a **ring.** Better sideband signals are produced in the ring modulator than in the bridge-type. Again, the speech input is in the same mode as the output. The rf signal is fed into transformer center taps, so it is balanced with respect to the output.

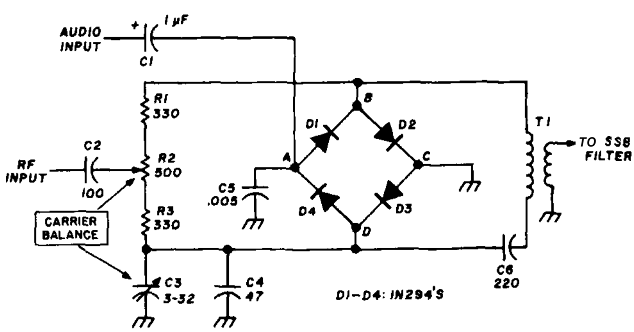
Efficiency in the ring modulator is high, and the four diodes should be carefully matched. One way to check them is with an ohmmeter, by measuring their forward and backward resistances. All four should match within 2%–even better, if possible. If they aren't matched, a certain amount of the rf carrier will slip through. Furthermore, the sidebands themselves will be unbalanced, which will create distortion when you try to recover them at the receiver.

The purpose of the 500-pF capacitors in both modulators is to keep audio and rf separated except in the mixer diodes. The capacitors pass rf energy quite easily, but present a fairly high impedance to the speech signal. In **fig. 3A,** the sideband output transformer would act pretty much as a short circuit for the speech signals; instead, the capacitors keep them out and they are forced to go into the bridge.

In **fig. 3B,** the two capacitors merely assure a low-impedance path for the rf signal in both directions to the ring circuit. The capacitors hardly affect the speech input signal at all, because of their low value.

About the most popular balanced modulator for ham equipment is the one shown in **fig. 4.** It is a variation on the ring circuit already described. Don't be confused by the

fig. 4. The most popular diode modulator used in ham transmitters.



way it's drawn, because it isn't a bridge. You can tell it's a ring by the fact that the diodes all are in series with one another; in a bridge, there are always two cathodes together and two anodes together.

This circuit has some other peculiarities, because it is designed to eliminate the expensive input transformers. Furthermore, both the speech and the rf signals are fed into the ring diode circuit from stages in which one side is grounded. The thing to do, to understand this particular balanced modulator best, is to analyze the action on the rf carrier alone first, and then study the effects of unbalance created by the speech input signal.

Next consider what happens when a speech signal is applied at point A. The capacitor there has little effect on the speech frequencies. Therefore, the path to ground for positive half-cycles of the speech signal is through D1 and D2. For negative half-cycles, it is through D3 and D4. The speech signal thus "turns on" these diode pairs alternately, at the speech-signal rate. You can see that the speech signal determines which diodes are conducting and which not conducting during a given half-cycle period.

As an example of the effect: when the speech signal is causing D1 and D2 to conduct, it has reverse-biased D3 and D4. Consider positive excursions of the rf signal (many rf excursions occur during each audio half-cycle) Their only possible paths to ground at that Instant are through D2, which is made conductive by both the rf and the speech signal, and through D4 which is made conductive only by the rf signal Obviously, the greater rf current flows through R1 and D2 Much less flows in R3 and D4, because conduction in that diode is opposed by the speech-signal excursion For negative excursions of the rf signal, the path aided by the speech-signal excursion is through R1 and D1; the R3-D3 path is opposed because D3 is still reverse biased

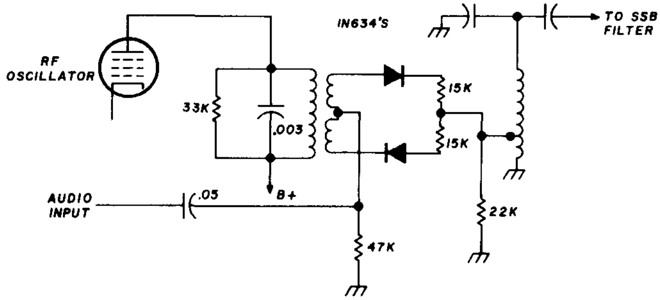


fig. 5. Two-diode balanced modulator is one form of the ring-type.

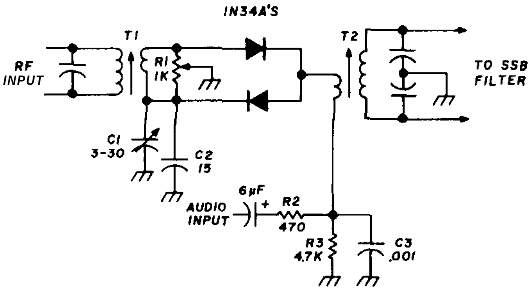
The first thing to notice is that two corners of the ring are grounded as far as rf is concerned. Capacitor C5 keeps point A at rf ground; point C is grounded directly. That being the case, the rf signal is applied to the ring effectively in parallel. It goes in both directions through R1 and R3, trom balancing potentiometer R2. When it reaches B and D it splits up, with both segments being shunted to ground through whichever diode happens to be conducting on that particular half-cycle or excursion.

The important thing is that, because of the way the diodes are connected in the ring, the rf signal is behaving **inside the ring circuit** almost as if it were in push-pull. From point B, it goes through D1 on one half-cycle, then through D2 on the next, seeking ground. From point D, it alternates going through D3 and D4. With zero input from the speech circuit, the rf signal is continually shunted to ground on both excursions by one or another of the diodes. The result is that no rf reaches the primary of transformer T1 and consequently there is no carrier output.

If you carry through the reasoning for both positive and negative speech-signal excursions, you'll see that the path for all rf signals is through R1 on positive half-cycles of the speech signal, and through R3 on negative ones This unbalancing means that some of the rf is not canceled and causes rf current to flow in T1–first predominantly in one direction, then in the other Since, during modulation, this unbalance is varying at the speech-signal frequencies, the output is a pair of sidebands resulting from mixing the carrier and the speech frequencies; there is no carrier.

Capactitor C6 serves the same purpose it serves in the other circuits–to make sure only sideband signals reach T1; its value is such that it virtually blocks speech signals. C3 and C4 are balancing capacitors that make up for any stray capacitance in the stage; C3 is adjusted for minimum carrier output with zero modulation.

fig. 6. Simple two-diode balanced modulator is also ring-type.



two-diode modulators

An exceptionally simple variation of the diode-ring balanced modulator is used in one transmitter. If you examine its circuit carefully (fig. 5), you'll see that all the requisites of a balanced modulator are there. The diodes are in series with each other, as in a ring circuit. The rf signal is fed to the modulator circuit in push-pull by the secondary of the input transformer, and the resulting sidebands are taken out in parallel via a tap between the two balancing resistors. The speech signal is fed in parallel, being applied to the center tap between the two secondary windings of the input transformer.

With no modulation, each excursion of rf is applied to the diodes, but only the ones that make the top of the transformer secondary positive and the bottom negative can make the diodes conduct. Current then flows through the two balanced resistors. However, the output is taken off **between** the two resistors, so the voltages across the two resistors are in opposite phase with respect to ground, and they cancel. The result: no output. During the other excursion, there is no output because the diode5 aren't even conducting.

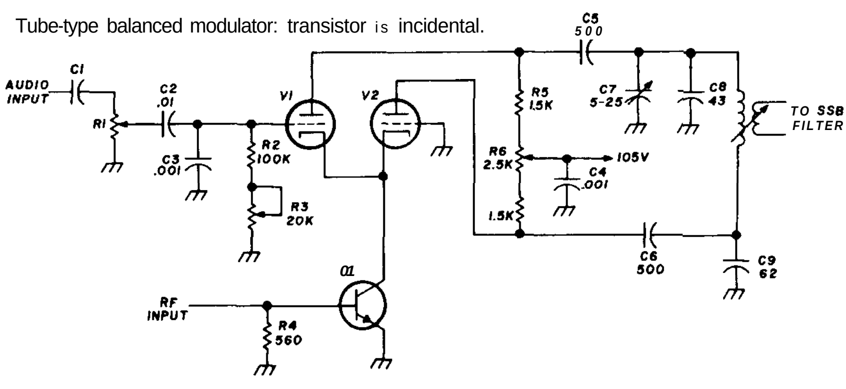
As the speech signal switches first one and then the other diode "on," the rf signal that is trying to flow in both diodes is either opposed or aided. The constantly shifting unbalance at the speech frequencies produces the sidebands at the junction of the two resistors, with the carrier suppressed.

Don't let the tap on the output coil confuse you. This one is strictly for matching the low impedance of this modulator to the higher impedance of the filler that follows.

There are other relatively simple two-diode ring circuits. One that has been popular in some home-brew rigs is shown in **fig.** 6. You don't need much explanation of this one; you can figure out its operation from your knowledge of this type of balanced modulator.

The rf is applied to the two-diode ring in push-pull. R1 can be adjusted for a "center-tap" ground that allows the speech to be fed in at a tap between the two diodes–therefore in parallel. The output is taken in parallel at the same point, through a coil which couples

fig. 7. Tube-type balanced modulator: transistor is incidental.



the sideband signals through a tuned secondary to the sideband filter.

C1 and C2 are the capacitance-equalizing part of the carrier-balancing network. Sometimes C1 is moved to the opposite end of the coil. R1 is of course the carrier-balancing potentiometer.

tube-type balanced modulators

In **fig.** 7 is a tube-type balanced modulator used in at least one commercial ham transmitter. A transistor is used, too, but it is primarily an impedance-matching input device. The balanced modulator has the usual configuration. The rf is fed to the stage in parallel, to the two cathodes, and the output is taken out in push-pull.

R5, R6, and R7 make up a balancing network to equalize conduction of the two tubes. The speech signal must be applied in push-pull, which is accomplished by grounding the grid of one tube and feeding the signal to the grid of the other. This is, in effect, push-pull.

Operation of this circuit is very much like the one in **fig 2A.** The two 500-pF capacitors keep speech signals out of the output circuit. C7, **C8,** and C9 are the capacitance-balancing capacitor,. C7, the adjustable one, rnay be at either end of the output transformer winding, depending on which position does the best job of suppressing the carrier. Capacitor C4 keeps the carrier-balance control slider at rf ground, providing the rf "center-tap" ground necessary to make the output circuit push-pull.

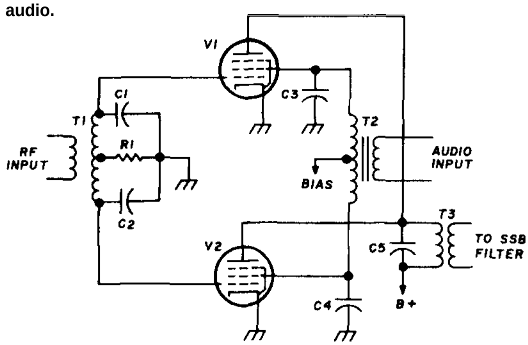
Pentodes may he used in place of triodes for this circuit. When pentodes are used, the rf signals may be fed to the control grids while the speech signal is fed to the screens. Speech and rf are in push-pull, and the outputs are paralleled. **Fig.** 8 shows an example of this particular hookup. The schematic is simplified to show merely how it works; the system is seldom used in commercial ham transmitters.

deflected-beam-tube modulator

RCA has a tube, the 7360, that is designed especially tor balanced rnodulators and balanced detectors. Its cathode gives off electrons that form a beam which travels between two deflection plates toward two output plates. The schematic diagram detailed in fig. **9** looks pretty complicated, but it isn't if you keep the fundamentals you have already learned about balanced modulator operation in mind.

One big difference in this particular circuit is that it is self-oscillating. The carrier is generated internally. An external oscillator can be used, but there is little need, unless a separate oscillator is already part of another circuit. This one is a crystal-controlled Colpitts.

fig. 8. Pentode system uses screen grids to accept the audio.



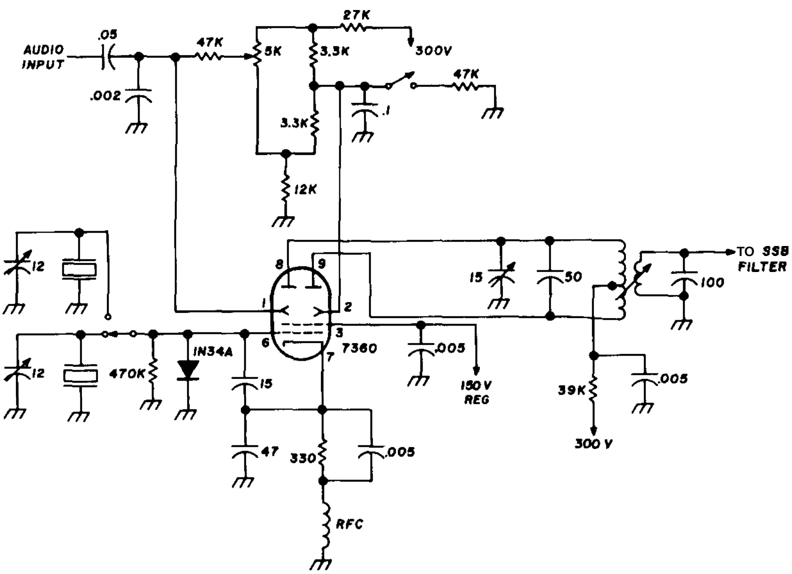
The structure of the tube is such that the internally generated rf signal modulates the beam, but the beam doesn't strike either output plate. It goes right between them. So, you have that old familiar balanced-modulator characteristic: no rf-carrier output signal. You can call this type of carrier-signal injection parallel, since it has the capability of reaching either plate, as you will shortly see. The output, of course, is in push-pull.

The speech signal is applied to the deflection plates of the tube, effectively in push-pull. One of the deflection plates is at ground potential for audio signals—the 0.1-μF capacitor does that job. The speech signal is fed to the other deflection plate.

The dc voltages on the two deflection plates must be equal when there is no speech input, so the rf beam misses both plates. That is done by a voltage-divider network across the 300-volt supply line, and the 5k potentiometer. Then, when speech modulation is applied to the deflection plates, the beam is pulled hack and forth so it strikes the output plates, first one and then the other, at an audio rate. The speech signal thus produces an output that contains only the sidebands.

The beam-deflection balanced modulator has several advantages. It is easy to adjust and doesn't require a lot of tricky balance adjustment—just one potentiometer. A regulated supply is a good idea lor the screen grid, which acts as a sort of plate for the oscillator. The circuit is capable of at least 60 dB of carrier suppression. This is beyond thr most stringent requirements.

fig. 9. Beam-deflection balanced modulator; tube is a RCA 7360.



There you have the story of balanced modulators. Keep in mind the basics: the output mode is the opposite of the rf input mode, so the output will be devoid of the carrier frequency. The speech signal is fed to the stage so it can control gain, switch diodes, or deflect a beam that already contains the rf. The result of this mixing of signals is an output that contains the sidebands produced by the mixing process, without the carrier.

ham radio

two-year novice licenses

A number of **rumors** have been circulating around to the effect that Novice licensees who held a one-year Novice license which expired 22 November or thereafter, would receive a one-year extension to their license term. This is not true. According to the Federal Communications Commission in Washington, D. C., no extensions will be granted to Novice licensees because of the new two-year Novice-license term.

If you know any Novices whose licenses expired on or after 22 November 1967, check to make sure that they are not under the mistaken impression that their license has been extended for one year. If you hear other amateurs spreading this rumor, please tell them that there are no one-year extensions to Novice licenses. If they have any questions, have them check with their local FCC office.

ham radio

folded mini-monopole antenna

Space problems?
The mini-monopole antenna
offers high performance
and efficiency
in a small package

William I. Orr, W6SAI, 48 Campbel Lane Menlo Prk, Calfornia

Compact, low-frequency antennas are characterized by low input resistance, high reactance, Inherently high Q and low radiation efficiency. In addition to this melancholy catalog of features, the compact antenna requires a matching network to compensate for antenna reactance and to transform the input resistance to a value that is compatible with the transmitter. Unhappily, in most cases, the matching network Inevitably introduces additional circuit losses. As a result, the performance of a compact low-frequency antenna system often leaves much to be desired.

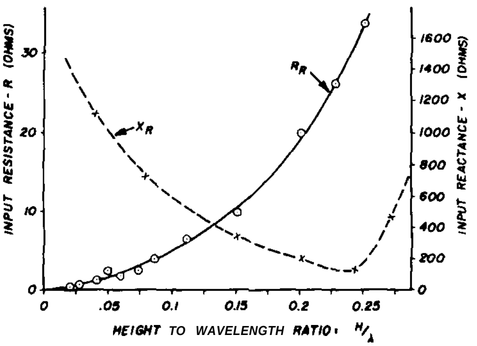
the low-frequency whip antenna

The size of an efficient low-frequency antenna presents a problem to the amateur whose backyard is small and hemmed in by telephone and power lines. In many cases, the only clear direction for an 80- or 160-meter antenna is straight up. Mobile operation poses much the same problem, with the additional handicap that vertical antenna height is limited to eight feet or so. Mobile antennas and practical back-yard whip antennas of reasonable height are not noted for their efficiency; moreover, ground losses in the back-yard antenna and body losses in the automobile installation further compound the problem.

The input resistance of a simple whip antenna drops rapidly as the length is decreased below one-quarter wavelength **(fig. 1)**. An ordinary quarter-wave whip has an input resistance in the neighborhood of 32 ohms or so, with minimal reactance.

At a length of 0.2 wavelength, the input resistance has dropped to about 20 ohms, and the reactance has risen to -200 ohms. When the whip length is further reduced to 0.1 wavelength (about 25 feet at 80 meters), the input resistance is only 4 ohms and the reactance has sky-rocketed to about -600 ohms. Worse still, the eight-foot mobile whip has an input resistance of less than 0.4 ohm at 40 meters and less than 0.1 ohm at 80 meters.

fig. 1. Input impedance of whip antennas.



In each case, the input reactance is extremely high–over -1000 ohms. Since the input resistance is extremely low, a considerable amount of current must flow in the antenna if any power is to be dissipated as radiation. In order to accomplish this, the reactance must be cancelled out by adding equivalent positive reactance (inductive) in the form of a loading coil to tune the antenna system to resonance.

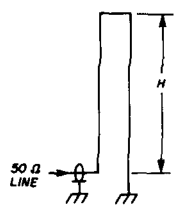
The tuning network or loading coil adds the losses of the coil to that of the antenna. Even with a high-Q coil (say, 300 or so), the rf resistance of the coil is from 20 to 200 times the input resistance of the short whip, and, as a result, the greatest percentage of transmitter power heats up the coil instead of increasing the signal at the receiver of an eager listener. Antenna system efficiencies on the order of 0.5 to 2 percent are common in mobile installations, and you can expect efficiencies of 5 to 10 percent in short, loaded, back-yard whip antennas.

the folded mini-monopole antenna

The folded mini-monopole antenna was developed in an attempt to obtain a portion of the impedance transformation within the antenna itself, eliminating the high losses of the customary matching network.[1,2]

The folded monopole is merely one-half of a folded dipole working against an image ground system as shown in **fig.** 2. Folded mini-monopoles less than a quarter-wavelength in height have been tested in the low-frequency region of 1.8 to 7 MHz in an attempt to devise a relatively efficient antenna system of reasonable size. The input impedance of a folded mini-monopole was measured on a vhf model and plotted in **fig.** 3. Both the resistive and reactive components of the input impedance vary sharply as the height-to-wavelength ratio is varied. The curves show several interesting input combinations which exist at discrete heights.

fig. 2. The folded monopole antenna.

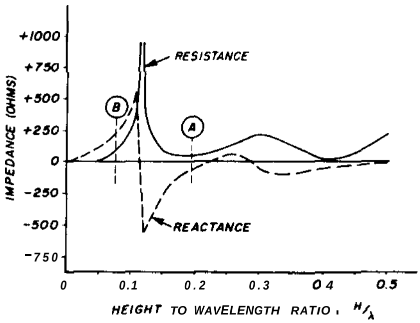


At a mini-monopole height of 0.19 wavelength (point A, **fig. 3**), the input resistance is about 50 ohms, with a reactance of -50 ohms. This is a reasonable value to feed directly with a 50-ohm line with no matching system; the SWR would be about 1.5:1. Not bad at all for a simple antenna! At a center frequency of 3.8 MHz, it would require a mini-monopole height of about 48 feet.

An even more compact mini-monopole may be constructed 0.075 wavelength high (point B, **fig.** 3). The input resistance of this antenna is about 50 ohms, with a reactance of +150 ohms. The positive reactance may be easily cancelled out by adding a series tuning capacitor. Happily, the tuning capacitor has much less resistive loss than even the best loading coil or matching network. For a center frequency of 3.8 MHz, the 0.075-wavelength mini-monopole is only about 19-feet tall.

This version of the monopole is very attractive for 160-meter operation. For a lower-band segment (1.8-1.9 MHz), the antenna height is 41 feet, while for the upper band (1.9-2.0 MHz), the antenna is 37 feet high. For 40-meter mobile operation, moreover, antenna height is about 10 feet–well within the capability of a modest installation since no center loading coil is required!

fig. 3. Input impedance for folded mini-monopole antenna.



the 0.19-wave mini-monopole

Dimensions for the 0.19-wavelength mini-monopole antenna for 160, 80 and 40 meters are given in **fig.** 4. While the antenna itself presents a reasonable match to a 50-ohm line, you can accurately match the transmission line with an SWR meter by adding a very small, adjustable loading coil.

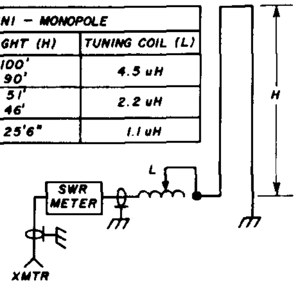
While a height of 90 to 100 feet may be prohibitive for 160 meters, you can see from **fig.** 3 that the antenna may be shorter than the dimensions given in **fig.** 4, down to 0.15 wavelength or so, without appreciably affecting results. It is merely necessary to add inductance to the tuning coil (L) to compensate for the restricted height. A minimum height of 80 feet is suggested for 160-meter operation, while half this would be suitable for 80-meters.

the 0.075 wave mini-monopole

The 0.075 wavelength mini-monopole is a very interesting antenna; the dimensions for this mini-wonder are given in **fig. 5.** The chart

fig. 4. The 0.19 wavelength mini-monopole.

0.19λ MINI – MONOPOLE		
FREQ.(mc/s)	HEIGHT (H)	TUNING COIL (L)
1.825 1.925	100' 90'	4.5 uH
3.5 4.0	51' 46'	2.2 uH
7.0	25'6"	1.1 uH



in **fig. 3** shows that this antenna presents an inductive component at the input terminals which can be tuned out with a series capacitance. For 160 meters, the over-all height is 40 feet or less, while for 80 meters the height is only 20 feet, or less. Since antenna height is only 10 feet on 40 meters, mobile operation is feasible. The series capacitor should be a transmitting type for moderate power levels, since the voltage across it is quite high.

building a mini-monopole

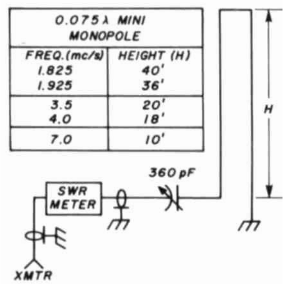
As with any antenna system, you must take care in building a mini-monopole or the advantages of the antenna will be offset by sloppy and inaccurate construction. For the 80- and 160-meter versions, the antenna may be built of aluminum tubing. Telescoping lengths of l-inch and 7/8-inch tubing may be used if it's well guyed, or larger tubing can be used in a self-supporting arrangement.

The parallel tubes are strapped together at the top with a low resistance strap; a length of l-inch wide aluminum strap is best. The spacing between the whips is not critical and may be from ten to sixteen inches or so. It appears that too-close spacing will inhibit antenna operation and spacings of a few inches have provided puzzling results. Good, low-resistance connections between the sections of the antenna are mandatory, and liberal use of an aluminum anti-oxidizing compound such as Penetrox-A* is recommended.

A good ground is required with both versions of the mini-monopole antenna. Multiple ground rods may be used, or a simple counterpoise system is suggested. A satisfactory counterpoise may be made up of two or more quarter-wavelength **insulated** wires running close to, but not on, the ground. Insulated hookup wire is satisfactory for this purpose and the clever amateur can disguise them by running them through bushes, along fences, or around the framing of the house.

The antenna is tuned by adjusting the series capacitor or inductor for minimum SWR on the transmission line. If you're an experimenter, you may want to change the height of the mini-monopole to drop the SWR to near-zero at your pet operating frequency. The rest of us lazy hams, like myself, will probably forego this instructive experience and work the antenna as is.

fig. 5. The 0.075 wavelength **antenna.**



* Penetrox-A is made by the Rurndy Company and distributed through the General Electric Supply Company.

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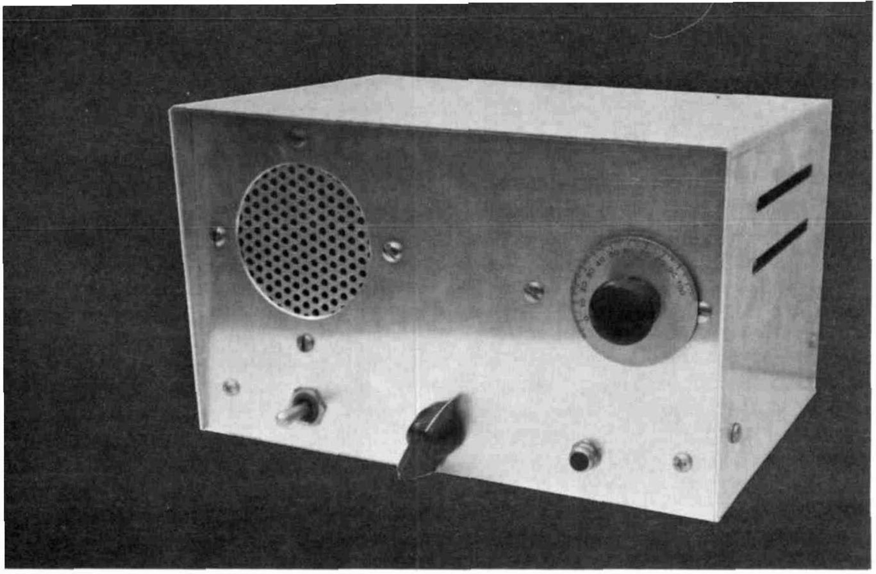
1. Giddis, "A Technique for Matching an Adjustable Folded Dipole," *IEEE Transactions* on Antennas and Propagation, May 1964.
2. King, "The Theory of Linear Antennas," Harvard University Press, Cambridge, Massachusetts, p. 341.

ham radio



162.55-MHz

ESSA weather receiver

using IC's and FET's

A tunable vhf receiver of modern design using IC's, FET's and a crystal filter

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

In addition to low-frequency broadcasts of weather information for aircraft by the FAA, there are Environmental Sciences Service Administration (ESSA) broadcasts of surface weather information on 162.55 MHz (FM). Since these surface-weather broadcasts are primarily intended for ships and boats they are only available near navigable waterways. However, from the list of the ESSA stations given in table 1 you can see that most of the country is pretty well covered. The broadcasts from these stations are continuous–by means of an automatic tape.

You can build a rather simple receiver for these weather stations by using some of the new semiconductor devices on the market. Integrated circuits cut wiring complexity, and field effect transistors and a crystal filter add performance features that can't be equaled—even by the multi-tube "kluges" in FM mobile service.

circuit

A block diagram of the 162.55 MHz receiver is shown in fig. 1. Note the use of FETs in the front end; the rf amplifier and mixer are TIS34's. The use of FETs in these two blocks of the receiver yields exceptional cross-modulation immunity. Another TIS34 is used for the local oscillator, but for a different reason–its inherent temperature stability over a bipolar transistor circuit.

The system bandwidth is established immediately after the mixer by a surplus 10.7-MHz crystal filter with a 40-kHz passband. Since the ESSA broadcasts are using FM with ± 15 kHz deviation, the 40 kHz passband of the crystal filter provides the required 30 kHz, plus 5 kHz "elbow room" on each side of the channel.

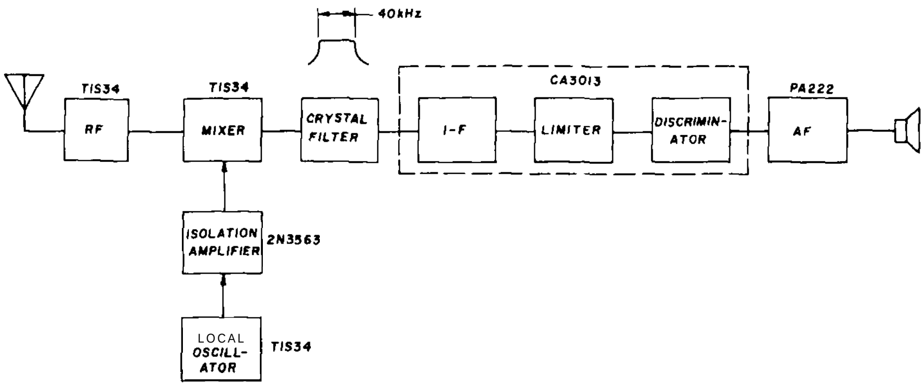
The i-f, limiter and discriminator consist of a single integrated circuit, an RCA CA3013. This ten-lead, TO-5 can IC has 75 dB i-f amplification and the discriminator diodes–all for $2.75! The CA3013 is apparently the same IC that is being used as the sound i-f and detector in many TV sets. All you need to complete the detector is a discriminator transformer and a few peripheral resistors and capacitors. If you have trouble finding a CA3013, it could probahly be replaced by the CA3014 or SK3022.

table 1. List of 162.55-MHz ESSA weather stations

Atlantic City, N. I.	KHB-38
Boston, Mass.	KHB-35
Bridgeport, Conn.	KHB-45
Charleston, S. C.	KHB-29
Chicago, Ill.	KWO-39
Corpus Christi, Texas	KHB-41
Galveston, Texas	KHB-40
Honolulu, Hawaii	KHA-99
Jacksonville, Fla.	KHB-39
Kansas City, Mo.	KID-77
Lake Charles, La.	KHB-42
Los Angeles, Calif.	KWO-37
Miami, Fla.	KHB-34
New Orleans, La.	KHB-43
New York, N. Y.	KWO-35
Norfolk, Va.	KHB-37
San Francisco, Calif.	KHB-49
Tampa, Fla.	KHB-32
Washington, D. C.	KHB-36

The receiver power supply is conventional, with a full-wave silicon rectifier, capacitive input and a dual regulator. Three voltages are supplied for this mixture of active devices: unregulated +18 and regulated +6 and +14 volts.

fig. 1. Block diagram of the 162.55 MHz ESSA weather receiver.



The audio section is similarly taken care of by an IC, the General Electric PA222. This chip, in a modified dual-inline package (DIP), provides 70 dB of open-loop gain and can put-out up to a **watt** of audio. This amazing little af package fits into a standard 14-pin DIP socket, even thuugh it only has 8 pins. The ninth (and largest) lead on the PA222 is a heat sink tab. It is soldered to a copper tab on the chassis.

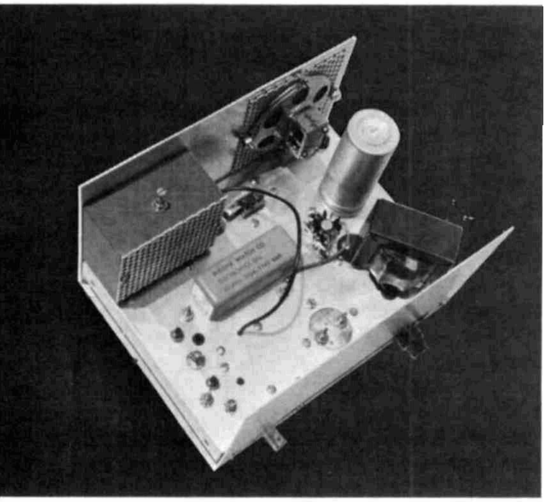
construction

My receiver is built in an LMB W1C utility case which affords more than ample room. You could probably get the electronics in the smaller LMB W1A and still not be crowded. The i-f amplifier-detector section was built on a 2 by 2-1/2-inch piece of copper laminate. I used this as I could easily solder clown the IC bypass capacitors and maintain minimum inductance in all leads. This is very important, since the more casual construction I tried at first produced an oscillator! After all, we are fooling around with an awful lot of gain in a small package. The CA3013 simply won't forgive a sloppy layout.

Each of the .05-μF bypass capacitors on pins 2, 3, and 4 are put in parallel with a 1000-pF standoff-type capacitor. In fact, the standard capacitors (which are soldered to the copper laminate) support the IC socket–a "spider" type, Nugent LP51710. Similarly, the .01-μF bypass at pin 10 is in parallel with the 1000-pF feed-through capacitor bringing +6 volts in from the top of the chassis. While parallel-bypassing sometimes has its pitfalls, it was the trick that "calmed down" this IC. I made the discriminator transformer more-or-less as described in the RCA spec sheet for the CA3013 and CA3014.[2] The spacing between L_5 - L_6 and L_7 was not specified, so you can play around with it. I used about 3/4-inch.

In addition, the PA222 is a high gain IC, and all the leads must be kept as short as practical. Since the PA222 has much less bandwidth than the CA3013, this job is a little easier. More details on the PA222 and how to use it arc described in references 3 and 4.

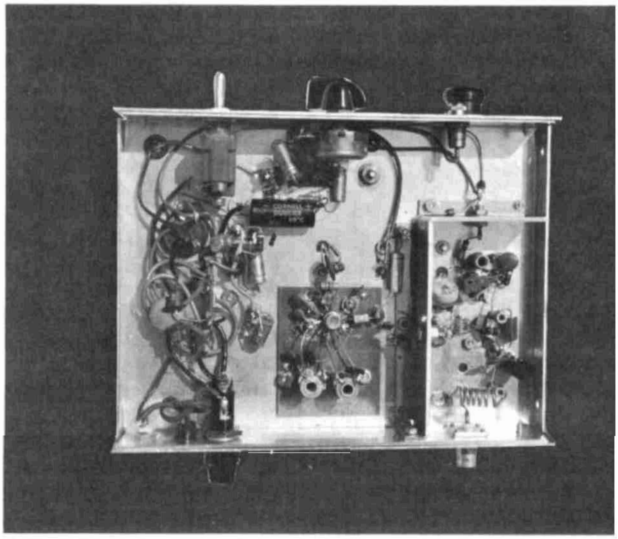
The vhf weather receiver. The tunable local oscillator is mounted on the box on the front panel just in front of the crystal filter.



local oscillator

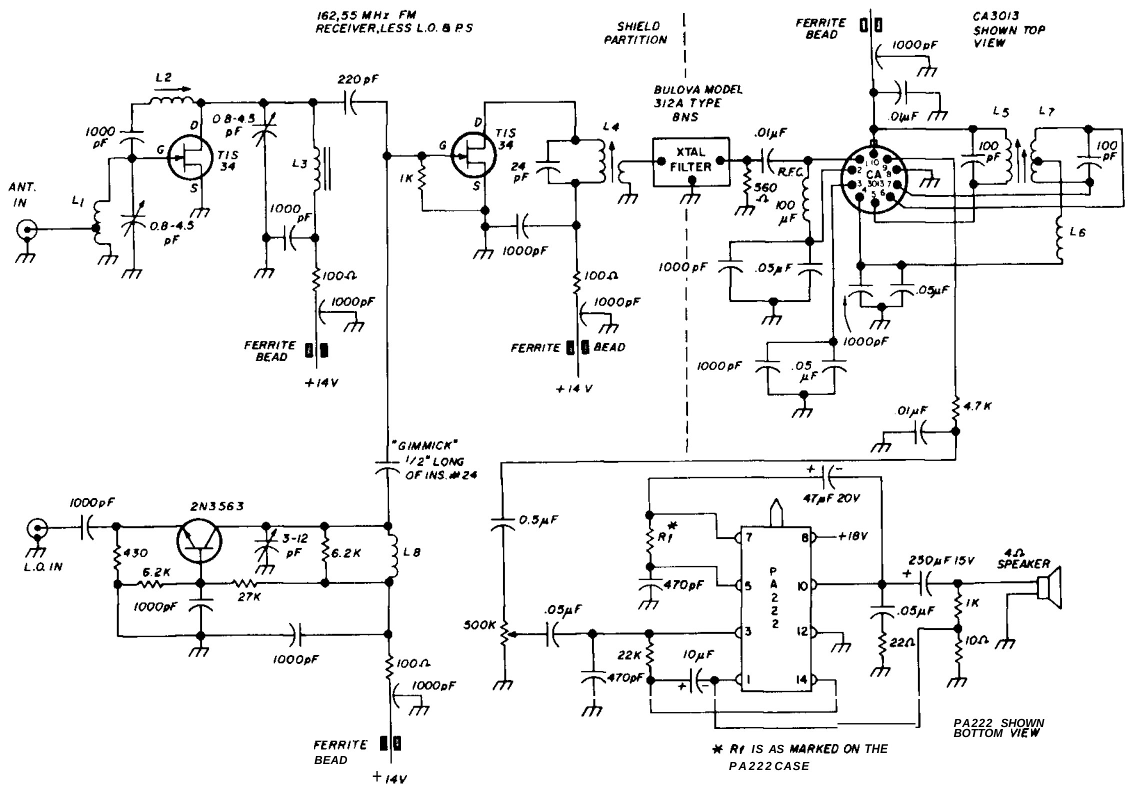
I used a tunable local oscillator because it was simple; after all, that's what every hi-fi FM tuner on the market uses. Stability is further improved by using an FET in the LO. Local-oscillator power is extracted from the FET circuit at a 50-ohm tap on the oscillator coil. A 10-dB pad and a common-base amplifier isolate the LO from the mixer and allow injection into the mixer gate (in parallel with the signal), through a small capacitor (less than 1 pF). Local-oscillator energy is delivered to the 2N3563 isolation amplifier through a 6-inch length of 50-ohm RG-174/U cable.

Below-chassis view of the receiver. The rf amplifier and mixer are located in the enclosure to the right.



The LO is built in a 3 x 2-1/2 x 1-1/2 box made of copper laminate with a perforated aluminum cover. The LO box is quite rigid and is mechanically tied to the front panel of the receiver. Flexure between the front panel and the chassis doesn't change the LO frequency because the box moves with the panel.

Several points about this design deserve additional discussion. The rf amplifier and mixer are quite similar to the circuits in K6HMO's two-meter converter.[5] It may be possible to eliminate the 2N3563 isolation amplifier by injecting the LO at the source,



L1 6 turns number 14, **7/8" long, 1/4"** ID. Tapped 1 turn from ground end.

L2 **0.35-0.52 µH (Delevan** 4000-04)

L3 5-1/2 turns number 22 enameled on **Amidon Associates T-30-13** toroid

L4 CTC **X2060-3;** secondary 10 turns number 28 enameled

L5 18 turns number 36 bifilar wound with L6 on 114" ceramic form

L6 18 turns number 36 wound bifilar with L5

L7 Two 9-turn number 36 bifilar windings on **1/4"** ceramic form; connected in series

L8 6 turns number 20, **3/8"** long. **1/4"** ID. air wound

fig. 2. Schematic diagram of the tunable vhf receiver. Although only one TIS-34 rf amplifier is shown here, another could be added for more sensitivity. The power supply and local oscillator are shown in fig. 3 and 4.

hut I haven't tried it.

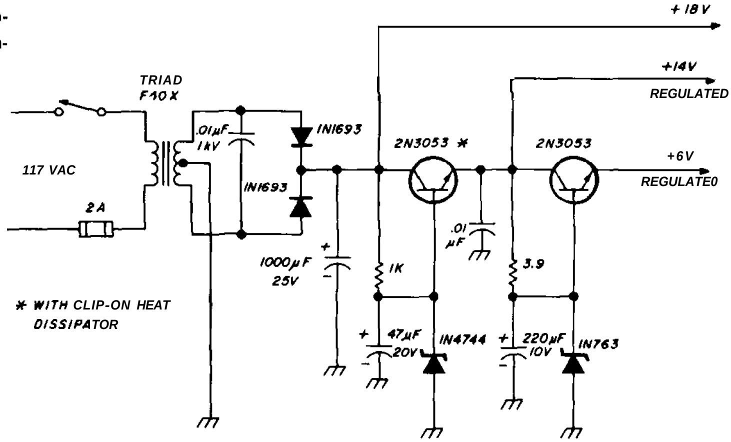
A crystal-controlled local oscillator would probably be aclvantageous for marine-mobile service, but it entails more circuitry if you want to suppress spurious responses. That is, if a 75.925-MHz crystal is used with a doubler to produce a 151.85-MHz LO, you must keep 75.925 out of the doubler output. Otherwise, the receiver **could** also receive TV channels 3 and 6.*

In K6HMO's two-meter cvnverter a similar effect can be noticed. Since he used a 58.5-MHz crystal with a doubler to generate a 117-MHz LO, the third harmonic (175.5 MHz) is also generated. The 175.5- and 117-MHz injection frequencies **each** convert 146.25 MHz to a 29.25-MHz i-f. However, at any other frequency in the two-meter band, two separate i-f frequencies will be found for the signal.

These two examples are used only to show that crystal control of the local oscillator is not without its hazards. The most direct method of using crystal control in this receiver is to use a seventh-mode crystal on 151.85 MHz. These crystals are available from at least one crystal manufacturer: James Knight.

* 75.925 MHz − 10.7 MHz = 65.225 MHz (channel 3) and 75.925 MHz + 10.7 MHz = 86.625 MHz (channel 6).

fig. 3. The power supply for the vhf weather receiver.



If you want to receive really weak signals, such as distant two-meter stations or mobile services (police etc.), it may be a good idea to use two TIS34 stages instead of just one. K6HMO's converter is designed that way. However, for the reception of KHB-49, San Francisco, some 20 miles away, this receiver is more than adequate in Palo Alto.

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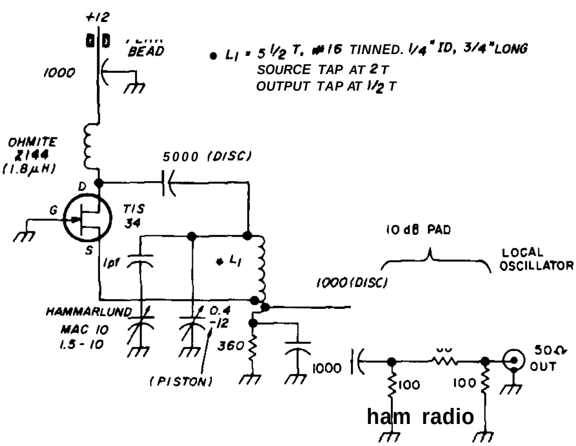
2. RCA, "CA3013, CA3014 Specification Sheet." 92 CM, February 1966.
3. General Electric, "PA222 Preliminary Specification Sheet," 85.20, March 1967
4. A. Petrie, "First Integrated Circuit Phonograph," *Electronics World*, December 1966, p. 28.
5. R. Friess, "A Low-Cost FET Two Meter Converter." 73, October 1966, p. 22.

fig. 4. With this tunable local oscillator, the receiver may be used for two meters and the various mobile services as well as the ESSA weather service. For single-channel reception, a crystal oscillator may be substituted as indicated in the text.



ham radio

phone patch legality

The legal status of phone patches may be clarified and improved if the Carter Electronics Corporation succeeds in its case against three major telephone companies.

In 1965, the FCC advised Carter Electronics that its cradle-type Carter-phone patches were prohibited hy Tariff 122. Carter filed an anti-trust suit against AT&T, the Associated Bell System and General Telephone. The Federal District Court requested a ruling hy the FCC, which ruled that the Carterphone indeed violated Tariff 122, but that the tariff was an "unwarranted interference with the telephone subscriber's right to use his telephone reasonably in a way that is privately beneficial without being publicly detrimental."

The eventual result of this case may be that if a customer wishes to attach gear to his line, his telephone company will have to show cause for denial, a complete reversal of the present situation.

Jim Ashe, W2DXH

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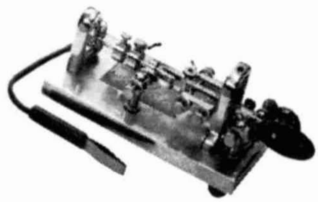

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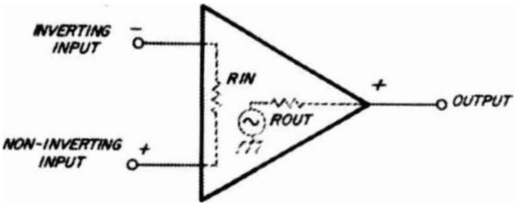


fig. 1. Basically, the integrated-circuit **operational amplifier** is simply an **amplifying device** with **two** inputs.

amateur uses of the MC1530 IC

Although these circuits were designed around the MC 1530, other types of linear IC's may be used in similar circuits

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

In the past, integrated circuits have app ared as winrlow dressing in some equipment when one or two transistor stages would have done the same thing at lower cost. However, the rapidly decreasing cost of IC's make them more applicable to many amateur uses. The Motorola MC 1530, a typical IC operational amplifier, tan be used as an audio amplifier, audio compressor, i-f amplifier or oscillator with better results than several transistor stages.

The MC1530 is typical of a class of integrated circuits known as operational amplifiers. It is rather expensive, about $15, but all types of surplus IC's are becoming available at prices ranging from one to three dollars. The applications described here are applicable to almost any operational-amplifier IC– as long as you consider the frequency range and power-supply requirements.

Operational amplifiers are essentially multistage transistor amplifiers wired so that the connection of external elements determines the frequency response and gain. By connecting external capacitors, coils, etc., IC's can be used as audio amplifiers, rf amplifiers or oscillator\.

Normally, the operational-amplifier IC uses dc-coupled amplifier stages so no internal capacitors are required. The lack of capacitive coupling permits the amplifier to be used as a dc amplifier. In addition, the frequency range is not restricted by any capacitive-coupling effects.

Integrated circuits offer a number of advantages over conventional, discrete-component amplifiers. Perhaps the greatest advantage is the large amount of signal amplification available in a very compact package. The MC 1530, for example, provides a maximum signal gain of 74 dB, yet it isn't any bigger than a small audio transistor. The input and output terminals in the transistor circuits are separated by several inches. Good input/output shielding is imperative when you use IC's. The IC also has a frequency response beyond that of a simple transistor-amplifier stage.

The MC 1530 will operate from dc to well beyond 14 MHz. Proper bypassing is very important if you want to avoid spurious responses and oscillations. The application notes available from the IC manufacturer detail the circuit precautions that are necessary for stable operation; they should be followed closely. If you follow the manufacturer's recommendations, you'll find that IC's are more

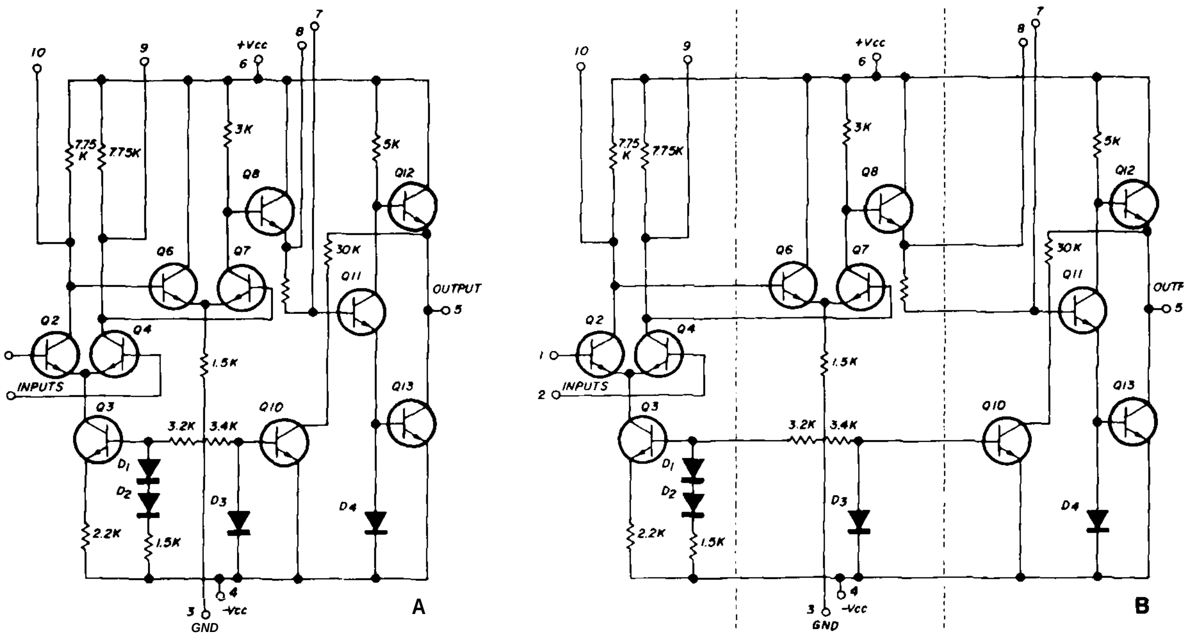


fig. 2. The complete internal circuit of an IC is rather complicated (A), but when it is broken into sections as shown in B, the circuit is much more straight forward.

temperature stability of integrated circuits is better than conventional transistor circuits because of the circuit configuration and construction.

The large amount of gain available in such a small package also indicates some of the problems it can cause For instance, you must remember that you're dealing with a device with input and output terminals which are only separated by 1/4 inch. Although the gain provided by the IC is equivalent to a four- or five-stage transistor amplifier, the in-stable and reliable than many conventional transistor amplifiers.

the MC1530 operational amplifier

The best way to look at the MC 1530 is as the amplifying block shown in **fig. 1.** Maximum gain is 74 dB, and the frequency response is essentially flat from dc to 14 MHz. It has a high input impedance (about 50k ohms) and low output impedance (about 500 ohms). Two supply voltages are required–plus and minus 6 volts. Power-supply filtering is about the same as that required for most transistor amplifiers. Maximum supply current is on the order of 10 mA.

Unlike conventional amplifiers, however, note that there is both a "non-inverting" and "inverting" input. The output signal is inverted (180° out of phase) when the "inverting" input is used. In other words, a positive-going input produces a negative-going output.

The output signal will be in phase with the input if the non-inverting input is used. If the two inputs are fed simultaneously, the phase and amplitude differences between the input signals govern the output-signal characteristics.

Although you can look at the operational amplifier simply as a compact amplifying device, it is better to understand what is going on inside. This is also helpful when you want to use the IC in a practical circuit.

The schematic of the operational amplifier shown in **fig. 2A** is confusing and it's difficult to distinguish between stage functions. However, if you divide the sections of the IC properly, as shown in **fig. 2B,** the relationships are not difficult to understand.

A description of the MC 1530 stages is informative, since it applies to many other operational amplifiers as well. In the input of the MC 1530, Q2 and Q4 form a differential-amplifier stage. If the base of either transistor is grounded, the output is inverted or non-inverted, depending on which transistor is grounded. If you use both inputs, the amplified signal is the difference between the two inputs. This feature is used in the compressor circuit shown in **fig. 8.**

The input to the MC1530 can vary from several millivolts up to a maximum of 2 volts. A constant current source, Q3, is connected to the emitters of both Q2 and Q4. This stabilizes the input stage operating points.

The second differential amplifier, Q6 and Q7, is driven by Q2 and Q4. The output of this amplifier stage is single-ended, however, and taken from the collector of Q7 through an emitter-follower stage, Q8. Transistor Q11 functions as a voltage-'evel translator stage and is necessary because the amplifier stages are dc coupled.

fig. 3. The high-frequency rolloff characteristics of this audio amplifier are determined by the capacitor connected between terminals 9 and 10.

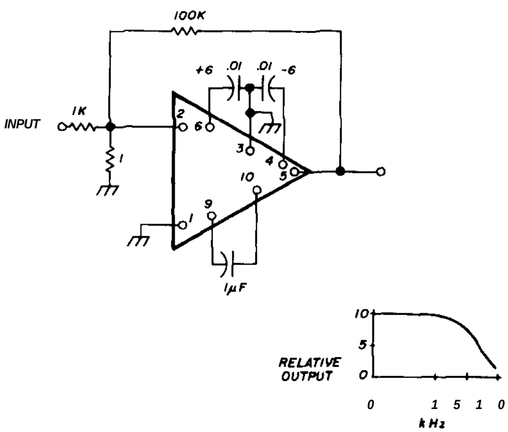
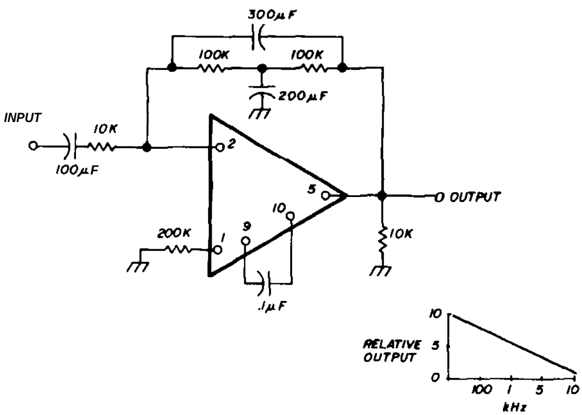


fig. 4. The IC integrator circuit provides constantly decreasing gain with frequency.



In conventional capacitance-coupled amplifiers, the capacitors block the dc voltage from preceding stages so that the signal at the input of a stage swings around zero volts. This cannot be done with directly-coupled stages, so level translators are used to accomplish basically the same function. The three-volt offset (approximate) at the output of Q8 is shifted to zero volts by Q11; Q10 serves as a stabilized dc current source for Q11.

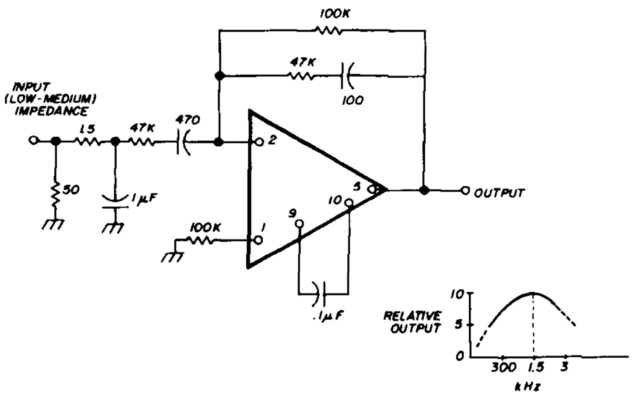
Transistors Q12 and Q13 form a conventional direct-coupled output stage. Terminal connections for the MC 1530 IC are fairly obvious from the diagrams in **fig.** 2. Terminals 7, 8, 9 and 10 are used for connecting external components between the internal stages of the IC so that you can control frequency response, gain and phase characteristics.

In a sense, working with an IC is like working with an amplifier where only the input/output, power supply and plate or load connections are accessible. It should be obvious that each connection has a purpose, and although all connections may not be used for any given application, you don't ground the unused pins! Some experimenters, who never bothered to understand the basics of IC circuitry, have done this with rather unusual and expensive results.

applications

While all of the following circuits are designed around the MC 1530, most of them are suited to other integratod-circuit operational amplifiers as well. Terminal connections will differ, of course, and you should consult the manufacturer's data sheet before using the device. Power-supply voltages are not exceptionally critical, but don't exceed the recommended values. Handle them like a transistor–but remember it doesn't take much overvoltage to destroy them in most cases. However, when they are used within their proper ranges, their reliability exceeds that of most discrete transistors.

fig. 5. By adding additional components to the simple audio amplifier shown in fig. 3, both low- and high-frequency gain rolloff characteristics can be obtained.



audio amplifier

A simple, low-level, high-gain audio amplifier is shown in **fig.** 3. Very few external components are required. With high-impedance microphones of reasonable output, this stage may be used as a microphone preamplifier or between a low-noise preamp and a high-level amplifier.

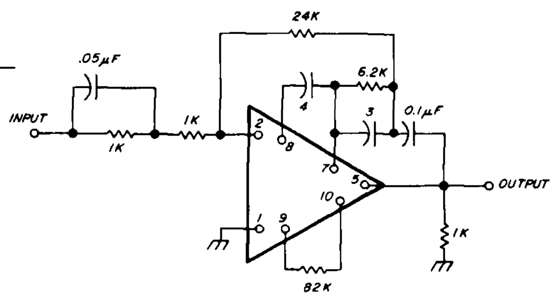
With the 1-µF capacitor connected between terminals 9 and 10, the gain is constant from dc to about 1 kHz; the gain is about 20-dB down at 3.5 kHz. You can vary the value of this capacitor for the upper frequency where you want the gain to roll off. If a .01-µF capacitor is used, for example, the gain will start to roll off just above 10 kHz.

integrator

The integrator circuit gives maximum gain at the low frequencies and constantly decreasing gain with increasing frequency. This is in contrast to the previous circuit where gain is relatively constant up to the roll-off frequency.

In communications circuits, integrator circuits are usually used as low-pass filters. For instance, an integrator may be used after an audio-clipper stage to block the high-order harmonics which the clipping action generates. If they're not eliminated, they may overload later stages or produce audio distortion. The circuit in **fig. 4** produces about 55-dB gain at 100 Hz and drops linearly to 25-dB gain at 3,000 **Hz** and 10-dB gain at 10 kHz. This is a gain reduction of nearly 45 dB between a 100- and 10,000-Hz signal.

fig. 6. This extremely broad-band amplifier works from dc to over 20 MHz.



bandpass audio circuit

By making a slight change in the external RC circuitry used with the IC shown in **fig. 5,** roll-off characteristics can be obtained which will favor the audio range from 300 to 3,000 hertz. With the values shown, the maximum gain of the amplifier is about 50 dB.

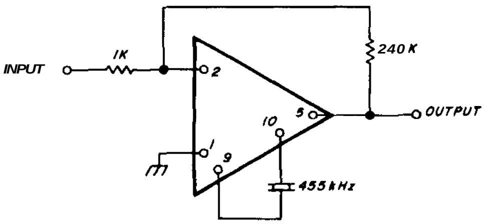
Gain peaks at about 1500 Hz and is 10 dB down at 400 and 4,000 Hz. With variable components, both the upper and lower frequency roll-off points can be adjusted.

broad-band rf amplifier

With the proper external circuitry, the MC 1530 can be used as an extremely broad-band rf amplifier. The circuit shown in fig. 6 provides a constant 20-dB gain from dc to 10 MHz and falls off to 15 dB at 20 MHz. This circuit may be used with various test instruments to extend their useful range or in such devices as broad-band monitors. You could even use it for additional i-f gain or as a broad-band rf preamplifier.

If the input and output circuits are tuned, the gain can be increased of course. As a tuned amplifier, you should be able to obtain at least 50-dB gain at any frequency in the zero to 15-MHz range.

fig. 7. A crystal-filter amplifier using the MC-1530.



crystal-filter i-f amplifier

Fig. 7 shows how the MC 1530 can be used with a crystal to provide a high-gain, selective amplifier. This stage would be very useful in a receiver or transceiver to provide additional selectivity for CW reception. The gain of the circuit is more than enough to compensate for the attenuation introduced by the crystal. The circuit can be connected between the i-f stages in a receiver and directly bypassed (or a capacitor switched in place of the crystal) whenever the additional selectivity is not desired. If you want to be more elaborate, the single crystal can be replaced by a crystal filter.

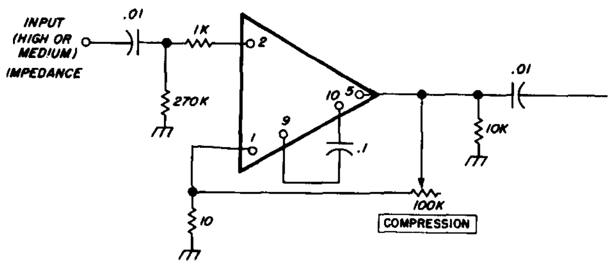
The photograph at the beginning of the article shows how this circuit can be assembled on a small piece of Vectorboard. The MC 1530 is practically lost when compared to the size of the crystal. The construction is typical of the circuits shown and demonstrates how compact circuits can be easily assembled.

audio compressor/amplifier

By taking advantage of the fact that the two inputs to the MC 1530 produce both an inverted and non-inverted output, an extremely simple but effective compressor circuit can be developed. Although the circuit values shown in **fig. 8** are those which I found to be most effective for an audio compressor, there is no reason why compression action could not be applied to an rf stage by changing the capacitor values.

This circuit is relatively easy to understand. The input is directly coupled to the non-inverting input of the IC. A small amount of the output signal is fed through the 100k-ohm **compression** control and 10-ohm voltage divider to the inverting input (pin 1). The amount of feedback voltage is determined by the value of the 100k pot and must be determined experimentally for the desired degree of compression. Tests I have made indicate that a compression range of at least 15-20 dB is easily attainable.

fig. 8. Simple audio compressor using the MC-1530. For minimum noise, this circuit should be preceded by a low-gain, low noise audio amplifier.

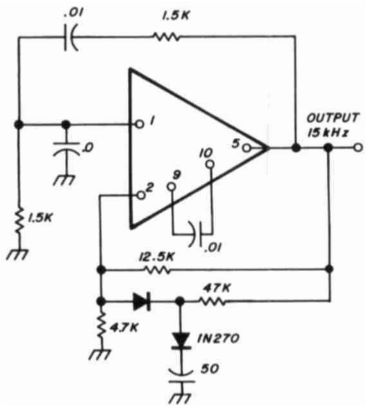


The capacitor between terminals 9 and 10 is used to provide a high-frequency roll-off at about 3,000 Hz to eliminate unwanted noise amplification. The value of this capacitor is not critical, but it should not be less than 0.05 μF if the signal-to-noise factor of the amplifier is to be maintained at a reasonable level.

audio/rf oscillator

Since the MC 1530 is a high-gain amplifier, there is no reason why it can't be used as an oscillator. Fig. 9 illustrates an MC 1530 oscillator which uses both positive and negative feedback to produce a highly stable oscillator operating at 15 kHz. The output waveform is almost harmonic-free. By varying the RC feedback circuit connected between pin 5 and pins 1 and 2 and pin 2 and ground, you can build an oscillator from a few hundred Hz to several MHz.

fig. 9. This 15-kHz sine-wave oscillator provides a very pure output waveform.



Proper bypassing between pins 2, 9 and 10 and ground is necessary to prevent any spurious outputs. The frequency can be varied over a ten-percent range by varying either the resistive or capacitive element of the RC network between pins 1 and 5.

summary

I have tried to demonstrate the extreme versatility of the integrated-circuit operational amplifier–from broad-band amplifiers to audio compressors and oscillators. It is only a matter of time before they are incorporated into commercial amateur equipment. In the meantime, you can use them to advantage to construct home-brew station or accessory equipment. The cautions are relatively simple and no more restrictive than those necessary with early transistors. The advantages are many–versatility, space and cost.

ham radio

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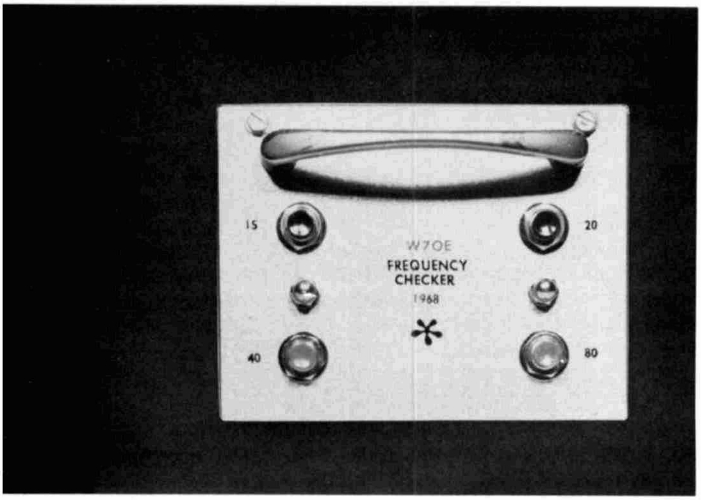
Turn to page 85



The mini-spotter frequency checker. The handle is the antenna! Four push-button switches choose the desired output frequency.

mini-spotter frequency checker

With this spotter you can hit your nets and schedules right on the nose

Howard S. Pyle, W7OE, 3434 34th Avenue, S. E., Mercer Island, Washington

For the dedicated ham who likes spot-frequency operation for schedules and nets, accurate dial setting is a must. A stab in the dark with the VFO or a vague mark on the receiver tuning dial may give you an approximate setting, but you're going to have to dance around a few kilohertz either way to spot your man. There's an easier way–it will put you right on the button every time.

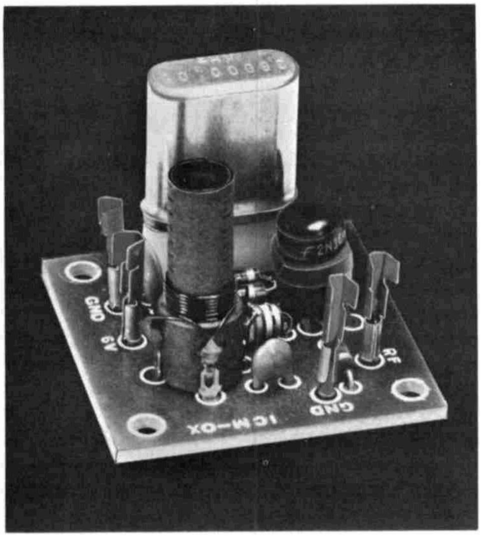
Maybe you work traffic nets–perhaps a frequent schedule with a friend in a distant state or across the pond. If you're right on frequency when the NCS or distant pal calls, it practically guarantees your QSO when hand conditions are right.

I had the same problem not long ago; regularly scheduled operation with other members of the QRP Amateur Radio Club-International on designated club frequencies on 15, 20, 40 and 80 meters. It's common practice at schedule time to set the VFO dial where you **think** it should be, zero your receiver, and call your man. No go. Diddle the VFO a bit and try again. Maybe you catch him this time, more likely not. Try again! Ah, there he is; you finally hit him! Meanwhile, you've been dancing around the receiver dial hoping he'll be calling you so you can zero in the VFO.

All that monkey business is childish. How much traffic do you suppose the commercials would move between ship-and-shore or point-to-point if they had to fuss around like that to find a guy who is supposedly right on the bottom?

I licked my transmitter problem by switching to crystal control for these contacts, leaving the VFO for casual excursions around the bands. The receiver relied on the crystal oscillator for a zero beat. However, the normal output from the crystal oscillator (final switched off) was too much for the receiver. Even with the rf gain backed down, the zero beat was too wide because of over-loading.

An International Crystal OX crystal-oscillator board.



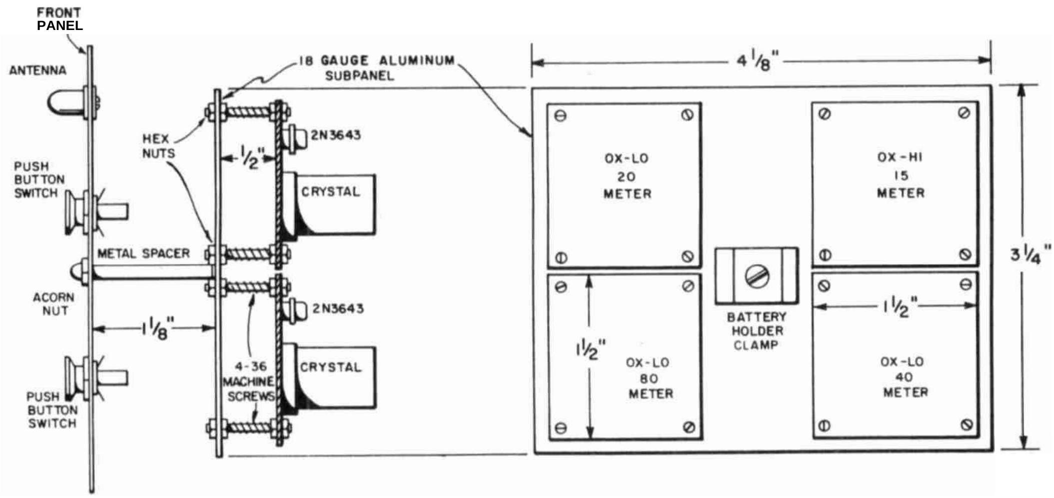
The solution was simple: a low-powered oscillator on each of the frequencies I was interested in spotting. Press a button for the selected frequency, zero in on the receiver and, if the transmitter is VFO controlled, you've got a perfect spotting signal on which to zero. Better yet, if you're standing by for a call on one of your spot frequencies, or if you monitor certain frequencies a good deal of the time, you don't have to keep the transmitter hot for occasional frequency checks. With this little frequency spotter just push the button and check your receiver zero at any time!

crystal oscillators

I maintain spot-frequency schedules on four frequencies, so my little spotter is equipped with four separate crystal oscillators. You can cut this down or add more as you see fit. While it may sound costly to provide separate oscillators, actually it's not. Each oscillator cost me $2.35; the cabinet and miscellaneous hardware were about $3 more. Using crystals I already had on hand, :he whole four-frequency business came out well under $15! With less oscillators of course, the cost will be even less.

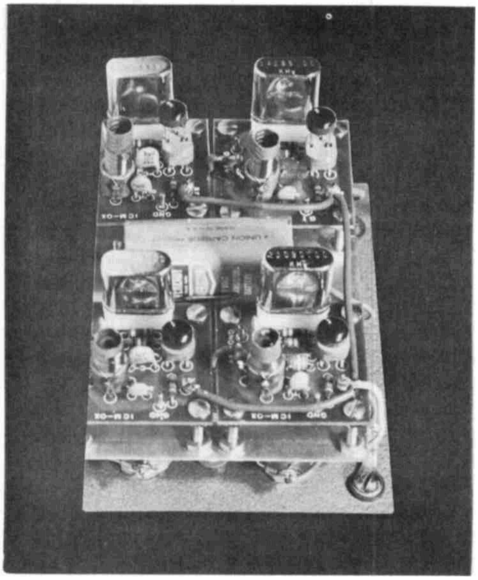
Each oscillator is one of the OX types recently introduced by the International Crys-

fig. 1. Layout of the mini-spotter sub-panel.



tal Company*. These are supplied complete with the transistor, printed-circuit board, all components and hardware; everything you

Internal layout of the mini-spotter. The battery is mounted in the center between the oscillator boards.



need except the crystal. You can use one you already have or you can buy an EX crystal from International.

Assembling the kit is duck soup; I averaged 20 minutes for each oscillator including the soldering and cleaning off the resin residue! All the holes are drilled in the printed-circuit hoard and the position of all components is plainly labelled–you can't go wrong.

construction

Each little oscillator is mounted with four corner screws (yes, they even come with the kit). I didn't want to "cheese-hole" the front panel with sixteen mounting holes and screw heads, so I used a piece of scrap aluminum for a subpanel to mount the oscillators on. This way, the subpanel was mounted to the front panel with only two screws. Spacers hold it one inch away from the rear of the front panel. The 9-volt transistor battery which powers all oscillators is mounted between the oscillators on the sub-panel with

• International Crystal Company. Inc., 18 North Ler, Oklahoma City, Oklahoma 70102.

a small wiring clamp as shown in the photograph.

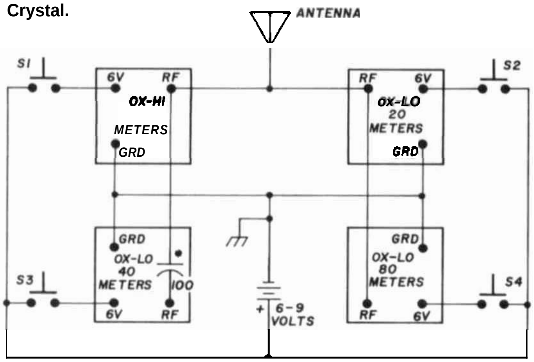
I housed the whole business in a Bud AU-1028* 3 x 5 x 4-inch utility cabinet. If you use a different number of oscillators, you'll probably want to use a different cabinet.

Don't let the handle on the cabinet face fool you! Sure, it's a convenient way to move the little spotter around, but it is actually the antenna for the oscillator outputs! It's mounted to the front panel with shoulder-type insulating washers slipped over the mounting screws so that it doesn't make electrical contact with the panel.

The rf output terminals from the oscillators are paralleled as shown in the wiring diagram so there is no need for any switching. Pushing the selected frequency button simply applies a positive voltage from the battery to that particular oscillator. Its output then feeds the handle antenna.

Battery life is indefinite. I've used my spotter for many months and it looks as though there are still years of soup left! The current drain is only momentary and is approximately 29 mils when a button is depressed.

fig. 2. Wiring diagram for the mini-spotter. Each of the oscillator boards is en OX kit from International Crystal.



There you have it, a frequency spotter that will put you right on the nose for schedules and net operations simply by pressing a push-button. This is about the handiest little gadget I have in my shack and I wouldn't be without it.

*** Gray hammertonr, $2.05 plus postage from Allied Electronics, 100 N. Western Avenue, Chicago, Illinois 60680. Ordrr catalog number 42 E 8549.**

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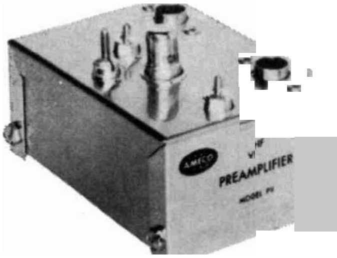

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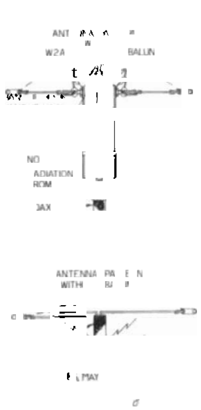

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updating the BC-603 tank receiver

Here's how to breathe new life into an old receiver

Jim Harvey, WA6IAK, 1401 E derby Way, Sunnyvale, California 94087

For three or four years I've been meaning to modify my old RC-603 to see if I couldn't put some life back into it. In the meantime, I have used it as a tunable i-f and demodulator for a number of vhf converters. When I bought it several years ago, it was in pretty sad shape -a previous owner had, among other things, tried to change the basic tuning range from 20 to 28 MHz to 50 to 54 MHz. This operation butchered up the rf and mixer-oscillator coils and removed one of the plates from the oscillator-tuning capacitor.

For some reason, the squelch circuit never worked reliably and the receiver was about as noisy as a one tube regen–it had little, if any, limiting. To top off this unhappy state of affairs, the sensitivity left plenty to be desired; at times it seemed to require a full volt to get full quieting.

After examining the existing circuit to see what would give me the most immediate improvement, I concluded it would be easier to rebuild the set than try to fix it. This also seemed to be the best way to increase overall performance.

There were other things about the receiver's condition that bothered me: aging resistors that were probably noisy and scores of large mica bypass capacitors similar to ones which had caused me so much grief in other old gear. Finally, I wasn't very impressed by the 6J5/6AC7 tube lineup–not to mention the 6H6's which must have gone out with high-button shoes!

Rebuilding the receiver would not only eliminate these old parts, it would make it easier to include both a-m and fm capability. The decision to rebuild the BC-603 eliminates this as a modification article.

Instead, I will try to give you a few ideas and suggestions on rebuilding it. If a complete rebuild doesn't appeal to you, you may still get some good ideas; each part of the new circuit will stand on its own as a BC-603 modification.

disassembly

The mechanical components, i-f transformers, rf coils and tuning assembly are well built. With some new capacitors and resistors, they are all used in the final receiver. The first task is to remove all the wiring. It isn't easy to start if your old BC-603 is still in working order, as mine was. Once you've started, there's no turning back. The front panel is saved, wiring and all. Take out the screws holding it to the receiver chassis and set it aside until later.

A word of caution: if you have one of the BC-603's that was fungus proofed, don't breathe the fumes that are produced when you're desoldering the old solder joints.

Remove all the octal tube sockets except the one near the front of the chassis right next to the jack for the front panel V9. Remove the phenolic strip along the side of the chassis near the demodulator, squelch and audio sections. Remove all the rf and i-f coils and set them aside; they will be modified before they are re-installed. I also took out the 2-μF electrolytic capacitor near the front panel jack because I couldn't find any good use for it.

Since I didn't need them, I took out several other items to reduce weight:

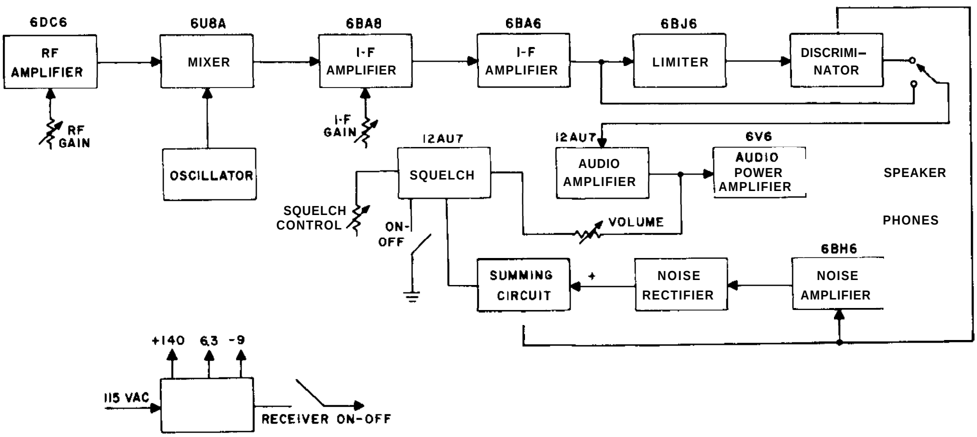
1. The three bathtub capacitors on the side of the main-tuning-capacitor housing, C15, C23 and C24.

2. The metal cable guide that runs across the side of the chassis underneath the BFO and discriminator cans.

3. The multipin rear connector that I could never find a mate for (not the one to the dynamotor housing).

4. The antenna and ground jacks which stick out from the front of the chassis.

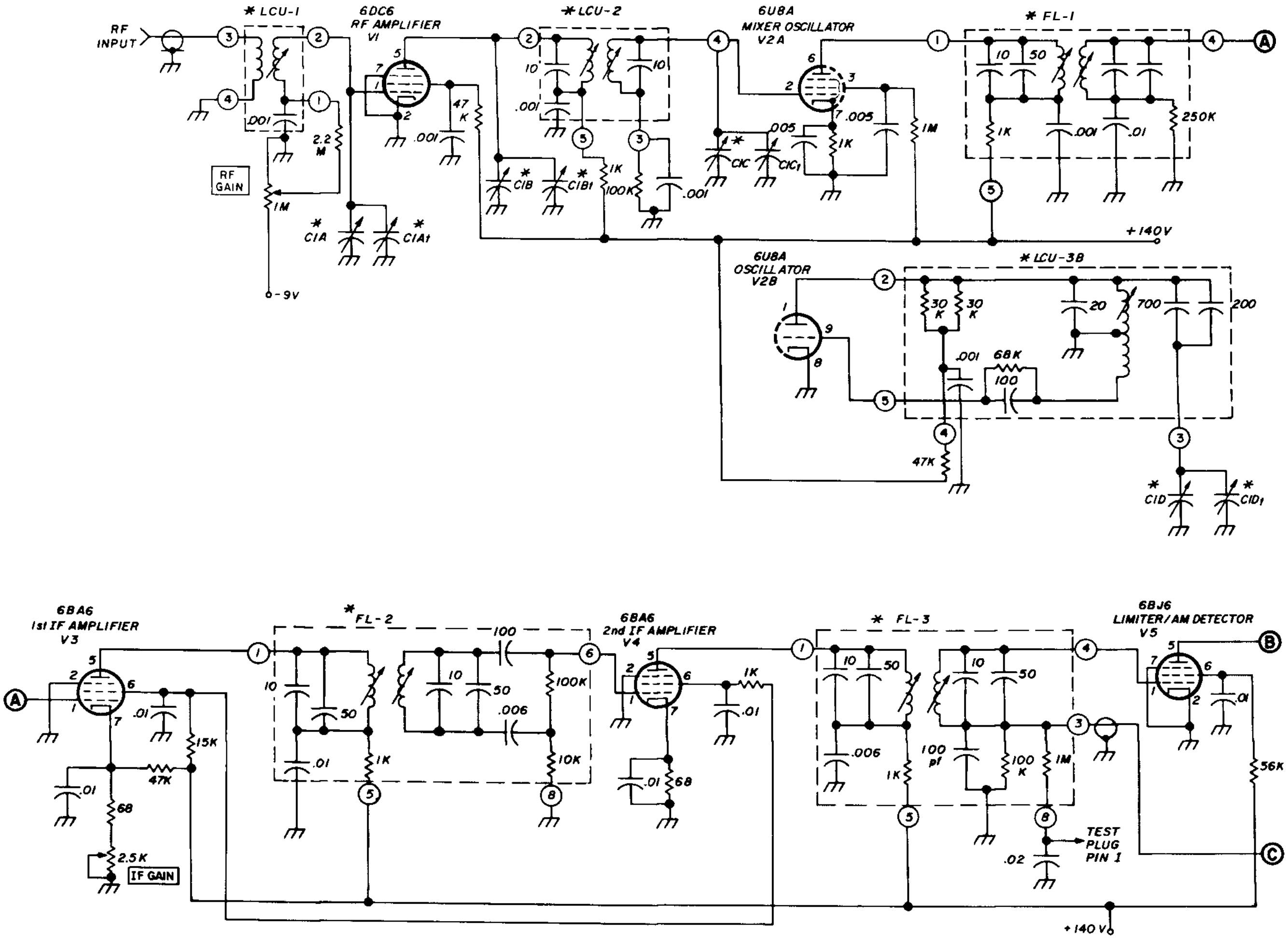
fig. 1. Block diagram of the updated BC-603.



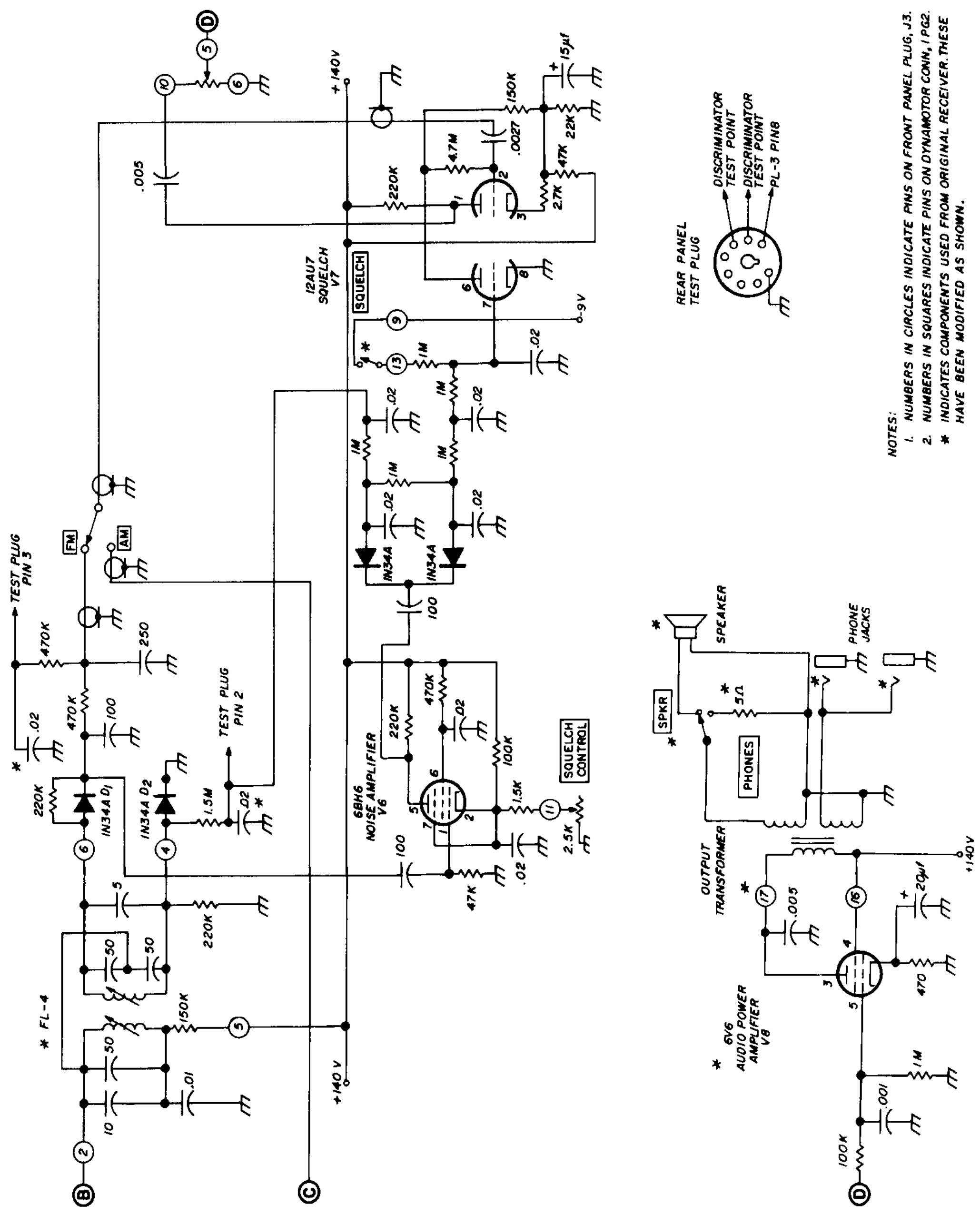
construction hints

I won't attempt to describe every detail of how the receiver is built. I'll just describe the sticky areas that might hold up progress on the various stages during construction.

The mechanical work is very slight; this is one of the advantages of using an old receiver as a starting point. Mounting plates for the



RF INPUT
* LCU-1
RF GAIN
-9V
6DC6
RF AMPLIFIER
V1
* LCU-2
6U8A
MIXER OSCILLATOR
V2A
* FL-1
+140V
6U8A
OSCILLATOR
V2B
* LCU-3B
6BA6
1st IF AMPLIFIER
V3
IF GAIN
* FL-2
6BA6
2nd IF AMPLIFIER
V4
* FL-3
6BJ6
LIMITER/AM DETECTOR
V5
TEST
PLUG
PIN 1
+140 V



TEST PLUG PIN 3
TEST PLUG PIN 2
FL-4
IN34A D1
IN34A D2
FM
AM
+140V
12AU7 SQUELCH V7
SQUELCH
6BH6 NOISE AMPLIFIER V6
SQUELCH CONTROL
IN34A
-9V
REAR PANEL TEST PLUG
DISCRIMINATOR TEST POINT
DISCRIMINATOR TEST POINT
PL-3 PIN8
SPEAKER
PHONE JACKS
SPKR
PHONES
OUTPUT TRANSFORMER
6V6 AUDIO POWER AMPLIFIER V8
NOTES:
1. NUMBERS IN CIRCLES INDICATE PINS ON FRONT PANEL PLUG, J3.
2. NUMBERS IN SQUARES INDICATE PINS ON DYNAMOTOR CONN, I PG2.
* INDICATES COMPONENTS USED FROM ORIGINAL RECEIVER. THESE HAVE BEEN MODIFIED AS SHOWN.

Schematic

smaller 7-pin and 9-pin tube sockets were cut from thin gauge aluminum with a pair of shears using the pattern in **fig. 4.** These adapter plates are installed in the holes left when the old octal sockets were removed. Use a ground lug under each of the screws that hold a plate to the chassis. If you do this on all of the plates, it will provide a convenient place for grounding things later on.

Install miniature 7-pin tube sockets in the chassis holes previously used by the rf amplifier (V1), first i-f amplifier (V4), second i-f amplifier (V5) and limiter (V6). Install miniature 9-pin sockets in the vacant 6AC7 mixer socket (V2) and 6V6 audio-output socket (V8).

Several multi-lug terminal strips help to route B+ and heater voltage; they also support the discriminator and noise-amplifier diodes and other components. Other parts are either inside the cans, as shown on the schematic diagram, or tied directly to tube or coil pins underneath the chassis.

The important thing to remember is to keep all unbypassed signal leads as short as possible.

fig. 3. Power supply for the BC-603 tank receiver.

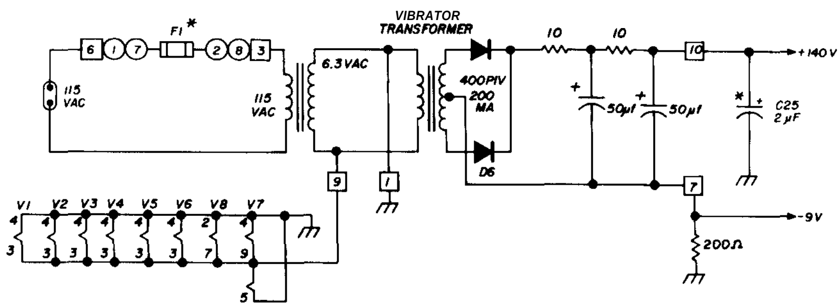
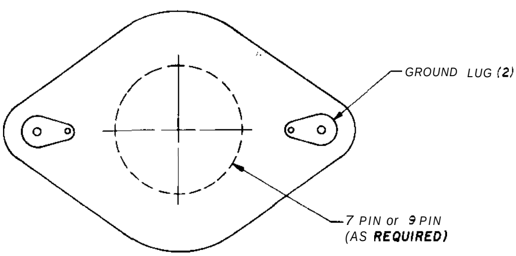


fig. 4. Mounting plate for the new 7- and 9-pin miniature tube sockets.



coil modifications

A block diagram of the finished receiver is shown in **fig. 1.** You can see from this diagram that the receiver is quite conventional. The rf stage is a low-noise 6DC6 semi-remote cut-off pentode. Both rf coil cans, LCU-1 and LCU-2, were modified as were the rest of the coils in the receiver. The schematic, **fig. 2** shows these coils in their modified configuration.

My BC-603 has a schematic glued to the bottom of the cabinet; I assume this was standard practice with these receivers so I will not present the old schematic here. Also, you don't have to have the original schematic to construct this receiver. Just make sure that the coils are modified as shown in the schematic.

Except for the oscillator coil, all of the coils are easily removed from the cans. The oscillator-coil can is sealed and must be cut open near the bottom. Be careful–make a clean, straight cut, modify the coil as shown in the schematic, and resolder the can. I used the soldering attachment on a small butane torch to tin the can and to do the soldering.

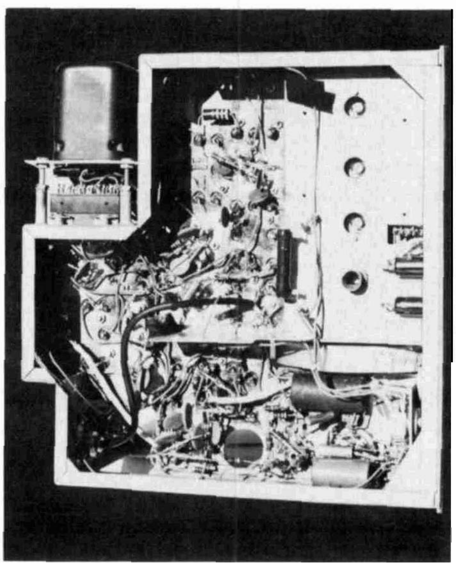
the new circuits

Not all receivers will have the same tuning range after you put in the new lower-capacitance tubes. The rf, oscillator and i-f sections must be tuned after completion. My unit covered 21 to 30 MHz with oscillator injection on the high side of the i-f by 2.65 MHz. This was a result of the missing oscillator-tuning capacitor plate mentioned earlier. No coupling, other than inter-electrode capacitance, is required for good mixing action in the 6U8A. Most of the headaches of this mixer stage, and there were a few, were cured by a book on receivers.'

The two 6BA6 i-f amplifier stages are quite conventional and need no explanation. They worked the first time power was applied and have given no trouble since.

I wanted to cover both a-m and fm modes in my unit. The grid of the limiter tube is the a-m demodulator. Grid rectification due to the unhypassed 100k grid resistor provides very acceptable a-m detection. The fm discriminator transformer FL-4 is extensively cleaned up, and a more conventional circuit using 1N34A detectors is substituted.

Inside the rebuilt BC-603. The power supply is mounted in the old dynamotor space.



squelch and noise amplifier

The squelch circuit deserves some explanation. It may seem like a lot of trouble to go through, but it gives reliable results and really isn't as bad as it looks. The circuit is made up of a noise amplifier, a noise rectifier, and a keyer tube which turns the first audio amplifier tube on and off. When no signal is present, the noise at the output of the discriminator is filtered to remove any low-frequency components and then amplified by the 6BH6 noise-amplifier tube; the noise is then rectified by the 1N34A's. This produces a positive potential, which, when applied through a summing circuit to the grid of the squelch-keyer tube, half of the 12AU7, causes it to conduct heavily. With heavy conduction, its plate voltage is about 20 volts lower than its normal non-conduction voltage.

The plate voltage of the keyer tube is used to bias the grid of the other half of the 12AU7 –the audio preamplifier. In the full-conducting state, the grid bias on the audio amplifier is approximately -20 volts with respect to its cathode. The tube is cut off in this condition, and the audio is squelched.

When a signal appears in the bandpass of the receiver, the noise in the discriminator output disappears because of quieting, and the positive voltage which was being generated by the noise disappears at the summing circuit. At the same time, a negative potential is present in the secondary of the discriminator transformer across diode D2. This negative potential appears at the summing circuit and cuts the squelch-keyer tube off. When the squelch keyer is cut off, its plate voltage rises and turns on the audio amplifier stage. Voltage from a minus 9-volt bias supply can be switched from the front panel to cut the squelch keyer tube off and effectively disable the squelch circuit when desired.

power supply

The power supply uses a 6.3-volt filament transformer and a 6-volt vibrator transformer. The 6.3 Vac supply runs all tube heaters and provides an input voltage to the 6-volt vibrator transformer. I used a surplus 6.5-volt filament transformer with an 8-amp rating.

You can find a vibrator transformer in almost any car radio, old mobile equipment or in the junk box, or you can buy a new transformer that has the output you need. However, make sure the transformer will fit in the available space and has sufficient rating.

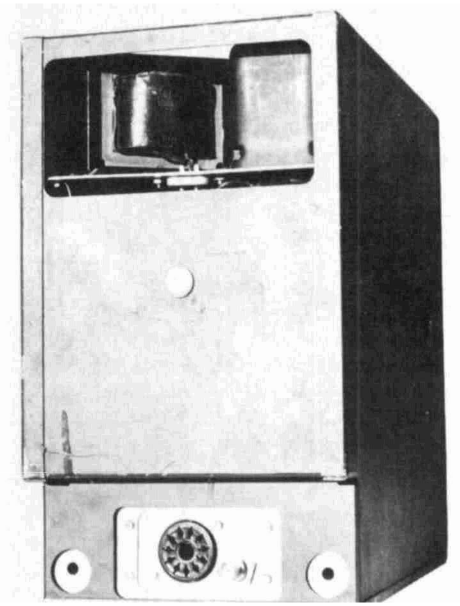
The power supply is mounted on an aluminum plate which is supported by the threaded risers in the dynamotor housing as shown in the photograph. I used a TV cheater cord and receptacle for ac line voltage for easy removal of the chassis from the cabinet. The power and switch leads are brought into the receiver through the existing Jones plug on the dynamotor housing. Pin numbers for the Jonesplug are marked in the square boxes on the schematic diagram; I used the original BC-603 Jones plug, PG2.

The target voltage is 140 to 150 volts at 60 mA. The minus 9-volt bias is obtained through a 200-ohm dropping resistor in the power-supply return lead.

front panel

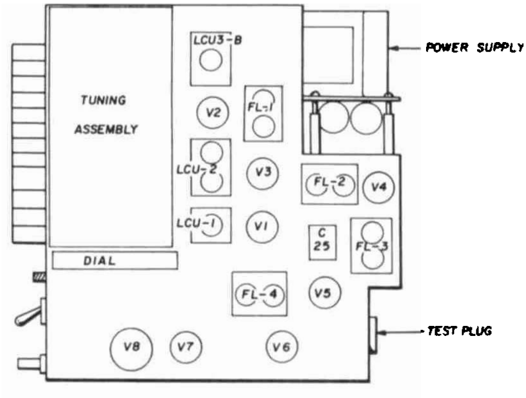
Front panel wiring is essentially untouched. However, the tie point which connects the front panel squelch on-off switch, the volume, and the squelch controls to ground must be removed. Then the panel is wired for the front panel plug. The pin designations are shown in the circles on the schematic diagram in fig. 2. The volume control and squelch pots were replaced with the values shown in the schematic.

Back view of the reborn BC-603 with the cover on.



The speaker in my receiver was had and was causing distortion in the audio so I replaced it with a shallow 4-inch speaker. The heavy iron grillwork was taken off and replaced with a light aluminum grill made by bending some scrap aluminum in a vise and drilling a pattern of holes in it.

The rf and i-f gain control pots are installed side by side in the holes left by the removal of the antenna and ground posts. Small diameter pots must he used to allow clearance; I used a 718-inch diameter Centralab units.



VIEW PROM RIGHT SIDE

tracking adjustment and tune up

Tuning up a receiver is not an easy task without some test equipment. An rf signal generator, preferably one with an attenuator, that covers the i-f frequency of 2.65 MHz and the range from 20 to 35 MHz is almost a necessity. A VTVM or a sensitive, 50-microamp meter is also needed. To tune the i-f amplifier and detector stages, remove the 6DC6 rf amplifier, and disable the mixer oscillator by removing B+ from pin 4 of the oscillator can, LCU-3B.

Tune the signal generator to 2.65 MHz, couple it to pin 2 of the 6U8A mixer through a capacitor and tune FL-1, FL-2 and FL-3 by observing a peak indication on test-plug pin 1. Switch the meter to test-plug pin 2 and tune the primary of FL-4 for peak indication. Finally, switch the meter to pin 3 on the test plug and adjust the secondary of FL-4 for zero indication. These adjustments have some interaction so you'll have to go through this procedure several times. Reconnect B+ to pin

4 of LCU-3B and replace the 6DC6 rf amplifier.

Set the oscillator frequency range 2.65 MHz higher than the received range. Tune the receiver to 21 MHz and adjust the oscillator coil slug until the oscillator frequency is 23.65 MHz. Set the dial to 30 MHz and adjust the trimmer capacitor, C1Dt for 32.65 MHz. Repeat these steps several times to insure proper tracking.

Put the meter on pin 1 of the test plug, set the rf signal generator at 21 MHz and plug it into the rf input jack. Tune the receiver to 21 MHz and adjust the coil slug of LCU-1 and LCU-2 for maximum indication on the meter. When tuning LCU-2, swamp the side of the coil not being tuned by connecting a .01 μF ceramic capacitor and a 1k resistor in series and attaching it across the coil you want to swamp.

Change the signal generator and receiver frequency to 30 MHz and peak the indication using the trimmer capacitors C1At, C1Bt and C1Ct. The adjustment of the coil slug and the trimmer capacitors at opposite ends of the tuning range will interact, so this procedure must be repeated several times to insure proper tracking.

summary

It is not my intent to present a rigid set of instructions for building this particular receiver, but it gives you an idea of my approach to the problem. By starting out with all the difficult mechanical work completed, you are way ahead in any receiver project. There are a lot of surplus receivers on the market at very reasonable prices, and when they're fixed up and modernized, they will give many years of good service before the transistors and IC's move in.

references

1. J. Kyle, K5JKX, "Rrceivers," 73, 1964

ham radio

aluminum's new face

Here are some helpful hints for cutting and finishing aluminum panels and chassis

J. B. Hood and E. L. Klein, W4BRS, 6814 Criner Road, S.E., Huntsville, Alabama 35802

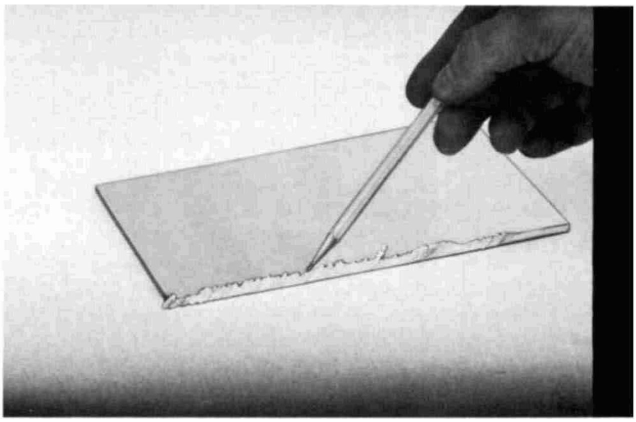
Are you proud of that new receiver or linear that you just built? Although it works well, its looks may leave a little to be desired! Painting aluminum panels flat black helps a little, but chips and scratches detract from the looks of an otherwise excellent product. Saw cuts frequently have to be smoothed with a file.

It used to take a lot of time to put a nice-looking finish on an aluminum panel, but with the simple techniques I discuss here you can get a professional-looking finish with minimum effort.

cutting aluminum

You can save a lot of time and trouble by cutting sheet aluminum with a table saw. Hold an old candle or stick of paraffin against the edge of the saw blade just before making a cut. The wax lubricates the blade and prevents galling of the work due to aluminum build-up. Lubrication is the single most significant technique for producing a smooth, accurate saw cut. Watch your fingers

Old Method. . . notice how the edge of the aluminum sheet has been galled by aluminum build-up on the saw blade.



-stop the blade before applying the wax!

When you're cutting aluminum sheet which is 1116th inch or thinner, it is a good idea to place a piece of masonite or plywood under the aluminum when running it through the saw. This supporting material is cut at the same time as the aluminum and the thin sheet is not bent down at the cut edge.

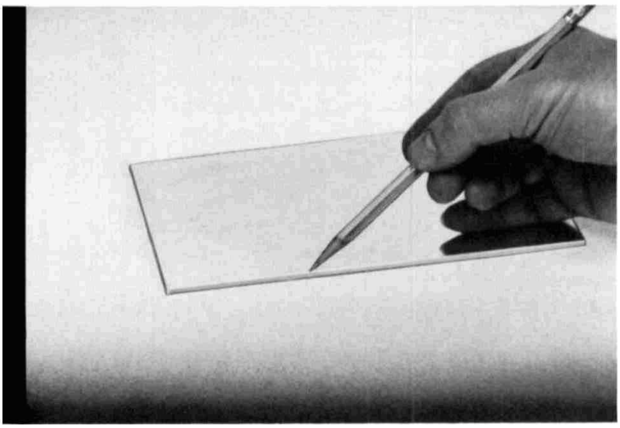
The type of blade used for sawing aluminum sheet is important. While a regular fine-tooth wood-type cross-cut blade works quite well, its life is limited somewhat. A non-ferrous metal-cutting blade is the best choice for cutting aluminum on a table saw. They are available from Sears for under three dollars.

Don't use coarse blades with tungsten-carbide inserts because they produce a very rough edge, especially on thin sheets. The same thing applies to hack saws and saber saws. A saber saw with a fine-tooth blade is very good for cutting thin aluminum, particularly when curves are desired. Also, if you use wax lubrication with these blades, very little, if any, filing will he necessarv after you make the cut.

surface finishes

Now, how about that surface finish? Would you like to improve it? A good paint job will go a long way toward improving the appearance of front panels. Aerosol spray cans are convenient to use and are available for primer paint as well as just about any type and color

New Method . . . smooth, clean edges by lubricating the saw with an old wax candle or parafin.



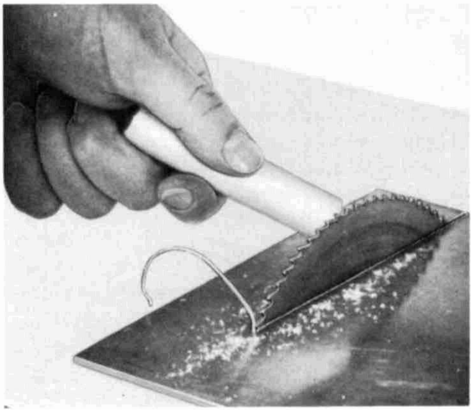
of finish coat. Many builders prefer flat black. After several coats are thoroughly dry, the flat-black surface should be rubbed with a soft cloth to provide a uniform sheen and minimize finger marks.

Bare surfaces, such as chassis and the backs of panels, can also be greatly improved in appearance after all the drilling is done. There are three very simple treatments for obtaining various surface effects—the brushed finish, orbital pattern and etched-satin finish. Each of them is very attractive on sheet aluminum.

brushed finish

This finish is very popular on professional equipment. It is commercially produced with

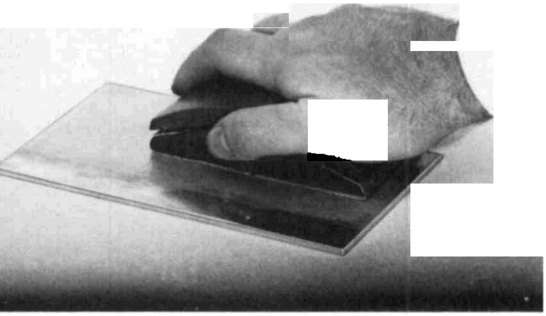
A used candle, held to the saw blade just prior to or during a cut, provides the needed lubrication. Danger—stop the saw before putting the wax on the blade!



a coarse, slow-turning wire brush; all scratch marks are parallel. All the amateur needs to simulate the brushed finish is a sanding block and a sheet of coarse sandpaper, preferably the type used for machine-sanding oak floors.

The coarse paper is moved back and forth in a straight line over the aluminum. A strip of wood clamped to the table as a guide makes sure that all the brush marks are straight. As the sanding progresses, the coarse grit should produce a series of tiny parallel grooves. Continue this back-and-forth movement until the entire surface has a uniform appearance.

You can simulate a brushed finish by using coarse sandpaper in straight parallel strokes.



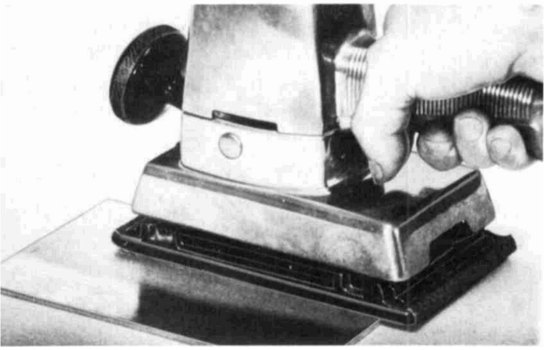
orbital pattern

For this effect, you nerd some fairly coarse sandpaper and an orbital electric sander. Steady the piece of aluminum and sand it while moving the orbital sander around in a circular motion until the surface has a uniform pattern. If the piece of aluminum has a badly scarred surface, begin sanding with a much coarser grit than you use for the final sanding; reduce the coarseness until the desired finish is procluced. For hcst results, use light pressure on the sander.

etched-satin finish

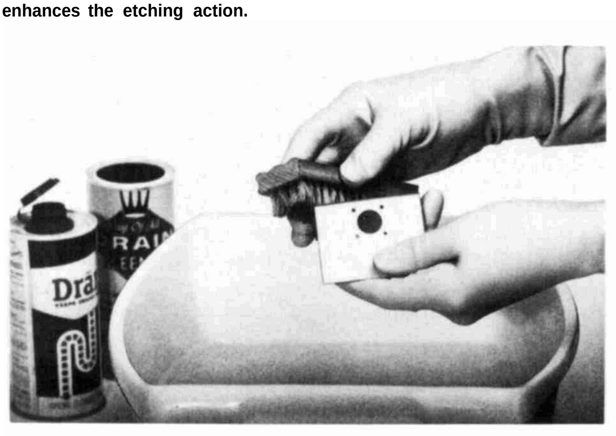
For this treatment, you need a large plastic container, such as a dish pan. You also need an alkaline chemical, a small scrub brush, a pair of rubber gloves and plenty of plain water.

An orbital surface pattern is obtained by using an electric orbital sander with medium-grade sandpaper. Light pressure produces the best cutting action.



The large container should be plastic. The alkaline chemical reacts with some metal pans and may damage the pan. Because of the fumes that are produced by this process, you should work outdoors or in a well-ventilated room. For the chemical solution, sodium hyrlroxide is mixed with plain water. If you can't find any sodium hydroxide, there are several brand-name household chemicals available that will do the job–Drano is one.

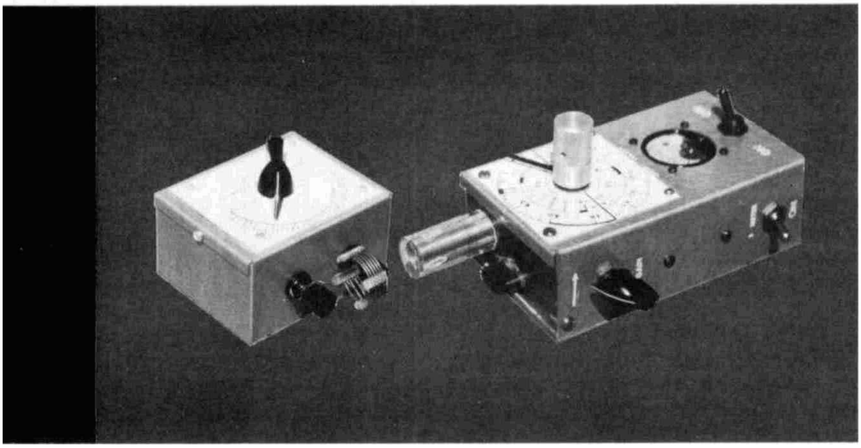
Any one of several common household caustics can be used for etching aluminum parts to provide a satin finish. The brush removes oxides and enhances the etching action.



Before putting the aluminum in the chemical bath, remove any deep scratches with sandpaper, working from coarse through successively finer grits. Mix the chemical and water in the plastic pan. When the sodium hydroxide is dissolved, dip the aluminum into the solution. Scrub it vigorously with a small bristle brush to remove the aluminum oxide film. In a few minutes, the solution will etch a very fine pattern into the aluminum. When the pattern is uniform, rinse the panel thoroughly in cold running water.

None of these techniques require any special equipment, nor do they require a great deal of experience. The next time you build a piece of equipment, try one of them–your hand-crafted equipment will be something you can be proud of.

ham radio



transistorized dipper

without plug-in coils

A handy instrument

for measuring frequency,

field strength,

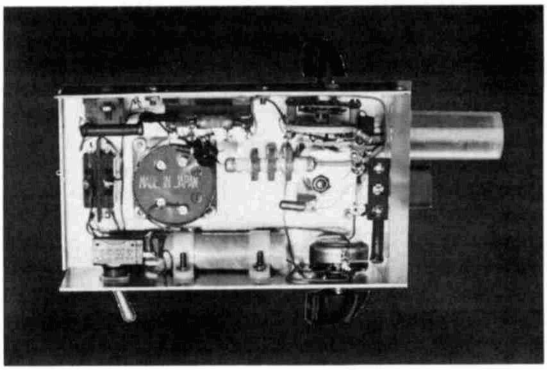
or checking crystals

Ed Marriner, W6BLZ, 528 Colima Street, La Jolla, California

It is often difficult to make transistor oscillators oscillate over wide frequency ranges. Dial calibration is also a real problem without expensive calibrating instruments and lettering sets. This article attempts to solve some of these problems, but you must use the same parts that I have, especially the coil and tuning capacitor.

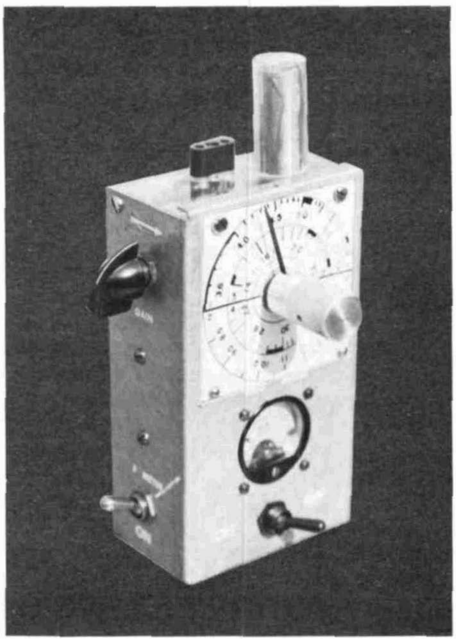
One of the joys of this dipper is the convenience of band switching. There are no plug-in coils to worry about. Plug-in coils are a source of trouble when they wobble around in the socket and cause loose connections. Resides, they are often difficult to push in and take out.

This particular instrument functions as a dipper, a field-strength meter ancl crystal os-



cillator/checker throughout the 3.5- to 30-MHz range. To make it do all of this properly, you must follow the circuit and layout very closely. A 2N1742 transistor was chosen for the oscillator because of its high frequency response, but other types, such as the HEP 2, GE-3, TRO6 or JR30X will also work. If you're only interested in the lower frequencies, the RCA SK3008 does a nice job.

The band-switching dipper may also be used as a field-strength meter or for checking crystals.



construction

The entire bandswitching dipper is housed in a miniature LMB 138 chassis, 6-1/4 x 3-1/2 x 2-1/8 inches. As you can see from the photographs, nothing is particularly crowded. The meter is an imported 0-1 mA unit. If you use the same parts and layout I did, you can obtain reasonable accuracy with the same dial calibration—I can furnish glossy prints of the dial I used for 35c. Put a piece of sixteenth-inch clear plastic over the dial to protect it.

In this circuit, the transistor oscillates easier above 6 MHz, so you have to switch in various values of feedback capacitance on bands one and two. I did quite a hit of experimenting to set the values for these capacitors and to determine the type of coil. I recommend that you use Air Dux 532T for the coil. I tried the next larger size, but the oscillator wouldn't work on the 3.5-MHz band. On hand four (22 to 40 MHz), the 2N1742 provides nearly full-scale deflection, while the SK3008 deflects the meter about one quarter. Almost any general-purpose audio NPN type will work in the meter-amplifier circuit.

When turning on the dipper, always put the sensitivity control to the off position to avoid pegging the meter needle; this is most critical on the 3.5-MHz hand. Notice the 25-pF silver-mica capacitor used on the 3.5-MHz switch position. This value was required for oscillation on this band. The next hand re-

fig. 1. Schematic of the dipper. On Band 1, the entire Air Dix coil is used. The 100-pF variable is a Cardwell ZU-100AS.

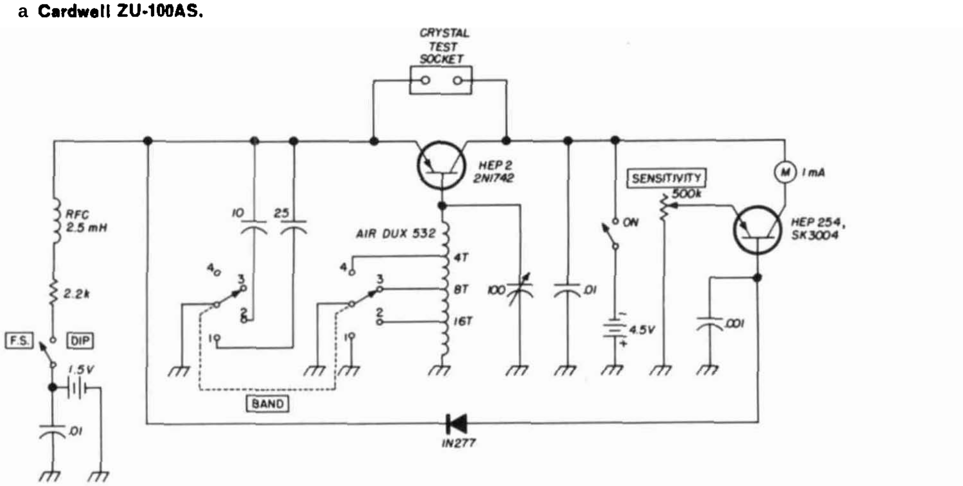
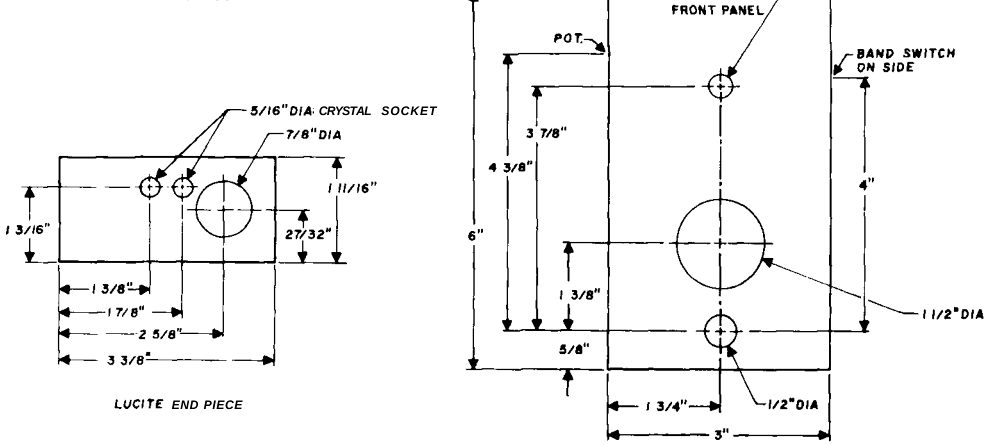


fig. 2. Mechanical details of the band-switching dipper.



quired 10 pF, and the rest of the bands didn't need any additional feedback.

The best way to check for oscillation is to listen to a receiver or by putting a VTVM probe on the emitter of the 2N1742. If a large positive voltage is indicated, the circuit is not oscillating. Check this on band two. The meter may still deflect upscale even if the circuit is not oscillating. The meter amplifier can be checked by feeding an audio or rf signal into the 1N277–large enough to drive the meter-amplifier transistor.

If you use mercury cells for power, be careful to get the polarity connected properly. Above all, don't forget the .01- and .001-μF bypass capacitors across the batteries; battery internal resistance may prevent oscillation. Ground the bottom of the coil; if it is left open, the stray capacity results in a tuned circuit, and the dipper will show a "false" dip.

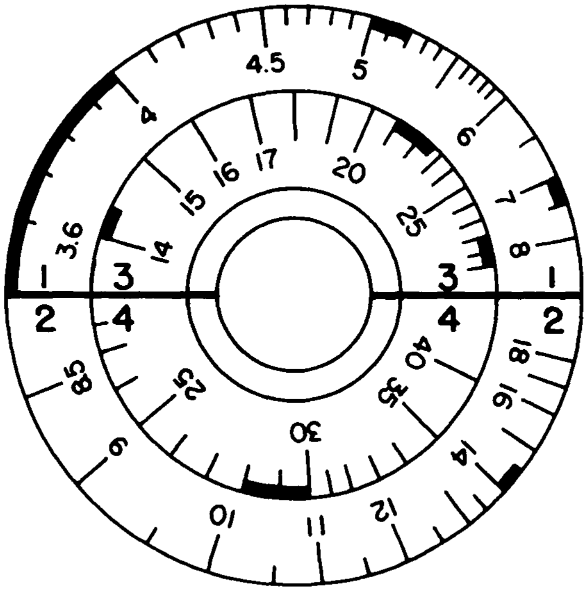
operation

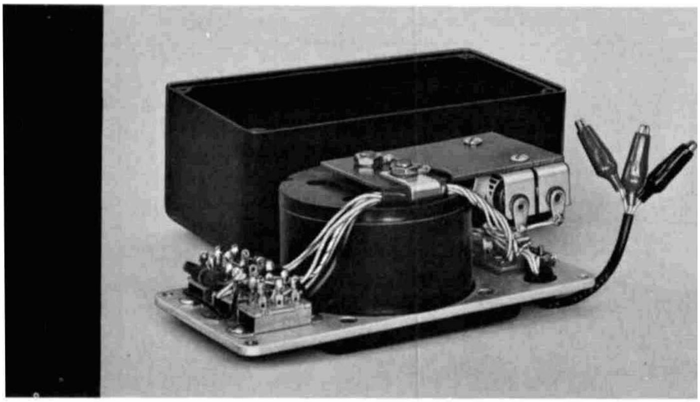
The single pole toggle switch in the battery lead is used for field-strength measurements. This switch converts the dipper into a field-strength meter for antenna measurements or transmitter tuning. When it is held close to an oscillator coil, it will indicate rf energy. The sensitivity control works in both positions–dipper or field strength. With the sensitivity turned up, the tank radiation from a 12-watt transmitter can be picked up three feet away.

In the dipper position, the dipper coil is placed near another coil and tuned through its range. When the dipper frequcncy is the same as the resonant frequency of the coil, the meter will dip to zero.

The distance you hold the dipper from a test coil depends on the selectivity of the coil and the frequcncy. For best results, use the minimum amount of sensitivity that is possible. Bandspread is sufficient, so a vernier dial is not necessary.

fig. 3. Full-size dial for the dipper.





Internal construction of the simple transistor checker.

simple, low-cost transistor tester for checking leakage and gain

The construction of this tester will be paid off by the use of low-cost, surplus transistors

L. Klein, W4RS, 68 Criti Rd., S.E., Huntsville, Alabama 35802

More amateurs and electronic experimenters are using transistors today than ever before. They are available at relatively low cost. Indeed, the bargains advertised in the back of this magazine make these versatile and compact devices universally attractive to builders who used to depend on tubes for a myriad of electronic applications. Today, most ham shacks boast a collection of transistors neatly stored in their original cartons; but what about transistors which have been used in a breadboard or the bargains which can't be identified?

The answer is a simple and easy-to-use transistor checker. When he looks at the published characteristics of transistors, the average amateur is overwhelmed by the abbreviations used by different manufacturers and by a maze of interrelations between the transistor's three elements–emitter, base and collector. Usually, the electronic experimenter is interested in only a few basic factors:

- Is it a PNP or NPN?
- What is the relative emitter-collector leakage?
- What is the approximate gain under normal conditions?

The transistor checker described in this article is intended to provide instant answers to these basic questions. When these questions are answered, the transistor circuit builder knows that his transistor is "good". However, in some critical applications, individual selection may be needed, particularly in high-frequency circuits.

circuit

You can see from the diagram in **fig. 1** that it's a simple circuit. Two sockets provide connections for the most commonly-used transistor configurations. As an additional convenience, a set of test leads lets you test transistors with other types of terminal arrangements. A double-pole, double-throw switch is included to reverse the emitter and collector terminals so that checker will take care of both PNP and NPN transistors. The 3-volt battery and 3.3k resistor provide emitter-to-collector current which is measured as "leakage" current on the 0-1 millammeter. By switching in either the 2.2k resistor or 102.2k resistor (2.2k plus 100k), a nominal bias voltage is applied to the base terminal. Now the milliammeter reads a current which is representative of "gain". If you want to, you can put in a third switch position to measure the battery condition; in this case, a suitable voltage-dividing network must be included An "off" switch is not used because no battery current can flow if a transistor isn't plugged in.

The "high-low" switch performs two functions. It shunts the meter and the 3.3k dropping resistor with a 15-ohm resistor. This permits a heavier current to flow through the transistor without deflecting the meter off-scale. Secondly, this switch selects a lower value of collector-to-base bias resistor which is common for power-type transistors.

construction.

I used a 5-3/4-inch bakelite instrument box for the case, but a mini-box or other home-made box will serve just as well. Because of its rigidity, 118-inch aluminum was used for the front panel. Since the 3-inch meter requires a hole which takes the greatest part of the panel width, this added rigidity is recommended. I used miniature slide switches due to their low cost, but regular toggle switches would be easier to mount in conventional round holes.

The two transistor sockets accommodate most transistors in use today. Smaller transistors of the TO-5 or TO-18 variety fit into the Elco 05-3304 socket, and power transistors of the TO-3 class are accommodated by the Cinch-Jones 2-TSI socket. In the latter case, the collector terminal is not insulated from the panel with a mica washer. This permits a more convenient collector connection when testing power transistors. Miniature mini-gator clips are soldered to the ends of the three short flexible leads. Different colored insulating boots assist in identifying the emitter, base and collector leads; Brady wire labels with a bold E, B and C complete the identification job.

A 1116-inch thick piece of phenolic, approximately 2 by 3 inches, is used t~ mount the dual battery holder (Keystone #140) to

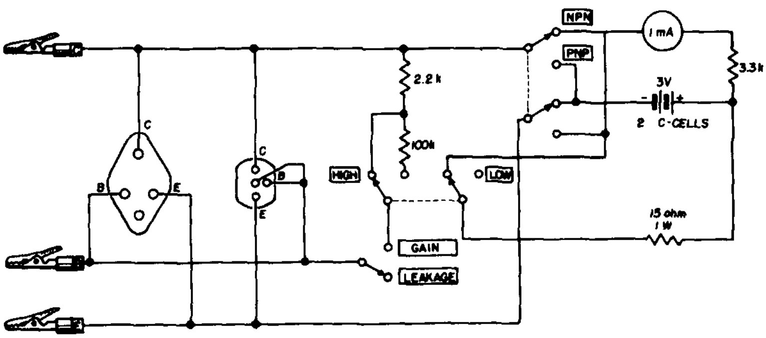


fig. 1. Schematic diagram of the transistor . fester. Mini-gator clips permit connections to transistors that do not fit the two standard sockets.

the meter terminals. The resistors used in the circuit are soldered directly to the switch lugs as shown in the photograph.

meter face

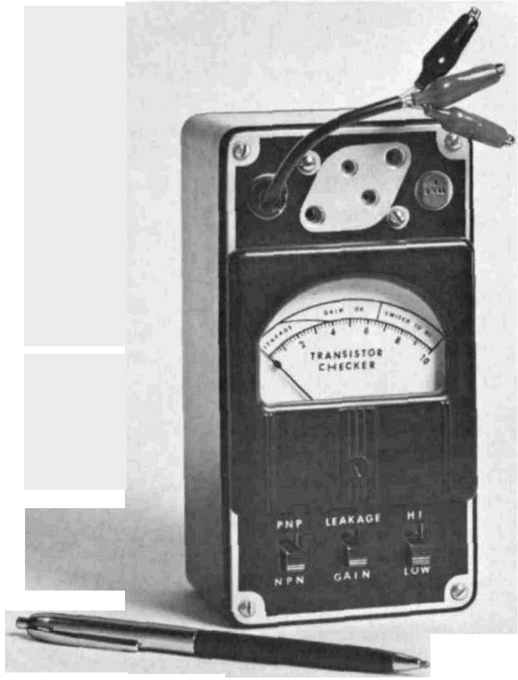
Most meters can be disassembled by removing three screws from the back. Then you can remove the enamelled-metal meter face. Don't bend the needle or distort the hair spring when removing the meter face. It's a good idea to turn the meter face over and cement the new scale on the reverse side. Then the original scale is preserved in case you want to re-use the meter in some future job.

Calibration of the new scale is not particularly critical. You can cut out fig. 2 and glue it on the metal plate or draw a new scale on good white paper stock with India ink. A very professional-looking scale can he made by first drawing the scale double-size and then reducing it photographically.

fig. 2. Full-scale layout of the new meter face for the transistor tester.

panel construction

The master drawing for the panel nameplate is shown in fig. 3A. The transistor tester illustrated in this article boasts a unique photo-chemically produced nameplate using Scotchcal from Minnesota Mining & Manufacturing Company. A proper technical dis-



The simple, low-cost transistor checker--ready to go to work.

cussion of this fascinating material would clemand a complete article. If you don't have access to professional nameplate fabricating facilities, you can use more conventional techniques for panel identifications such as decals or pressure-sensitive labels.

One word of advice–if a label is to be put on an aluminum panel, the surface should first be etched with Drano or otherwise worked to a uniform sheen with fine sandpaper or steel wool. A finish coat of clear Krylon will complete the job and give it long-term durability.

operation

When you're testing a transistor of unknown type or quality, simply plug it into one of the two sockets. The leakage-gain switch should first be set to "leakage" and the high-low switch is set to "low". A defective transistor usually has a very high emitter-collector junction resistance. Therefore, a bad transistor will seldom cause excessive current to he drawn through the meter.

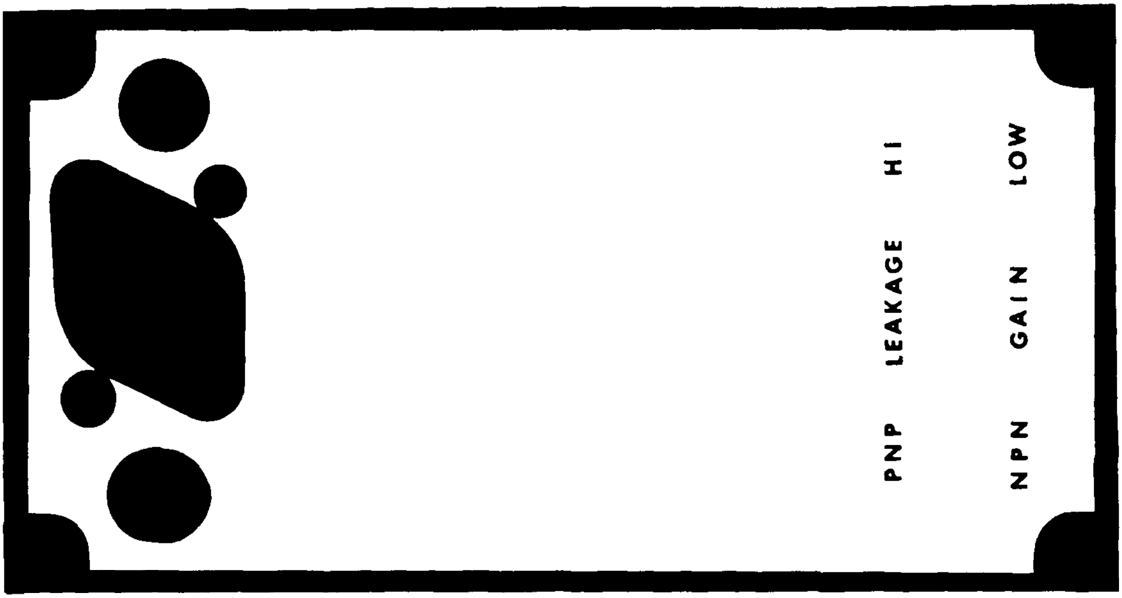


fig. 3. Full-scale photo master of the front panel for use with 3M Scotchcal. If you don't have access to photo-engraving, normal construction may be used with rub-on letters.

If little or no meter reading is evident, the transistor is either defective or the PNP-NPN switch is in the wrong position. Operate the switch; if a reading is now obtained, you have determined what type of transistor you have. If you still don't get a reading, the transistor is shot.

When switching from "leakage" to "gain", meter readings should fall into the "gain ok" area of the meter scale. The important thing to remember, however, is that with the same transistor, low leakage and high gain readings indicate merit. If the difference between the leakage and gain readings is small, you should consider the transistor to be below normal–a good candidate for the junk heap.

The absolute meter readings are not too meaningful if we wish to remain in the less technical domarn, which was the first condition of this article. However, the graduations from 0 to 100 were included on the new meter face as a convenience in recording relative values for different transistors within the same category or type number.

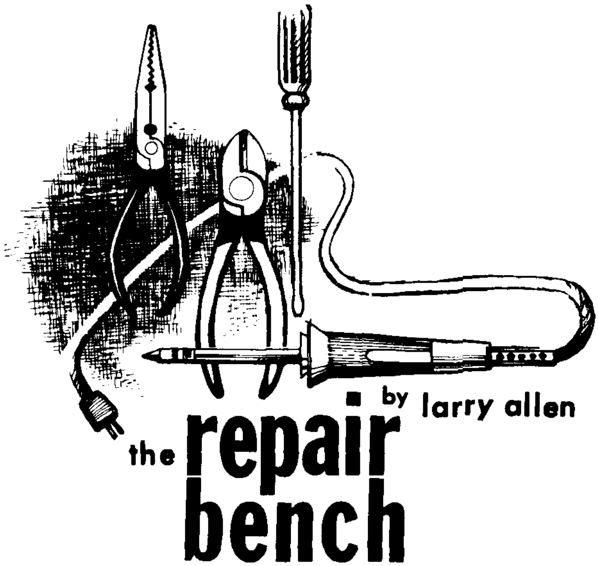
Generally, when power transistors are being tested, the "high" switch position is used. The "switch to high" notation on the meter lace reminds the operator to change the position of the high-low switch.

When your transistor checker is all completed and functions as it should, it will prove to be one of your most handy instruments around the shop–it certainly is at W4BRS

ham radio

next month in ham radio magaine:

Broad-Band Baluns
Single Sideband Oscillators
Amateur VHF FM Operation
Simple Milti-band Antennas
Return to Rack and Panel
An Amateur Aneometer
MOSFET Converter
Solid-State Transmitter Switching
Receiver Alignment
Plus much more!



signal injection in ham receivers

There's more than one way to skin a cat. And there's more than one way to track down trouble in a ham receiver in a hurry. Last month in this column I wrote about how to get your receiver back on the air fast by using a signal tracer. This month I'll explain another way–signal injection.

If you remember, the signal tracer lets you go through a set stage by stage, checking at each point for the signal. With signal injection, you don't even connect the antenna. Instead, you use a signal generator to put a signal into the receiver at various points, and you evaluate operation by whether or not the signal makes its way on through to the speaker. In most ways, signal injection is as effective as signal tracing. Often, it's even faster.

There are several different kinds of signals that can be used for injection. One is an audio-rate pulse produced by a very simple instrument called a harmonic generator. One example is a companion instrument to the pocket-type signal tracer I mentioned last month. It puts out sharply-spiked waveforms, generated at an audio frequency. The leading and trailing edges of the waveform are so sharp that the pulses are full of harmonics and can go through rf and i-f as well as audio stages. They can even shock-excite a tuned circuit into flywheel action. If you know the nature of the harmonic generator, you can use it throughout a receiver for signal injection. With a leftover transistor and a few parts, you can even build a little harmonic oscillator of your own.

Chief attractions of the small harmonic generator are simplicity and low cost. One alternative, which is more versatile yet more expensive, is the standard rf/af signal generator. With its audio and modulated rf outputs, you can inject exactly the correct signal for any stage in the receiver. The stability and accuracy don't matter unless you use the generator for alignment. For signal-injection troubleshooting, an inexpensive rf/af generator is adequate. The photo shows one of the better units, available in kit form. Several companies produce such instruments already wired. They range in price from $20 to well over $100, depending on accuracy and stability.

You can even use a square-wave generator, because of the sharp rise and fall of its waveform edges. You see, a square wave at 1000 kHz contains a lot of harmonics that extend Into the upper kilohertz and megahertz range. The result–and use–is similar to that described for the pulse-type harmonic generator.

the method

There are two acceptable ways to use signal injection for troubleshooting. One is the divide-and-conquer method described last month for the signal tracer. You start about half-way through the set. If your tests tell you the trouble is in the front half of the set, you divide that in half and check again. You keep this up until you have isolated the defective stage or circuit. If the fault is in the last half, you divide that... and so on.

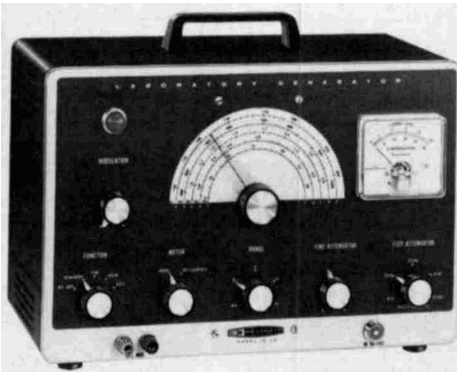
In the other system, you begin at the speaker and work your way back to the front end of the receiver. This is the method I prefer for signal injection, and it's the one I'll describe here. It's about as quick, and it has the added advantage of direct step-by-step logic. If you want to take shortcuts, you can easily skip a couple of stages, then back up if you find you've gone too far.

Turn the receiver on and let it warm up.

Turn the volume control to minimum, and disconnect the antenna. Set up the signal generator to give you only its audio output–not modulated, but audio alone. Connect the output lead of the generator directly across your speaker. The diagram shows this first test point as A. You may have to turn the output control of the gcnerator wide open to get enough signal to hear clearly, but if you hear a signal, the speakcr is probably okay. If there's no sound, the speaker is obviously bad.

Move the hot lead of the signal generator to point B, the plate of the audio output tube (or the collector of an output transistor). (Be sure there is a capacitor in series with the generator lead, so dc voltage won't burn out the attenuator in the generator; some instruments have a blocking capacitor built in.) You should still be able to hear audio signal in the speaker. What you're checking here is the output transformer. If you hear the sound, it is coupling energy to the speaker. If not, the transformer, or some connection between it and the speaker, is faulty.

A **popular** kit-type **signal generator.**



Next, move to the grid of the audio output tube (or the base of an output transistor)–point C. The sound you hear in the loudspeaker should now be very loud. The audio signal is going through the output tube, which should boost it plenty. Probably, you'll have to turn down the gcnerator to keep from rattling heck out of the speaker (or annoying anybody within earshot). If sound in the speaker is missing, or is not much louder than when the generator was connected to the plate, the output stage is at fault. You can make parts tests with the generator (they'll be described later) or use dc voltage tests to pinpoint the trouble.

While the generator is still connectecl to the output grid, turn the output level clown so that sound is just barely audible in the speaker. This will make it easier to evaluate the next test.

Connect the generator, still putting out an auclio signal, to the plate (or collector) of the audio stage preceding—point D. This still feeds the audio signal to the grid of the output stage, but does it through the coupling capacitor. If that capacitor is open, the signal will disappear. If it is okay, you'll hear the auclio signal in the rpeaker at about the same level you set it for before moving the connection.

Move the generator lead to the grid–point E. The signal you hear should be loud, because it now has the added amplification of the first audio stage. Again, you may have to turn down the output of the generator if the sound is too loud in the speakcr. If the sound isn't louder than before, the stage isn't doing its job. The tube may be bad, or an operating voltage may be at fault. Use whatever method you like to troubleshoot within the stage, but signal injection tells you whether the stagc is working or not.

If there are components between the audio-amplifier grid and the detectors of the set, you can move hack along those with the generator. None of them shoulcl reduce the output sound very much from the level at the first audio grid. Re sure you consider the volume control setting. You turned it down at the start, and you'll want to check its operation by injecting the audio signal at point F and running the slider up and down. With the volume control full up, the sound should be exactly as loud as it was with the generator at point E. The same goes for signals injected at points G, H, I, and J.

Next, check the diode detector. For this, you'll have to reset the signal generator to produce a modulated rf signal. For the first test, frequency of the generator isn't important, but you may as well set it to the i-f of the receiver. Clip the generator output to a point just preceding the diode detector. This may be across the secondary of the last i-f transformer, or at the plate of the i-f tube. In **fig. 1**, it's point K. Adjust the output of the signal generator for comfortable listening. Have the receiver audio gain (volume) turned up fairly high, but not high enough that you hear too much noise in the speaker with no signal injected. The setting of the rf gain control (if the receiver has one) doesn't matter at this point.

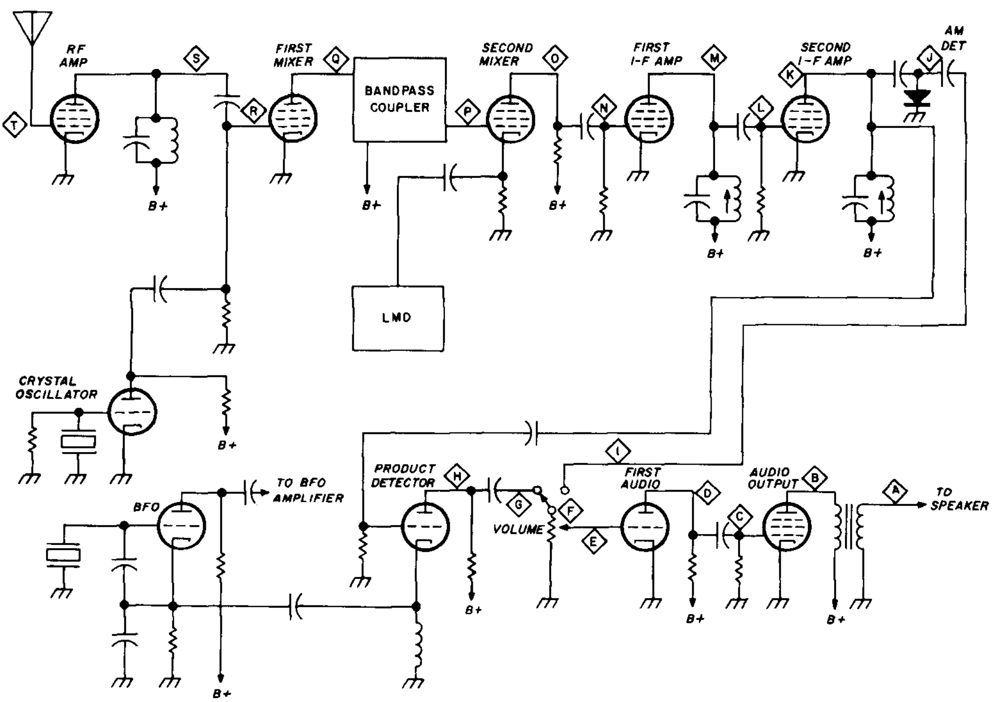
You can also, from this point on, stop using the loudspeaker as your indicator. A dc VTVM connected at the output of the detector diode makes a good one. Changes in signal level will show up as changes in the dc voltage at point J.

Adjust the signal level from the generator only high enough to cause a convenient, readable (or listenable) indication. Make sure the generator frequency dial is set for the i-f. Move the output lead hack to the primary of the last i-f transformer, which is at the plate of the tube. (Where impedance coupling is used, as in the diagram, the generator lead is already there; disregard this step.)

Signal indication should change very little, if at all. If it does, the i-f transformer is not coupling the signal through as it should. If you're using the dc VTVM as an indicator, reduced signal is indicated hy a lower dc voltage from the diode.

Before you move on hack to the grid of the i-f stage (point L), reduce the output level of the generator until you just barely hear a sound in the speaker. Then, with the signal generator shifted to the grid (or base), the signal should get a tremendous boost, and show a much Increased output from the diode or the speaker.

fig. 1. Schematic diagram of a typical communications receiver. The test points are discussed in the text.



By now you have gotten the idea of moving backward step by step through the stages. If the receiver is a multiple-conversion job, be sure to change the generator frequency as you jump from the plate of a mixer to the grid. In evaluating a mixer by this method, first make sure you have a good signal going through the rest of the set from the plate. Then, change frequencies and inject the signal at the grid.

If the oscillator is bad, you'll get little or no mixer output because, since you changed the generator's frequency, very little signal can get through the i-f's. If the oscillator and the mixer are working normally, you will get a signal output just about as strong from the grid as you got from the plate. If attenuation is noticeable, the mixer is inefficient.

Keep in mind that, in some sets, the stages preceding the second or third mixer are broadband i-f stages. This needn't be any problem. Just feed in a signal at any signal generator frequency that is within the pass-band of that particular i-f stage. **However,** remember that, if it's a stage that precedes the tuning section, you'll have to turn the station dial until the tunable oscillator is set to receive whatever frequency it is you are feeding into the i-f. It doesn't matter what the dial reads, since you're not checking calibration or anything like that; you just want to know if the stages are working. This is the way to tell.

Finally, you reach the antenna input–point T. If there is trouble in the set, somewhere along the way the signal will have grown distorted or disappeared. The trouble lies between the place where it was okay and the place it got bad.

parts analysis by injection

You've already seen how signal injection checks the interstage coupling components. If your generator is reasonably accurate, you can even try out i-f transformer adjustments, just to see if they're working. If you find a lot of signal attenuation through an i-f strip, it might be well either to align the set or have it aligned by someone who has the equipment and knows how. (Next month I'll explain how **you** can.)

If you recall how to use a signal tracer for checking bypass components, you have probably figured out that you can do the same thing with a signal generator. Only the method of application is different.

For example, to test a bypass capacitor by signal injection, set the generator for whatever frequency the capacitor is supposed to eliminate. Connect the output of the signal generator across the suspected bypass capacitor. Unless you turn the generator output awfully high, you should hear no output in the receiver's speaker.

Here's a good way to gauge how much signal to use in a test like this: assume a bypass capacitor at the screen of an i-f amplifier. First connect the signal generator to the control grid of that i-f amplifier. Turn up the output of the signal generator barely enough for you to hear a weak signal in the speaker with the af gain control fairly well up. If you're using a dc VTVM as the indicator, use just enough signal to cause a definite reading. Now, without disturbing the output setting, move the signal generator lead to the screen grid. Thrre should be virtually no output inclication, either audibly or on the VTVM. Practically all the signal should be bypassed to ground. If not, the capacitor is faulty.

That takes care of practically all the components in the set. The signal injection technique has already checked out all the transformers, tubes, and coupling components. About the only thing left is filter capacitors, and they're always checked best hy bridging a known good one across any suspected one.

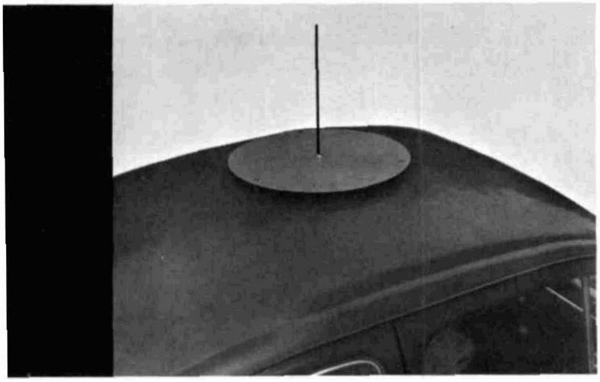
That's the story of signal injection. Next month I'll write about using the rf/af generator for alignment. Many of you have expressed the wish to align your own sets, and it really isn't too hard if you follow a few easy principles.

My mailbag was light this month, with little to pass along; I'm still waiting to hear from you. I'll be answering the best letters in this column.

ham radio

The **repair bench** is for you. Tell us about problems you have run into and solved, keeping your rig in peak shape. Questions you ask will be answered only if accompanied by a **copy** (not returned) of the full schematic diagram and a stamped, self-addressed #10 envelope. Larry will include some of the most interesting case histories in his column each month.

Editor



two-meter
mobile installations

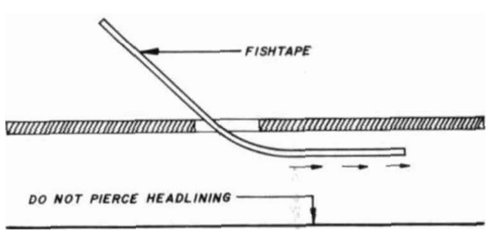
Here are some hints for two-meter mobile antennas, some without drilling a hole in the roof

E. H. Marriner, W6BLZ, 528 Colima Street, La Jolla, California 92037

There are many ways to mount a two-meter antenna on your car, and some are better than others. The best way is to punch a hole in the center of the roof, but not many hams have this sort of courage. Usually they look ro other methods. Another way is to hold a vertical ground plane on the roof with a magnet as shown in the photo, or to fashion an antenna from an aluminum disk and hold it on with suction cups.

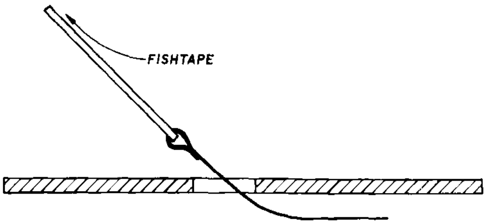
For those of you who **will** punch a hole in the car roof, there are several mobile antenna kits available for under $10.00. The Antenna Specialist model ASPR-1L or the Motorola Mobile rooftop model TU-316-1L are both popular.

fig. 1. Step 2.



The Motorola antenna comes with coaxial lead-in cable and all of the hardware for a complete installation. The cable is installed inside of the car between the upholstery and the metal car body. In some cars the coax can be pulled up to the top of the car by pushing a piece of stiff wire down between the body and the upholstery and fastening the coax to it. Installations may vary slightly with each car, but generally speaking, they are almost all the same.

fig. 2. Step 3.



planning the installation

The installation procedure following is for a two-door passenger car. First, select a location for the antenna as near to the center of the roof as possible. When you install the lead-in cable, it should be kept as short as possible to minimize rf loss. Running the coax from the antenna diagonally to the right or left side of the trunk compartment will keep it short. You can probe the headliner with your fingers to make sure that all obstructions are avoided.

installation procedure

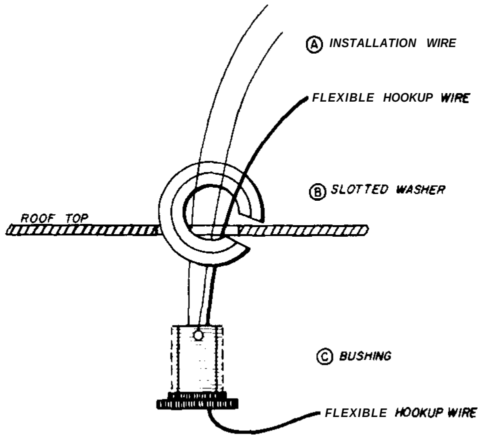
When you have all the tools you need, you can start.

1. Locate the desired position for the antenna, mark it with the center punch and drill a 3/4" hole. Use a sheet-metal drill or hole saw. Do **not use a standard twist drill.** Be careful not to tear the headliner. Scrape the paint off the top of the car roof about 1/16" around the hole to provide a good electrical bond between the antenna base assembly and the roof.

2. Carefully route the electrician's fishtape through the mounting hole in the roof top. In some cases it may be necessary to remove the window trim from around the left or right rear window, and loosen the headliner along the top of the window.

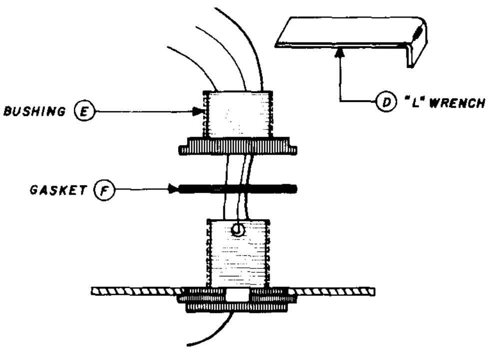
3. Attach a length of #22 or #24 flexible wire to the end of the fishtape and pull the tape back through the antenna mounting hole, leaving approximately 6" of flexible wire exposed at the mounting hole. Disconnect the fishtape.

fig. 3. Step 4.

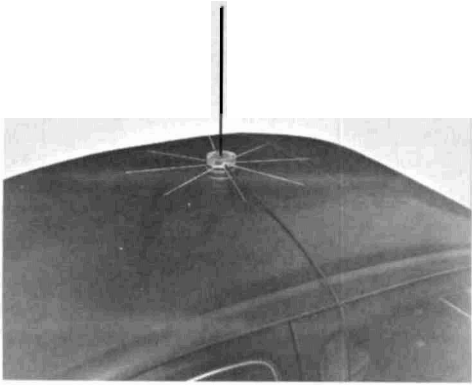


4. Attach the installation wire (A) to bushing (C), and pass the bushing over the end of the flexible wire and lower it through the mounting hole. Pass the slotted washer (B) over the installation wire and the flexible wire, keeping the shoulder of the washer

fig. 4. Step. 5.



This home made 2-meter ground plane is held on the car roof with a magnet.



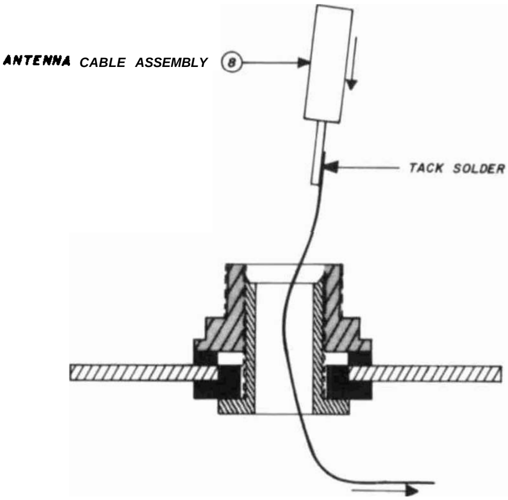
toward the top. Tilt the washer and pass the slot over the edge of the mounting hole. Push the washer through the mounting hole and onto bushing.

5. Put the gasket (F) (scam side downward) and bushing (E) over the end of the installation wire (A) and over the end of the "fish" wire. Pull the installation wire up until the washer is firmly seated in the mounting hole. Now, let the gasket drop into place over the bushing.

Put a little lacquer, glyptal or varnish on the threads of the bushing (E). Apply sparingly. Thread hushing E onto bushing C and tighten by hand.

Put the "L" wrench (D) in the slot in the top of the bushing and securely tighten it using a 718" open-end wrench. Take out the installation wire by breaking it in the center and pulling out the two pieces.

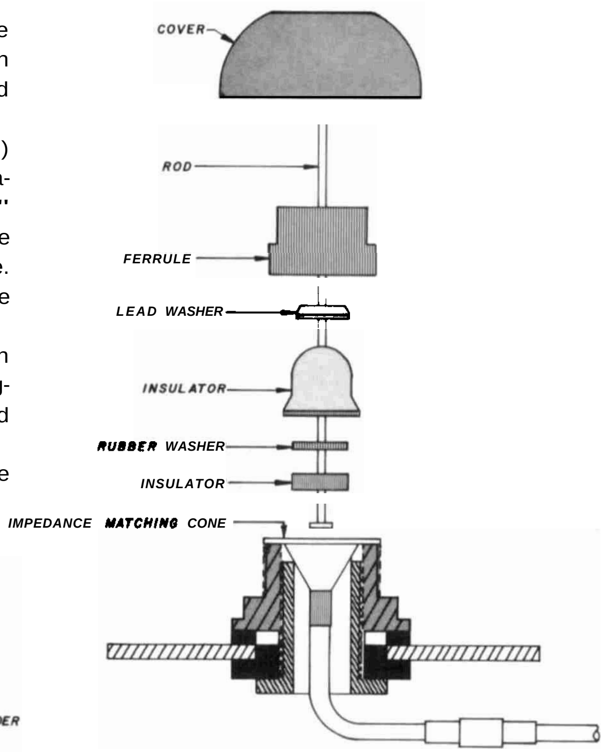
fig. 5. Step 6.



6. Tack solder the unconnected end of the antenna cable to the upper end of the fiexible wire and gently pull the lower end of this wire from the trunk or window frame until the curved tube on the end of the cable reaches the bushing (E). Work the curved tube downward onto the bushing until the impedance matching cone seats in the top.

7. Assemble the parts shown in fig. 6 onto the antenna rod in the order shown and pull the rod up until the lead washer, insulators and rubber washer are drawn completely into the ferrule. Now apply a small amount of lacquer, glyptal or varnish to the threads of the bushing. Make sure you don't get any inside.

fig. 6. Step 7.



8. Thread the ferrule onto the bushing. Tighten it by hand, then back it off one full turn. Gently pull on the lower end of the

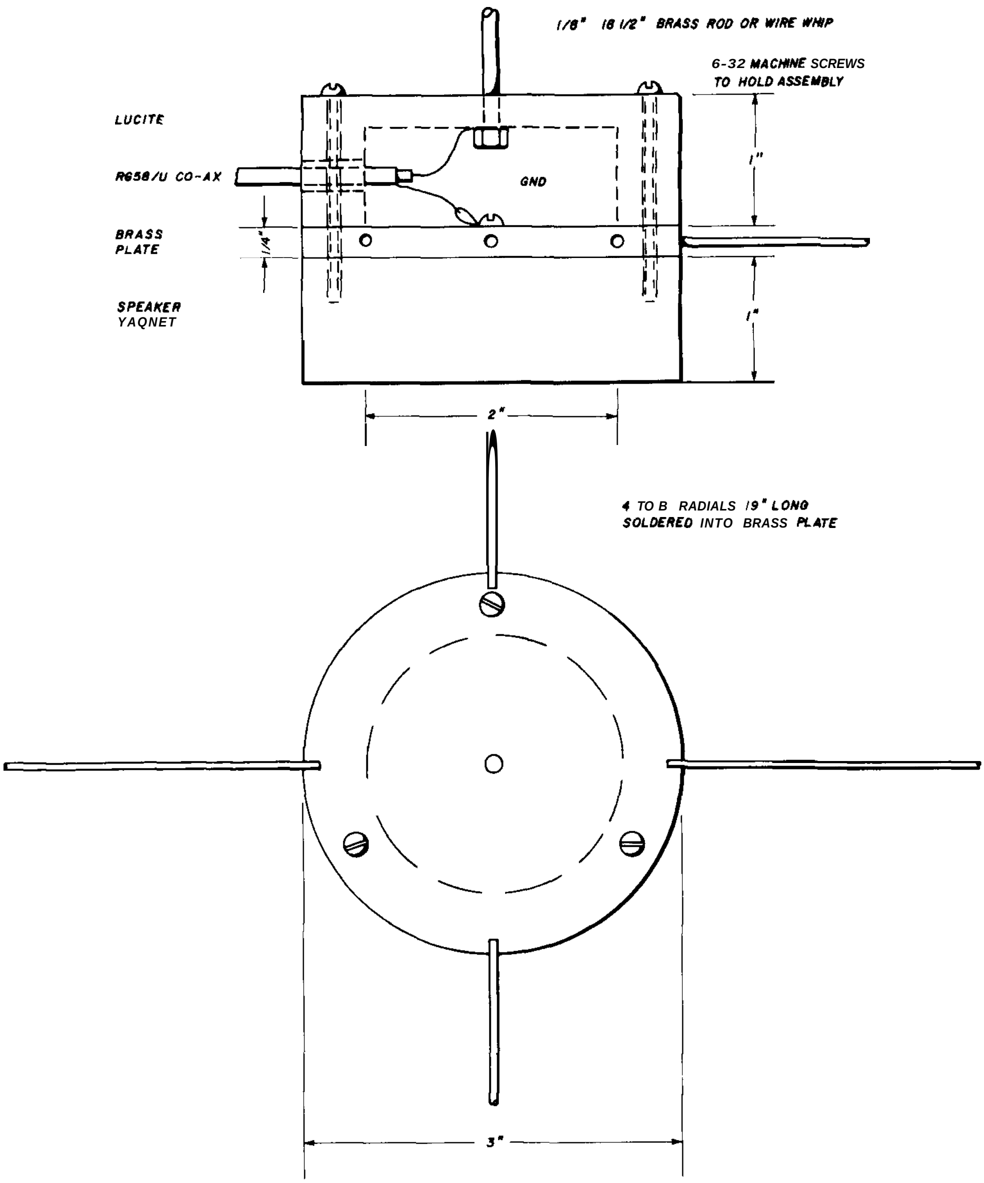


fig. 7. Holding a small 2-meter antenna on the car roof with a permanent magnet

lead-in cable. This will cause the curved tube to swing into line with the lead-in, minimizing strain at the junction of the tube and cable. Tighten the ferrule using a 3/4" open-end wrench. Be careful. Don't tighten the ferrule too much—the antenna insulator will crack or break. Now put the ferrule cover in place.

9. If it has been necessary to remove the window trim, route the electrician's tape from the trunk compartment to the place where the antenna cable sticks out and attach the antenna cable to the fishtape; pull the tape back into the trunk compartment and carefully pull all excess cable into the trunk compartment. Cut the cable to the shortest

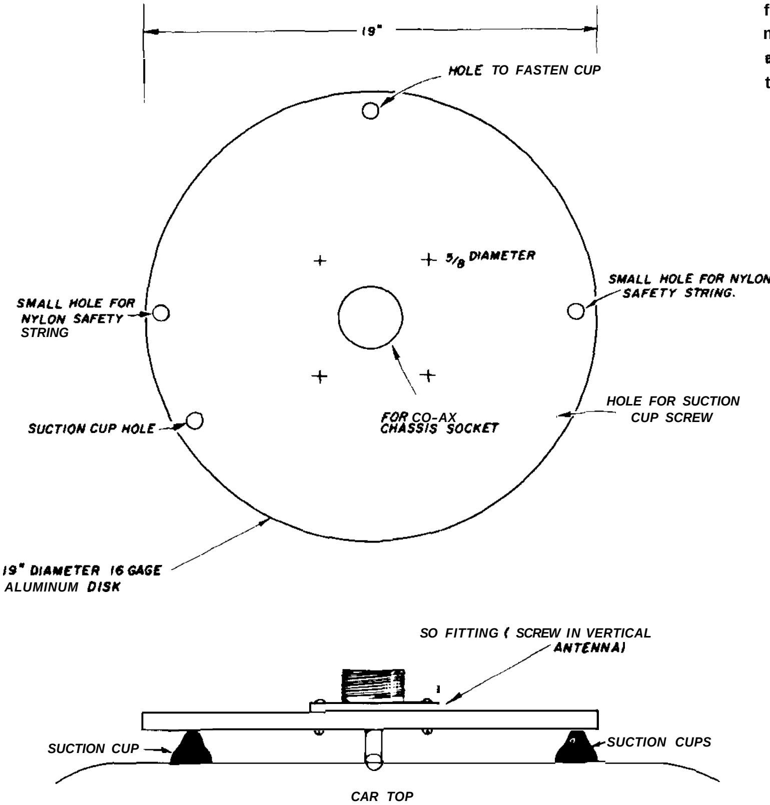


fig. 8. Two-meter mobile ground-plane antenna using suction cup mounting.

practical length that will reach the coaxial connector on your transmitter.

mobile antennas without holes

Now, how about the fellow who doesn't want to punch a hole in the car roof? Well, you can buy a commercial antenna such as the Hy-Gain mobile HH2BA with a HMBA mast, or you can build your own mobile antenna. If you build your own antenna, it can be fastened on the car roof with a permanent magnet or suction cups from the local auto parts store.

Fig. 7 shows another mobile ground-plane antenna. This one is held on the car roof with a magnet from an old loudspeaker. A brass plate three inches in diameter is screwed to this magnet. The ground radials are soldered to the brass plate; four to eight are enough.

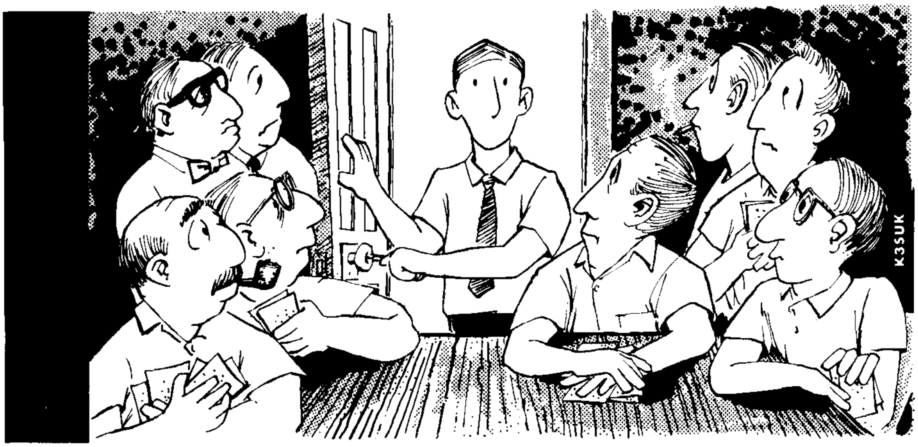
Fig. *8* shows how an antenna can be constructed from an aluminum disk and held on the car roof with suction cups. The same vertical which is used for a portable station can be used for the vertical part of the antenna or you can use a 19" piece of #12 wire.

ham radio

short circuit

There is an error in the schematic diagram of the discrete-component power supply on page 31 of the March issue. The supply will not regulate when connected as shown. The base of the 2N3053 regulator should be connected to the collector of the right-hand 2N3641, the base of the 2N4037 regulator to the collector of the right-hand 2N3644, **not** between the 5.1k and I k voltage-dividing resistors as shown. Thanks to W1OOP and W6GXN for bringing this to our attention.

Editor



the secret society

What will you do when you confirm over 300 DXCC countries?

Dr. J. Michael Blasi, W4NXD, 1490 Enota Avenue, N. W., Gainesville, Georgia 30501

Several months ago one of my very good amateur friends was passing through my section of the country. Since the welcome mat is always out, he didn't hesitate to stop by for a bit of Southern hospitality.

Jim works for an electronics firm and travels around the world, spending anywhere from a day to three months in any one place before he moves on. This brings him in touch with a great many amateurs, and he'll usually have an interesting story or two if I can get him going on the subject.

You all know how important DX has become in the last few years–DXpeditions, money for QSLs, pileups and publicity. When the DX fever hits, it's worse than the Asian flu! If there is a DX club in the area where Jim is staying, he gets in touch with some of the prominent members and attends a few meetings.

This story happened in Southern California. This makes everything very believeable, because you know how things are out there as far as ham radio is concerned. One of the clubs Jim visited had some of the top DX'ers in the world. There were at least eight fellows who had confirmed over 300 countries. Naturally this had no small effect on their personality; their noses tilted a bit upward, among other things.

He was supposed to be in that part of the country for a few months, so the club loaned him a key to the club house so he could operate the station. One afternoon he managed to have a hit of free time, so hc dropped hy to see if 20 meters was open. He didn't expect to see anyone there, so he was a bit surprised to find the eight club members who had more than 300 countries. They weren't using the rig–just sitting around the meeting table. They stopped talking as zoon as he came into the room.

Jim didn't give it much thought at the time, but about five days later, the same thing happened. The third time it happened he was sure that something was going on that he wasn't supposed to know about. They usually left as soon as he arrived, hut on the third visit Jim found something on the floor—a QSL card. After a few minutes, when he finally realized what was going on, he had a good laugh.

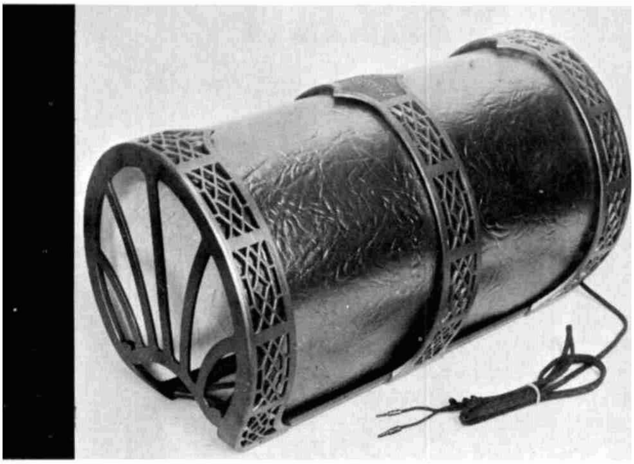
The next week he found them all there and mentioned the QSL card. He was right, they did have something cooking among themselves. The card was from a commercial station in South America thanking one of the hig DX'ers for his SWL report!

It seems that these fellows had just about worked every country in existence and just sat around the shack waiting for something to happen. One of them made a little one-tube superregenerative receiver and started to log foreign broadcast stations. Then he sent a report to them. He got more of a kick out of that SWL card than anything he had done in the past five years.

Before long, he and his select group of DX buddies were seeing who could get the most SWL cards using little one-lithe receivers. Naturally, they couldn't let this get out or they'd he the laughing stock of the club. The little meetings they had were to compare their new SWL cardz; Jim just happened to be there.

Jim and I can't tell you who these fellows are, and I'm sure they're not going to talk about it on 20 meters, hut if you want a few laughs, the next time you work one of the big time DX'ers in WG-land, ask how Radio Peking is coming through these days!

ham radio



what is it?

No,

it isn't an antique

footwarmer for

cold

ham

shacks

Ted Woolner WA1ABP, 30 Cedar Road, Shrewsbury, Massachusetts 01545

On November 3, 1925, Dr. Lee DeForest was granted a patent for a sound-reproducing device. At that time he was president of the Audalion Company, manufacturer of the speaker shown in the photograph. If you look closely, you'll see the name "Audalion" engraved on the center of the frame.

I got this speaker in a trade with a fellow member of the Antique Wireless Association who lives in Kentucky. He even threw in two bottles of "mountain tanglefoot" to boot! It's just as much a conversation piece as the speaker itself.

Since there are felt pads placed on the end and on the flat side, the speaker may be used in either the vertical or horizontal position. It is nineteen inches long, ten inches in diameter and uses a magnetic speaker unit. The coils were both open and I had to rewind them by hand. Most of the other parts of the speaker are made from white metal. Since it was 42 years old when I got it, you can imagine how much of the white metal was left in one piece. But, the best part of the whole project was, when all the repairs and replacements had been made, darned if it didn't work!

ham radio

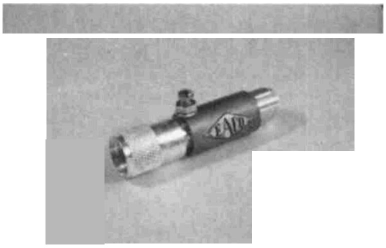
new products

DA Keyer

Omega Electronics has recently announced a new solid-state automatic keyer using integrated circuits, transistors and diodes, each where it will do the job best. It features "iambimatic" action (squeeze causes alternate dots and dashes), dot and dash memories, reliable long-life reed relay with heavy-duty 50-watt contacts and complete with built-in paddle. For maximum economy and low battery drain, milliwatt integrated circuits are used. Carbon batteries will last over fifty hours, or if preferred, the DA-3 option will provide operation from a 6.3-Vac supply.

The DA Keyer speed is continuously adjustable from six to sixty words per minute. The volume and tone controls for the built-in monitor are located on the rear panel. Jacks for an external paddle or straight key are also provided on the back. $85 from the Omega Electronics Company, 10463 Roselle Street, San Diego, California 92121.

EACO Lightning Arrestors



A novel ring gap is a feature of the new EACO Model *210* coaxial lightning arrestors. This type of gap is said to be self-correcting after a discharge of the type which would normally require replacement of point-type gaps. Zero insertion loss is claimed in a 50-ohm line, and the change in SWR cannot be measured at 150 MHz. The standard version has SO-239-type fittings, although other types are available. The Model *217* offers a loss-less electronic static circuit, but is limited to 50 watts of a-m power or 100 watts of ssb.

The Model *210* is priced at $3.15 while the Model *211* is $4.65. Available from the Electronic Applications Company, Route 46, Pine Brook, New Jersey 07058.

Design Industries Operating Desk

A new entry from Design Industries will help to solve the ever-present problem of trying to keep an XYL and a ham radio station under the same roof. The Senator I and Senator *II* series of operating consoles have been designed to accommodate many of the popular receiver-transmitter and transceiver combinations. Room is also available for various linear-amplifier combinations. If you have need for a custom-tailored installation, it can usually be worked into the Senator; the manufacturer will work with you to assure satisfaction.

Features include carpet rollers, an oil-dark walnut finish and removable front panel. The sides include storage and power-supply compartments and two letter-sized drawers. These consoles are designed so that when later equipment changes are made, it is only necessary to purchase a new front panel. The Senator I is $295.00 and the Senator II is $395.00 For further information write Design Industries, Inc., P. O. Box 19406, Dallas, Texas 75219.

Raytrak AutoLevel

An interesting new volume compressor has recently been made available by the Raytrak Company. This device is compatible with ssb exciters in that it is not a clipping-type compressor. It uses a small incandescent lamp which is optically coupled to a photoconductor (a light sensitive resistor) which regulates the output from the compressor. This approach allows a minimum of waveform distortion and insures a clean signal.

The *AutoLevel* Volume Compressor can provide up to 28-dB of compression. It is designed for use with either dynamic or crystal microphones and will operate on 115-volt ac power. Standard jacks are used and wiring permits push-to-talk operation through the unit. The price is $87.50. The unit can be purchased from the Raytrak Company, 2111 Springhill Drive, Columbus, Ohio 43221.

comments

Dear Jim:

I was going to write a book review type of article for the club paper; however, a good many copies of **ham radio** have found their way into the hands of our members. I haven't heard anyone knock it yet! This changed my mind about writing a review. Jim, if you can introduce a new publication into the ham-radio field and not have it lacerated, you and yours have a very bright future.

Most of your readers will admire the makeup and technical excellence of the magazine without realizing the endless thought and effort that went into your midwifery.

The content of the first issue is beyond reproach; you've set high standards for the future. Of almost equal importance however is the technical excellence of the magazine's production. One seldom sees better typography or printing. Your choice of a gothic type face is commendable and is especially appreciated by bi-focal wearers. Lastly, the schematics, technical drawings and cuts are beautifully sharp and crisp.

Congratulations, Jim, to you and your associates at **ham radio.**

Ed Howison, WA8AXB
Editor, CARASCOPE
Columbus Amateur
Radio Association

Dear HR:

Just received a copy of volume 1, number 1, and want to say how impressed I was—it has that fresh, clean look that's so hard to achieve in a new publication; and the layout, makeup and presentation are superb. Obviously, too, you are embarking upon a fresh, new approach to ham radio which has been needed for some time, and a standard of writing which should attract some competent professionals.

By the way, I would appreciate it if you could tell me just what type faces you use for the text and heads—I am an amateur printer as well as a radio ham. They are wonderfully modern and easy to read.

Don Holm, W7PFL
Outdoor Editor,
The Oregonian

The text is set in 8-point Optima; the captions in 6-point Akzidenz Grotesk; and the heads, 14, 18- and 24-point Standard.

Dear HR:

Ham radio has a very refreshing and needed approach — a good technical magazine. Best of luck. . .

H. W. Brown, W2OQN

Dear HR:

. . . Your approach to modern equipment is excellent. I have never understood ssb at all til I read the first issue and the excellent article on ssb theory.

William E. Harris, W5TVN

Dear **HR:**

If the next 12 issues are as interesting and refreshing as your first, you will be making a most creditable contribution to the published word on amateur radio state of the art...

John R. Esterly, **W8RAK**

Dear **HR:**

Congratulations! Your first issue is number one! Very fine articles, very well laid out, and very much needed. I hope that you will prosper in your labors and that the periods of difficulty will not weigh you down.

James R. Belt, Jr.

Dear **HR:**

...first issue looks great, keep up the good work.

George A. Wilson, Jr., **W1OCP**

Dear **HR:**

I just received your magazine in the mail today, and am writing this letter in haste. I am awaiting my General License and I am very anxious to "homebrew" my own ssb transmitter; in fact, all my gear. Mostly because I don't have the loot for an expensive appliance and because I would like to be able to say, "the rig here is homebrew," but unfortunately, I cannot find a reasonably presented article in any of the magazines on the market today—including yours. This is really frustrating. I guess I am going to give up looking.

When I saw the cover of your magazine I suddenly became excited when I saw the home-brew ssb exciter. The caption under the picture stated, "complete construction details"...you state that no tuning or alignment instructions are given, wow! ... You must present the article in kit form with available templates at a small charge and parts lists, "available parts of course," and also, large photographs and total presentation in easy-to-follow kit form...

Kenneth D. Brown, **WNØRXK**

If we were to publish the ssb-exciter article *in* kit-form style, it would require the entire issue of the magazine. Obviously, this is impractical.

Dear Jim,

I wondered where you had wandered. The answer is wonderful. Thank you very much for the sample copy of **ham radio.** How thorough and explicit your articles are. I couldn't wait to write a check and forward it to you for a year's subscription. Best of luck and success.

Gerald "Doc" Feinberg, W1TXL

Dear HR:

Congrats on the first issue of the new magazine. It's a real gem. It has opened the door for the return of ham technology long since gone from the scene. Heaven help the competition!

My best wishes for all the success in the future.

Robert W. Stankus, K2DX

Dear HR:

Thank you for copy of **ham radio** volume 1, number 1. The format is fine and I like the editorial philosophy.

I happen to hold original issue of **QST** December, 1915 and several others of the first year's issues.

W. A. F. Pyle, W3WR

Dear HR:

We took a swift glance through the new magazine and had to send you a little note of encouragement.

Will you have anything about RTTY? I am very interested in obtaining a printing model that can also be used to transmit. Any info concerning this phase of amateur radio would be greatly appreciated.

The sideband rig looks very interesting.

Danel P. Keech, KIFKX

How about *it RTTY* authors? *Let's have some* good *RTTY articles–my* cupboard *is bare!*

Dear HR:

Congratulations. The first issue is a real winner. Very glad to see you did not continue any of the articles on back pages...

Tom Lamb, K8ERV

Dear HR:

. . . I am most impressed with your first issue––the content is very good, but the type style and layout of the magazine is far superior to any other publication in the amateur-radio field—and, to most professional journals, I might add.

Douglas E. Westover, K6TZX

Dear HR:

Congratulations on the first issue of **ham radio.** Its format and layout are most attractive, the illustrations are of excellent quality and the printing is top grade. The contents are sufficiently diversified to provide something of interest to almost everyone. I have spent many years in the photographic illustration and advertising field, and I am sure that you are justifiably proud of your first issue. If the editorial policies of **ham radio** maintain the same high standards as its physical appearance, I believe its success is assured.

W. G. Blankenship, Jr., WA4GNW

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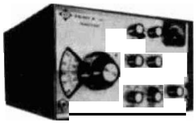

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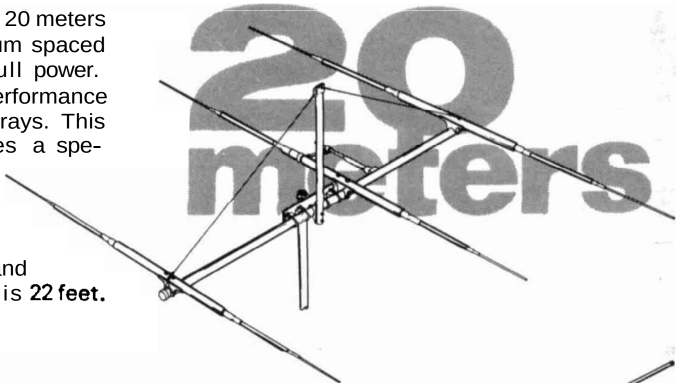

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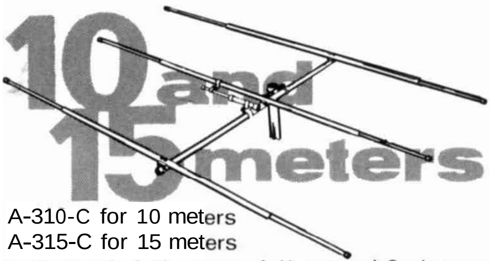

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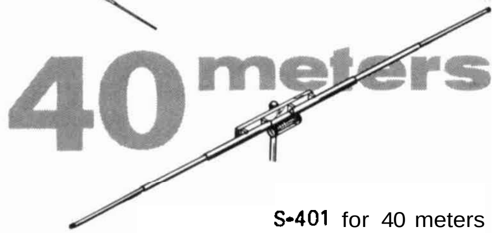

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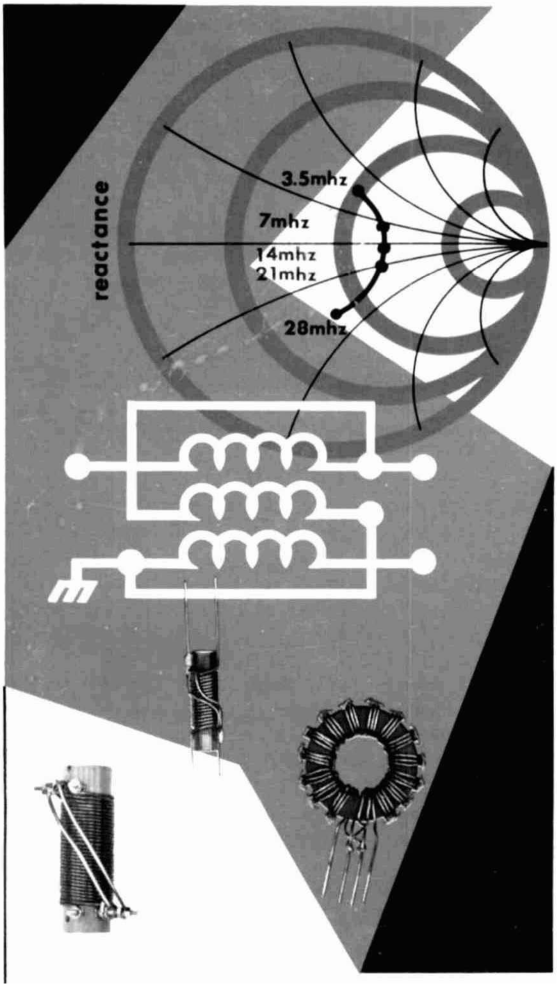
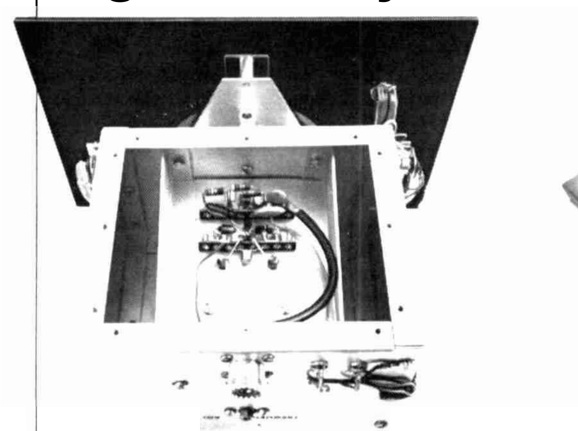


broadband antenna baluns 6

this month

Solid-State Projects for the shack

Build this high stability VFO

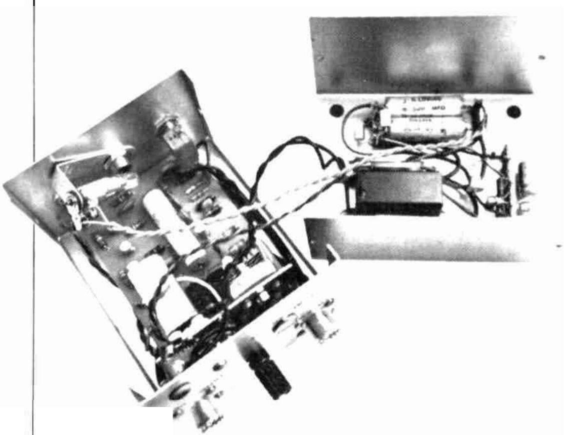


Here's an almost drift-free VFO built around the RCA-3N128 MOS/FET for flexible operation. After just 30 seconds warm-up, it tests out at less than 30 cycles drift in a two hour period.

Look in The Radio Amateur's Handbook, 1968 edition or write to RCA, Commercial Engineering, Sec. F175SD, Harrison, N.J. 07029 for full design details, including parts list, schematic, and building tips.

All listed RCA devices are available from your RCA Industrial Semiconductor Distributor

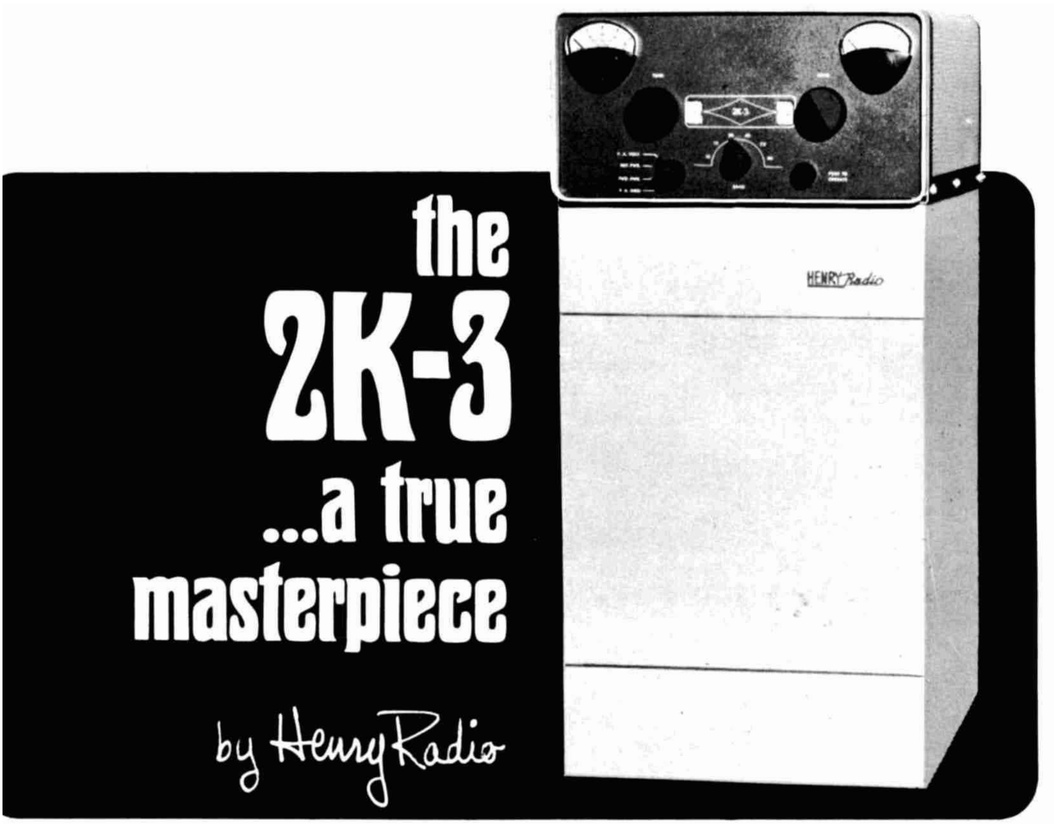
Build this VFO calibrator



If you're interested in MARS and have just a "ham-bands-only" receiver, this may be your answer to VFO calibration outside normal bands. It uses two RCA-1N3193 rectifiers; two 1N34A signal diodes; one RCA-2N2614 and seven RCA-2N3241A transistors—provides calibrating beats at 100 kHz points as well as 50, 33, 25 and 20 kHz. Handy, too, for calibrating test equipment.

Look in August 1967 QST or write RCA, Commercial Engineering, Sec. F175SD, Harrison, New Jersey 07029 for August 1967 "Ham Tips." RCA Electronic Components, Harrison, New Jersey.

RCA



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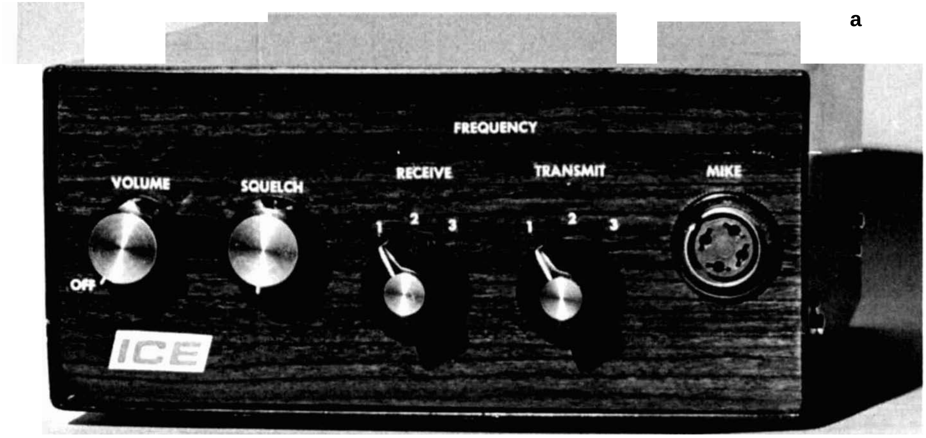

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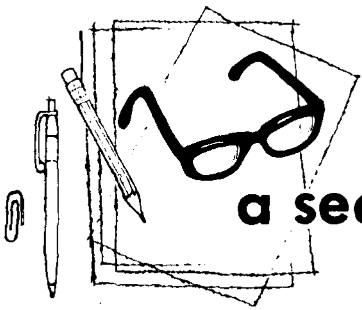

a second look

by jim fisk

Did you ever wonder why we are called hams? I suppose we all have at one time or another. We'll probably never know where the term really started, but there have been a lot of stories down through the years.

The first story I heard about "hams" related how the British amateurs referred to themselves as "am's." Because of the British pronunciation of the word, the Americans immediately picked it up as "ham." Since there were no other conflicting stories regarding the derivation of the word, I accepted it at face value.

However, in looking back over the story, this would mean that "ham" did not appear until radiotelephony became popular. Although old timers can't definitely remember the use of the word before the advent of radiotelephone, they're not sure exactly when it came into general use.

A story which was pretty widely circulated a couple of years ago indicated that H.A.M. was derived from the first letter of the last names of three prominent hams in the Boston area around 1910. This story seems to be concocted and apparently has no basis in fact.

The latest story I've heard tells of the derivation of the word ham as used in the theatre and how it was applied to the amateur radio operator. It seems that in the late 1880's there was a popular minstrel show on Broadway put on by one of the greatest blackface teams of all time, Heath and McIntyre. Their show was called "The Ham Tree." This show ran for a long time and eventually the word ham was applied to all minstrel players. Later on, any person who was not in the professional theatre was called a ham.

As many of you may remember, broadcasting on the ham bands was perfectly legitimate in the early days of amateur radio telephone. In fact, many of our famous radio personalities got their start by broadcasting on the amateur bands. They put on news shows, sports shows, music shows, and general-interest entertainment. These people were called hams, probably quite properly, and through usage, all amateur radio operators became hams.

Of all the stories I've heard, this one seems the most probable. I'm sure that there are a lot of other stories relating how the word came into use. If you have any stories along these lines, I'd be very glad to hear about them. They are just as much a part of ham lore as the story of the "wouf hong," the "retty snitch" or the "old man." These stories are a lot like the lumberjacks' Paul Bunyon. Some of them are pretty far-fetched, but they are always good for a few laughs. Hams we are, and hams we'll stay, but we'll probably never know the true story of the word.

Jim Fisk, W1DTY
Editor

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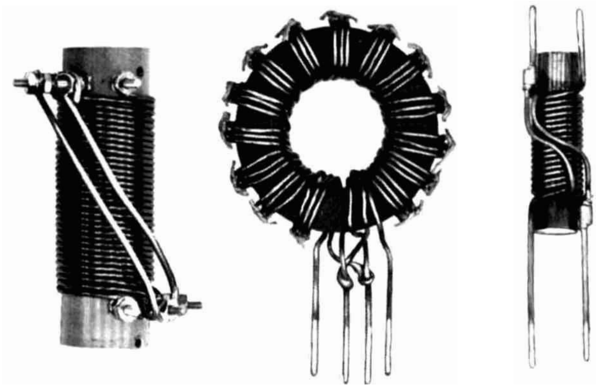


Three wideband baluns which **are** suitable for use in the hf spectrum. To the **left** is a simple **air-core** balun wound on a plastic form: in the middle is a toroid-core balun; to the right is a ferrite-slug **balun.**

broadband antenna baluns

A comparison of the impedance characteristics and power-handling capability of three popular types of broadband 1:1 baluns

William I. Orr, W6SAI, 48 Campbell Lane, Menlo Park, California

A balun is an electrical transformer for converting a balanced system to an unbalanced system or vice versa. They come in all shapes and sizes from the midget "ladder transformers" for television receivers to giant ferrite transformers for multi-kilowatt broadcast stations. The baluns that interest the radio amateur are used to match unbalanced coaxial transmission lines to balanced antenna systems.

An inexpensive narrow-band balun may be made with a quarter- or half-wavelength of coaxial line as shown in fig. 1. These baluns will cover a single amateur band, but they are of little practical interest if you're using a three-band beam antenna or multi-band dipole.

Various forms of ferrite toroid baluns that promise broadband operation have appeared recently,[1] as well as a simple broadband coaxial-wound balun.[2] This article discusses some of the more common ferrite balun transformer designs and describes a simple and inexpensive air-core balun that works well over the 7- to 30-MHz range.

the 1:1 balun transformer

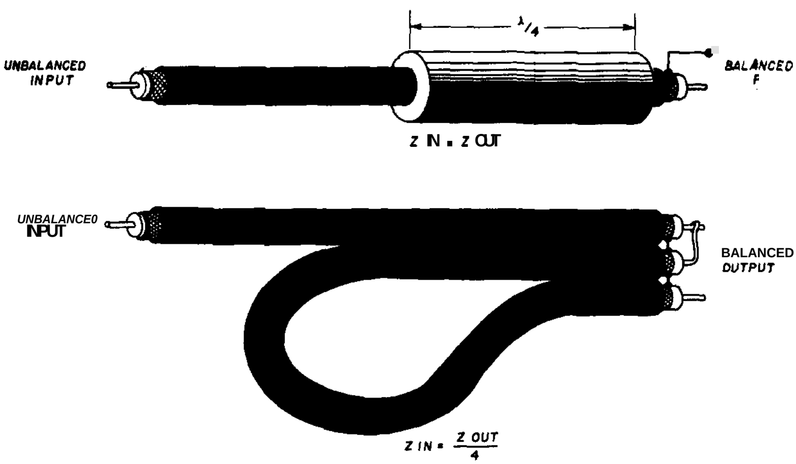
The 1:1 balun is well suited to the problem of feeding a split-dipole radiator with an unbalanced coaxial line. If this important balancing function is left out, it may cause transmission-line radiation and can result in confusing SWR readings when attempts are made to measure the characteristics of the antenna system. For the greatest benefit, a balun should cover the range of 3.5- to 30-MHz so that it can be used with three-band beams as well as "trapped" 80- and 40-meter doublets.' Useful, too, are baluns which cover the "tribander" antenna range of 14 to 30 MHz.

Sophisticated ferrite toroid baluns have been built which span the range of 50 kHz to 500 MHz,[3] and simple linear-coil versions of these exotic devices may easily cover the 80- through 10-meter bands. A well-built 1:1 balun is a noncritical device that provides near-unity transformation; it may be used with either 50- or 70-ohm coaxial transmission lines and most common antennas.

winding) inductor, and may be compared to a two-wire transmission line wound in a coil with an extra balancing winding provided to complete the magnetizing-current path. In effect, the balun is comparable to a transmission line transformer having a 1:1 ratio over the higher-frequency region of operation and to a 1:1 coupled transformer at the lower frequencies.

High-frequency response of the coil balun is largely limited by the distributed (shunt) capacitance and coupling between the windings (both of which are quite critical). The low-frequency response is determined by the primary inductance of the coupled transformer.[4] The most difficult part of the balun transformer to adjust is the distributed capacitance of the windings. Much effort must be expended to reduce this to a minimum value

fig. 1. Narrow-band balancing transformers may be made of lengths of corxial cable. The upper design provides a 1 to 1 balanced transformation and the lower design, a 4 to 1 transformation. Baluns provide good balance over approximately 0.05% of the operating frequency.



There's more to the balun than meets the eye however, and practical construction information is sorely lacking for these interesting devices. Some simple designs are shown in this article which work well over the h-f range and can be built easily and inexpensively in the home workshop.

The schematic of a 1:1 coil balun is shown in fig. 2. The balun consists of a trifilar (three-

without destroying the interturn magnetic coupling."

A coil balun may be constructed with an air core, and designs of this type work well. However, the number of turns required for good low-frequency response are large, and sufficient distributed capacitance exists in a simple balun of this type to limit the range of proper operation to a frequency span of

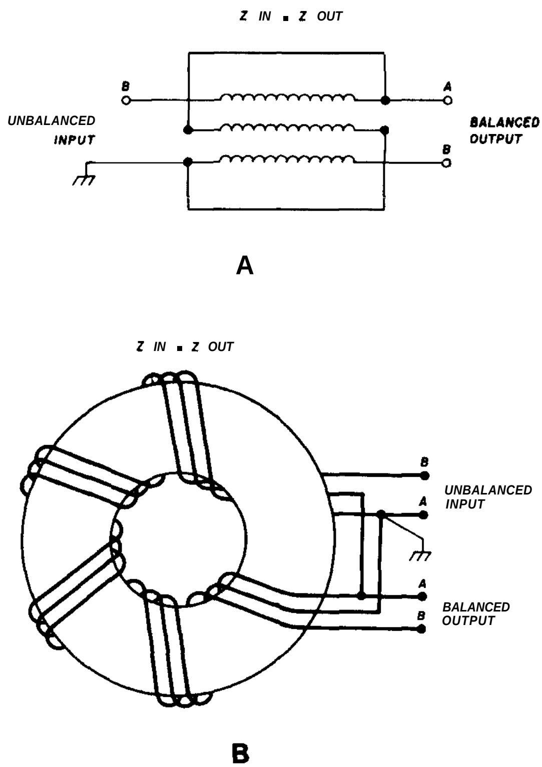
3 or 4 to 1. If a high-permeability ferrite core is used in the balun, fewer turns are required on the trifilar winding for a given low-frequency response. This reduces the overall distributed capacitance of the winding and improves high-frequency response. A frequency range of 10 or 20 to 1 is common with a balun of this type. The ferrite core, however, is power limited, especially at the high-frequency end of the operating range.*

the ferrite-core toroid balun

Toroid baluns have been described several times, and variations of the basic design shown in **fig. 2** are available in kit form. Several experimental toroidal baluns were built using available information. An expanded Smith-Chart plot** of one of the best designs is shown in fig. 3. The chart illustrates the reactive and resistive components of the balun when terminated in a 50-ohm load and examined with an rf impedance bridge. To plot the chart, a frequency run was made with a signal generator driving the balun through an rf impedance bridge; the bridge readings are transferred to the Smith-Chart. A typical test set-up is shown in **fig. 4.**

The curve for the toroid balun shows that it is not a perfect 1:1 transformer. It presents a termination of 49 ohms at 3.5 MHz and gradually rises to 55 ohms at 14 MHz. Thereafter, the impedance drops to about 44.5 ohms at 29.7 MHz. This balun presents considerable reactance at the lower frequencies,

fig. 2. A three-winding transformer provider a balanced I-to-1 transformation (A). Note that the center winding acts as the balancing coil, completing the magnetizing currant path. Connection of balancing winding to transformer winding at the input end must be taken as ground (point A) to preserve optimum balanca. The three-winding balun may be used with a toroid core (B).



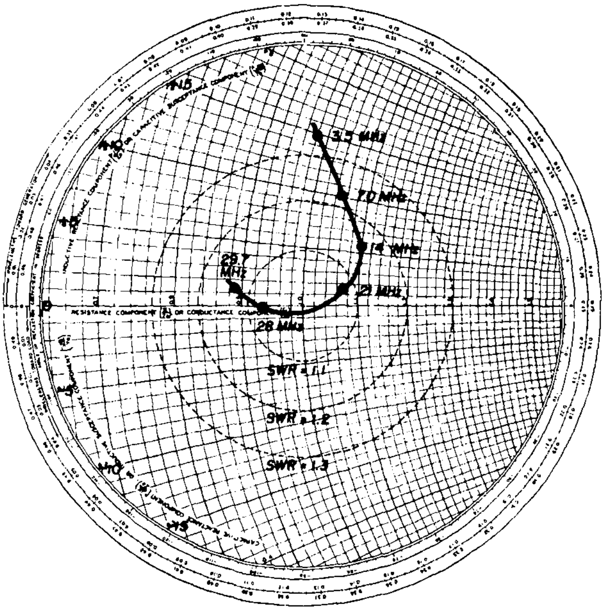
resulting in an SWR greater than 1.1 at frequencies below about 19 MHz. Above 19 MHz, balun reactance is slight. This chart shows that, when terminated in a "perfect" 50-ohm load, the balun will introduce a slight SWR (greater than 1.1 but less than 1.35) on a transmission line at frequencies below 19 MHz. Not indicated on the chart is the fact that the balance of the device begins to deteriorate above 30 MHz or so.

In summary, then, a ferrite toroid balun of this design is a good performer above 19 MHz and a fair performer down to above 3.5 MHz. The quality, or excellence of the balun, of course, is a subjective thing and an arbitrary SWR figure of 1.2 was chosen as a practical limit defining balun excellence. A really **good** balun, however, can plot a curve falling with-

* According to Sosin,[4] the bandwidth is limited at the low-frequency end by the low value of shunt inductance, and at the high-frequency end by a low-pass pi network formed by spurious shunt capacitance and leakage reactance. In order to keep the leakage reactance small, the winding must have a minimum number of turns. Consequently, the ferrite core is very heavily loaded. The power rating of the core depends on the cooling effectiveness, and the temperature rise of the core might become quite high. As the working temperature is increased, a runaway temperature is reached where operation is impractical because of unbalance; ultimately, the balun will be destroyed.

** The Smith Chart used for this series of tests is an expanded type normalized at 50 ohms. The resistive component of the measured impedance falls along the X-axis and the reactive component falls along the Y-axis in the normal manner. (General Radio #5301-7561-NE).

fig. 3. Expanded Smith Chart shows response of ferrite-toroid balun plotted with respect to 1.2:1 and a 1.1:1 SWR circles super-imposed on the chart. Balun consisted of 15 turns number 14 wire bifilar wound on 2.4" diameter, Q-1 toroid. Balun was optimized at h-f end of its range by squeezing and adjusting the spacing between adjacent turns and lacing the windings in place.



in the 1.2 SWR circle on the expanded Smith Chart.

Not shown on the chart is the difficulty I encountered in constructing a toroid ferrite balun capable of this degree of performance. Neatness of construction is a virtue, and the coupling, or spacing, between each turn and between the trifilar groups of turns is very critical. In the photograph you can see that each trifilar group of wires is tied in place with a length of lacing twine. Unhappily, if the trifilar groupings are too loose on the core, the high-frequency performance of the balun suffers. If, on the other hand, the lacings are too tight and the turns too tightly compressed together, the high-frequency performance suffers equally as before.

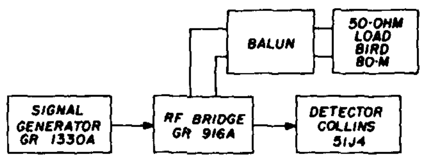
I spent a considerable amount of time adjusting the balun windings while it was connected in the measuring circuit. Only by juggling the windings while watching the rf bridge readings was I able to achieve a satisfactory transformation ratio and acceptable bandwidth. The experience was exasperating in the extreme, and I reached the conclusion that it was virtually impossible to construct a toroid ferrite balun of this type with acceptable characteristics without the assistance of an rf bridge.

The coupling between trifilar windings was very critical and had to be hand-adjusted to a fine degree to make the balun perform above 20 MHz or so. Regretfully, the toroidal construction was cast aside in favor of other designs that could be easily built in the home workshop.

the ferrite-slug balun

I next turned my attention to the ferrite slug balun, a number of which are available commercially in inexpensive versions. A simple trifilar balun was wound on a one-half-inch diameter ferrite core;' measurements were taken and plotted on an expanded Smith Chart as shown in **fig. 5.** The slug balun performed well over the 7- to 50-MHz range (within the 1.2 SWR circle on the chart), and

fig. 4. Balun tests were conducted with an r-f bridge and balanced load. The r-f bridge was compensated for h-f response to 50 MHz. Resistance end raactance measurements were made directly from the bridge and compensated for system errors.



seemed acceptable at 3.5 MHz where the plot fell just outside the 1.3 circle.

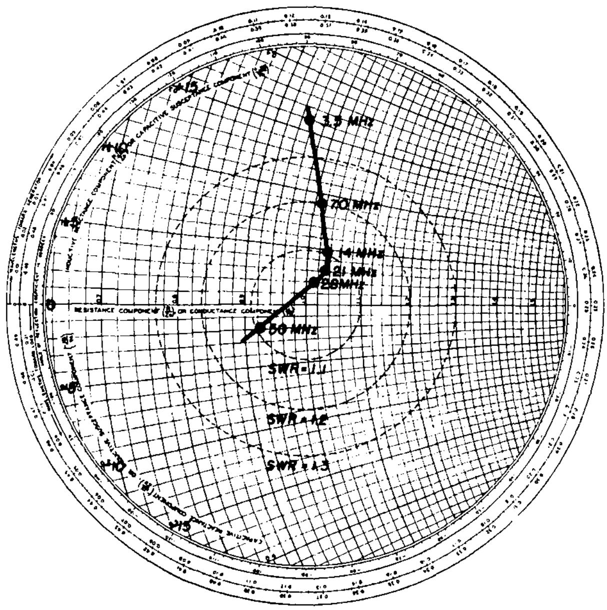
Wire spacing on the slug core did not seem to be nearly as critical as on the toroid form, and a neat, closewound coil performed very well. Anchoring the winding proved to be a problem; coating the wires with coil dope, Krylon or nail polish increased the distributed capacitance of the balun and disturbed the

• The Newark Electronics Corporation Industrial Catalog number 68 carries a partial listing of Indiana General Corporation Ferrites. A suitable 112-inch diameter rod is the Indiana General CF-503 which is 7-1/2" long. (Newark part number 59F-1521). The ferrite may be nicked with a file around the circumference at the desired length and broken with a sharp blow.

transformation above 20 MHz. The solution was to coat the end turns and leads of the assembly and to leave the winding free of adhesive material.

After I had built a few ferrite slug baluns and measured them on the rf bridge, they were tested with a transmitter and a dummy load. The balun with an Indiana General CF-503 slug core readily accepted 700 watts of power up to 14 MHz—the core became only slightly warm after three minutes of operation. Above 20 meters, however, core losses went up, and it was necessary to derate the

fig. 5. Chart shows plot of ferrite-slug balun with respect to 1.2 and 1.1 SWR circles. This simple, compact balun performed well over the 3.5- to 50-MHz range and the plot fell within the 1.2 SWR circle over the ranga from 7 to 50 MHz. The balun consisted of 6 turns of number 14 wire on a Q-1 slug and was patterned after a design by W8FYR. This balun provided the widest frequency response of all the baluns tested but war power limited at the h-f end of the range.



balun to 500 watts at 21 MHz and 400 watts at 28 MHz. I didn't run any power tests at 50 MHz, but I estimate that the balun is good for 100 watts or so at this frequency. With intermittent voice ssb operation, the power capability could probably be doubled with safety.

the air-core balun

Because of the power limitation and core cost of the ferrite baluns, I decided to explore the capabilities of a simple air-core balun. A suitable design evolved after several failures, and the characteristics of a prototype balun were plotted on an expanded Smith Chart (fig. 6). The balun plot fell within the 1.2 SWR circle over a range of 9 to 29.7 MHz and was just inside the 1.3 SWR circle over the 40-meter band. The power capability of the balun was better than 1000 watts at all frequencies within the range of operation. Ease of construction, low cost and simple adjustment plus the improved power capability emphasize the fact that this type of balun is well adapted to home construction

practical balun construction

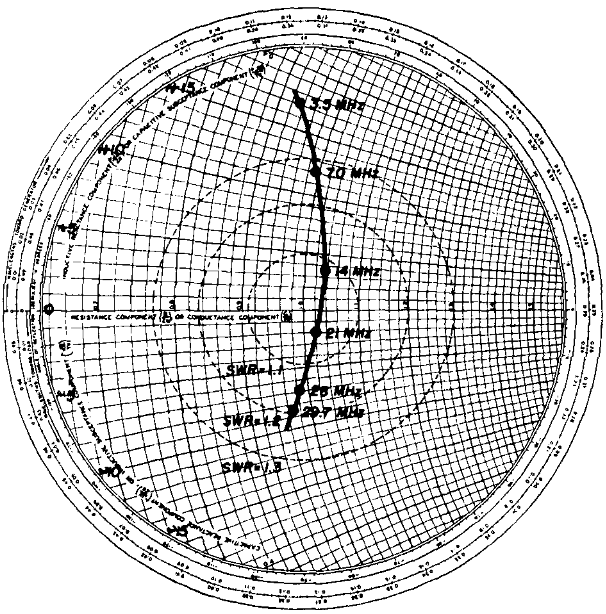
A practical air-core balun for the h-f range is shown in the photograph. It consists of three coils of number 14 Formvar* coated wire, ten turns to each coil. The windings are placed on a 4-inch long piece of l-1116-inch outside diameter polyvinyl-chloride (PVC) plastic tubing. This gray, plastic material is commonly used in most areas of the United

* Forrnvar (polyvinyl formal-phenolic resin) coating is superior to plain enamel because of its greater dielectric breakdown strength. Nyclad (polyvinyl formal with nylon overcoat) wire has even greater dielectric breakdown strength. It is manufactured by Belden Manufacturing Company. Nyform is another trade name for this coating (Anaconda Wire & Cable Cornpany).

States for water pipe, and a small chunk of it can be obtained from your local plumber at little or no cost. The trifilar winding is simple to construct after you get the hang of it, and expertise is readily gained after an experimental winding is made.

Three pieces of wire about 4-feet long are

fig. 6. Chart shows plot of inexpensive air-core balun shown in the photograph. While the frequency response is not as great as that of the ferrite-slug balun, the less expensive air-core balun works well over the range of 3.5 to 29.7 MHz. Response falls within the 1.2 SWR circle over the range of 9 to 29.7 MHz. Balun consisted of ten turns of number 14 Formvar insulated wire on a piece of 1-1/16" piece of plastic pipe, connected as shown in fig. 2A.



cut and smoothed to eliminate bumps and kinks. The wires are placed parallel to one another and the far ends held in a vise. The near ends are scraped clean of insulation and wrapped around 4-40 bolts placed in the PVC form as anchor points. The three wires are then wound side by side on the form as one, until ten trifilar turns are on the form. If you wind carefully and keep reasonable tension, the coil will adhere closely to the form. The other ends of the windings are scraped clean and attached to the proper bolts as shown in the illustration.

The last step is to interconnect the center, or balancing winding. As you can see in the photograph, the center coil cross-connects the outer coils. The terminal connections are reversed in physical position from one end of the coil to the other, and the proper balancing connection may be made by connecting the ends of the center winding to the outer coil winding at each end of the bifilar assembly. Neatness is important, and an evenly wound coil with the turns **just** touching each other will provide greater operational bandwidth than a haywire winding with uneven spots or lumps in the wires. Once the cross connections have been made, the assembly may be held in place by a few drops of epoxy cement, Krylon, or coil dope placed at the ***ends*** of the windings.

Note that the input terminals of the balun are non-symmetrical from an electrical point of view. That is, **point A** at the input end is taken as ground. The free winding, **point B**, is hot. At the output end, of course, the terminals are balanced with respect to ground. Transposition of the input connections will degrade balancing action. Either end of the unit may be used as the input, of course, provided **point A** is taken as ground.

encapsulating the balun

Air-core baluns of the types discussed here are sensitive to nearby capacitance, and some thought must be given to the problem of protecting the windings from sun, rain and weather without upsetting their electrical characteristics. I tried to encapsulate the baluns in epoxy resin, but the resulting increase in distributed capacitance degraded balun performance to a serious degree at the higher frequencies. A simple solution I finally arrived at was to place the balun in a cylindrical case made from a 2-1/2-inch diameter polyethylene "squeeze bottle" that once held hair shampoo. I cut the ends from the bottle and put in plywood discs with small wood screws. The balun was suspended inside the bottle by the wire leads which were connected to 10-32 bolts placed in the plywood discs. When the unit was completed, it was coated with epoxy resin to waterproof the joints.

A subsequent visit to the local plumber disclosed that there was a "welding" torch available which "welded" PVC pipe with a blast of hot air, using a strip of PVC material as plastic solder. Accordingly, a length of 2-114-inch outside diameter PVC pipe was cut to form an outer jacket for the balun. Two end discs were cut from a sheet of the same matkrial. Terminals were placed in the discs, and the balun was wired in place between the terminals. The assembly was then slipped within the larger PVC pipe and "welded" by the plumber with his unique hot-air torch. The result was a neat, compact, waterproof balun assembly ready to mount at the feed point of a beam antenna or hung from the center point of a dipole!

using the balun with your antenna

The air-core balun makes a fine balancing device for use between the coaxial line and the split dipole element of a three-band beam antenna. While nominally designed for a nonreactive 50-ohm load, the balun will happily accept the degree of mismatch presented by the typical amateur antenna. Most three-band beams using trapped elements present a different picture to the transmission line on each of the amateur bands, and rarely is this a "pure" 50-ohm termination.

Typically, such an antenna represents a frequency-sensitive, reactive load whose resistive component falls to about 25 ohms on 14 MHz, 35 ohms on 21 MHz and 45 ohms on 28 MHz. Because of the complex interaction of traps and elements, the chance of achieving a nonreactive 50-ohm load at more than one frequency in one of the three amateur bands is rather remote.

No matter; the operation of the antenna is not dependent to any great degree upon feedline match provided the SWR at the transmitter end of the line is not too great to prevent proper transmitter loading. It is permissible, therefore, to use the balun directly with a multi-band beam without additional matching devices other than those normally associated with proper antenna ooeration. Therefore, the balun may be used with varying beam terminations, possibly over the range of 15 to 100 ohms or so, while still providing worthwhile balancing and feedline isolation.

low-frequency air-core baluns

During the investigation of air-core baluns, two low-frequency units were built that may be of interest to experimenters. The first covered the range of 2.5 to 15 MHz and consisted of seven trifilar turns of number 14 Formex wire (21 total turns) wound to a length of I-314-inches on a piece of 2-3/8-inch diameter PVC pipe. The design center of this balun was 8 MHz, and the impedance plot fell within a 1.2 SWR circle on the expanded Smith Chart over the range of 3.4 to 15 MHz.

A second balun covered the range of 0.54 to 2.5 MHz. This balun was composed of eighteen trifilar turns of number 14 Formex wire (54 total turns) wound to a length of 3-5/8 inches on a length of 3-112-inch diameter PVC pipe. Design center was 1.2 MHz, and the impedance plot fell within a 1.2 SWR circle on the Smith Chart over the range of 0.7 to 2.1 MHz. This balun is well suited for 160-meter work, while the previously mentioned unit is designed for 80- and 40-meter operation. As it turned out, the smaller unit worked well on the 20-meter band as well and was subsequently used on a two-band 40- and 20-meter beam.

conclusion

The simple air-core balun is the easiest to make and adjust and the least expensive of the three balun types discussed in this article. It is capable of working at the full amateur power limit over its design range and greatly simplifies the operation and adjustment of any antenna system. Try one and see; you'll like it!

I'd like to thank Willie Sayer, WA6BAN, for his help in the preparation of this article and for his contribution of the hundreds of feet of copper wire that went into various unsuccessful designs along the way!

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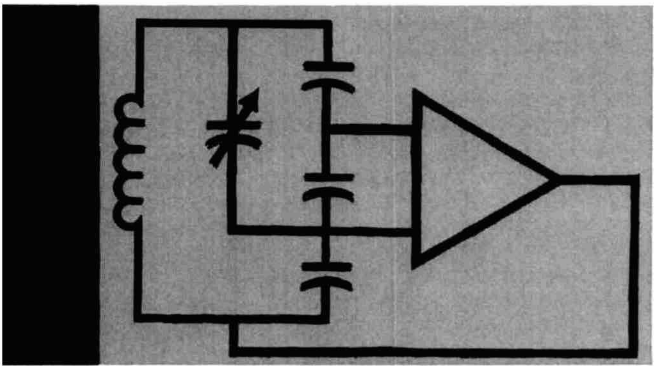
ham radio



stable transistor vfo's

A discussion of the Vackar and Seiler oscillator circuits

Jim Fisk, W1DTY, RFD 1, Box 138, Rindge, New Hampshire 03461

There has been a lot of interest in the Vackar oscillator lately because of a recent article[1] describing its many merits. Although there hasn't been too much information on this circuit in the American magazines, a wealth of information has been published in the **RSGB Bulletin.** In addition, there have been a number of amateur articles which have used a somewhat similar circuit—the Seiler oscillator.

Actually, both the Vackar and Seiler circuits are closely related to the Colpitts oscillator. The Vackar, named after its inventor, Jiri Vackar, a Czechoslovakian, was originally described in 1949.[2] The Seiler circuit, although almost forgotten, was described in **QST[3] in 1941.** Both of these circuits were designed to minimize loading on the tuned circuit, thereby increasing stability.

Most VFO's in use today use the series-tuned Colpitts or Clapp circuit; interestingly enough, Clapp based his design on the work of Vackar.[5] You can see from fig. **1** that the Colpitts, Clapp, Vackar and Seiler circuits are very closely related. The Colpitts circuit, of course, is the father of them all. Seiler added a third capacitor in the divider to lessen the load on the tuned circuit. Vackar did much the same thing, but put a variable capacitor across a portion of the tank circuit to increase the tuning range. Clapp went on to simplify the basic Vackar circuit.

Since the Colpitts and Clapp circuits have been covered quite well in the amateur literature,[17-24] the discussion here will be limited to the Seiler and Vackar circuits.

the Seiler oscillator

Until Seiler's article in 1941, most VFO's used the Hartley or high-C Colpitts circuit. The Seiler design permitted the amateur to voltage regulation, and by 1941 standards, the stability was very good.

One of the big advantages of the Seiler circuit is the large capacitors which are placed across the active device — in this case a transistor. These large capacitors tend to swamp out any reactive changes in the transistor and limit the harmonic output, thereby increasing frequency stability. Since capacitors C2 and C3 are usually much larger than C1 or the variable capacitor (Cv) in the Seiler oscillator, the frequency of oscillation may be simplified to:

$$f_{osc} = \frac{1}{2\pi \sqrt{L(C1 + Cv)}}$$

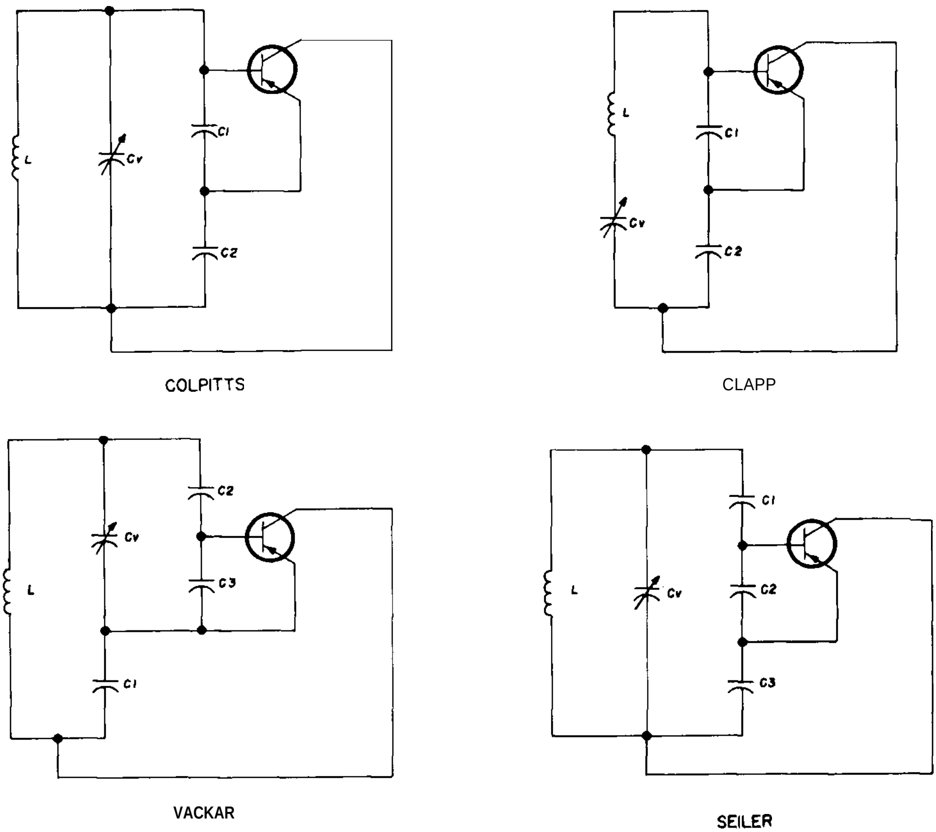


fig. 1. Circuit configurations of the Colpitts, Clapp, Seiler and Vackar oscillators. The Clapp. Sailer and Vackar circuits are derivations of the basic Colpitts circuit.

use a relatively low-C circuit that provided high stability and a tuning range of 1.8:1. A 6F6 was used in the original article, without

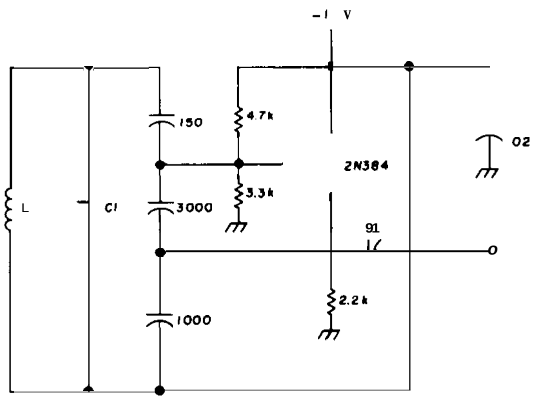
Several vacuum-tube versions of the Seiler oscillator have appeared in the amateur-radio magazines, but in most cases the designers weren't aware that their circuit was an extension of W8PK's original design. In at least one case, the author called his circuit a "ground-plate Colpitts type."[6]

To my knowledge, the first transistorized version of the Seiler oscillator was W3JHR's "synthetic rock" which was published in CQ in 1963.[7] 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[8] 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!

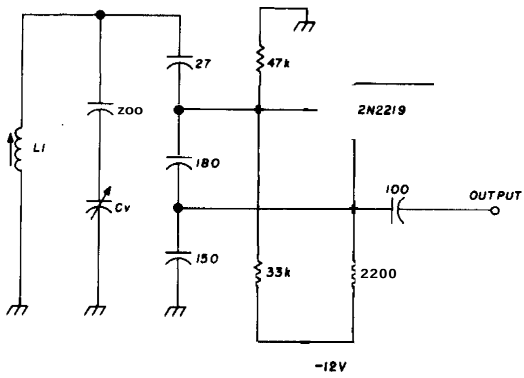
Don't let that 200-pF capacitor in series with the variable capacitor confuse you. It was used to set up the bandspread range of the variable capacitor.

fig. 2. W3JHR's "synthetic rock" —a Seiler oscillator—tunes from 4.9 to 6.1 MHz with tank componentr from an old ARC-5 transmittar.



Another transistorized Seiler oscillator was described by G3BIK,[9] 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 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!

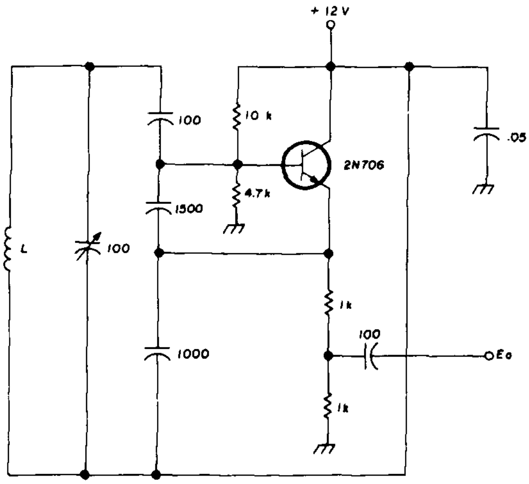
fig. 3. This Seiler oscillator, designed by K9ALD, tunes from 4.95 to 5.6 MHz. Total drift is reported to be 25 Hz. L1 is 2-1/2 turns number 16 on a 1-1/4" ceramic form. Variable capacitor C_v is a 100-pF variable in parallel with an 82-pF silver mica.



Since all the amateurs who have built transistorized Seiler VFO's 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 war 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 like it's crystal controlled. It far surpasses any VFO circuit I've

fig. 4. G3BJK'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.



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.[10] 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 C1 (fig. 1); 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.*

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.

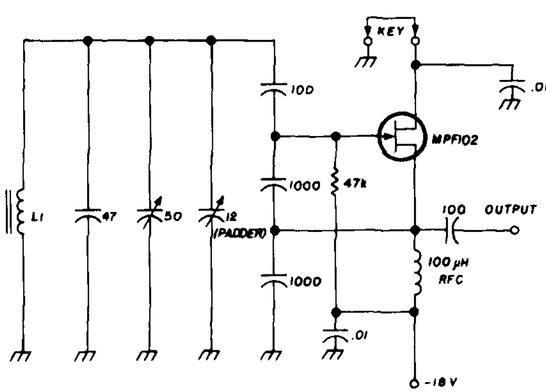
Choose a value of *Q* as high as possible, because oscillator stability is very closely related to tank-circuit Q. For all practical purposes, the Q of the tank will be determined by the inductance you select, so use the best coil you can. If you have lots of room, air-wound coils are very good; if you're 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, C2 and C3 (fig. 1), are not critical. However, they should be

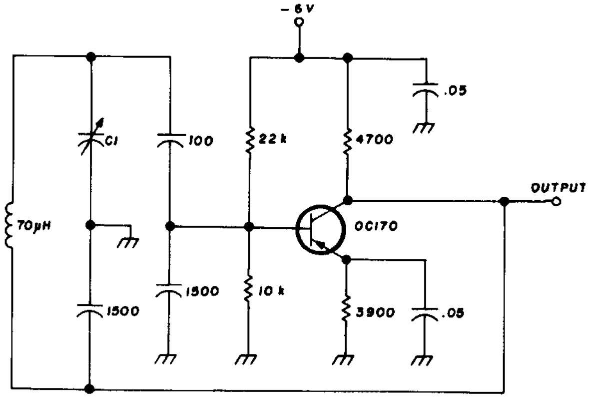
fig. 5. Stable Seiler oscillator using an FET. The tuning range of this circuit is 3.40 to 4.01 MHz. L1 consists of 44 turns number 30 on a 1/2" ferrite core (Amidon T-50-2*).

* 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.



quite a bit larger than C1 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

fig. 6. **Transistorized** Vackar oscillator designed by L. Williams, a British SWL. **C1** is a **30-pF** trimmer in parallel with a **75-pF** air variable.



still result 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 C3 to C2 less than the current gain of the transistor.

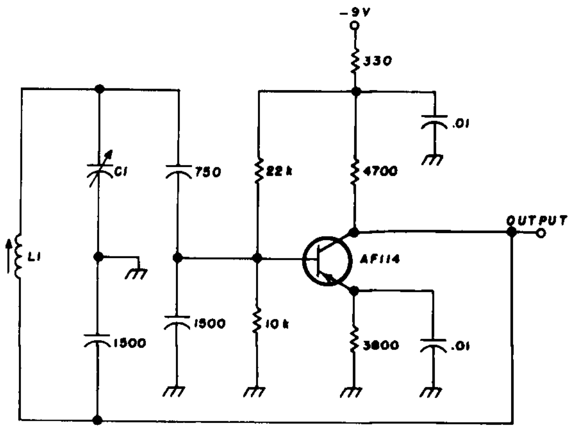
The variable capacitor, C usually consists of a variable in parallel with a padder. The padder can be adjusted so that the variable will cover the desired frequency range. Capacitor C1 determines the amount of drive to the transistor and is relatively small. The best approach here is to start off with about 100 pF at C1 and reduce it until the oscillator ceases to function. Add about 50% to this value as a safety factor for the final value of C1.

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.

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 impeclance and is very loosely coupled to the tuned circut. This 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

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.



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,[4] it has remained virtually unused. W9IK described a vacuum-tube Vackar oscillator built by W9TO[11], and a design appeared in **Radio and TV News,**[12] 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.[13] 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[15] **(fig.** 7). He experienced some difficulties with temperature drift, but felt they could be cured by putting the circuit in a die-cast 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 C1 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.** 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 I-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, 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

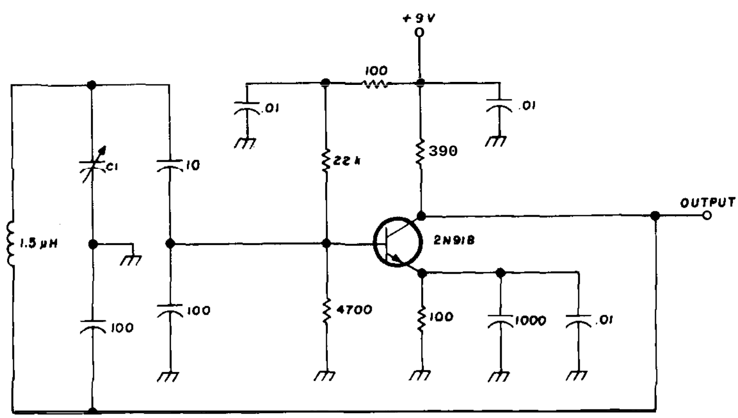


fig. 8. Thir Vackar oscillator designed by G. B. Jordan tunes from 16.9 to 34.7 MHz. C1 is a 12-pF air variable in parallel with a 6.2-pF silver mica.

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 C2, 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 1-1/2 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 VFO's 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 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 VFO's 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 shown here have indicated exceptional performance and stability with them. If you have done any experimenting along these lines, I would certainly like to hear about it—both of these circuits have been buried in the literature long enough. They seem ideal for transistor work, easy to design and a good choice the next time you're thinking about a new VFO.

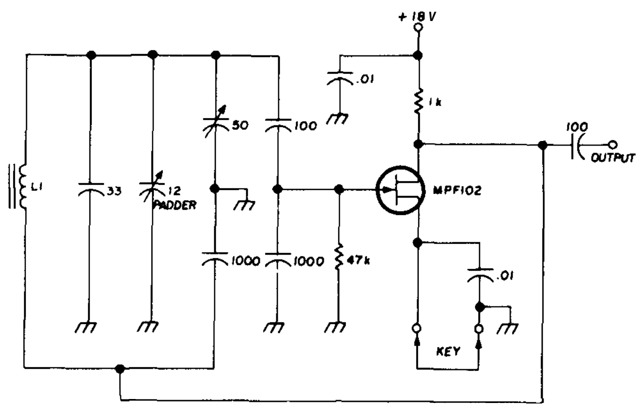


fig. 9. FET version of the Vackar oscillator is extremely stable. L1 is 48 turns number 30 on a 1/21' ferrite core (Amidon T-50-2)

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ham radio

some interesting aspects of increased solar activity

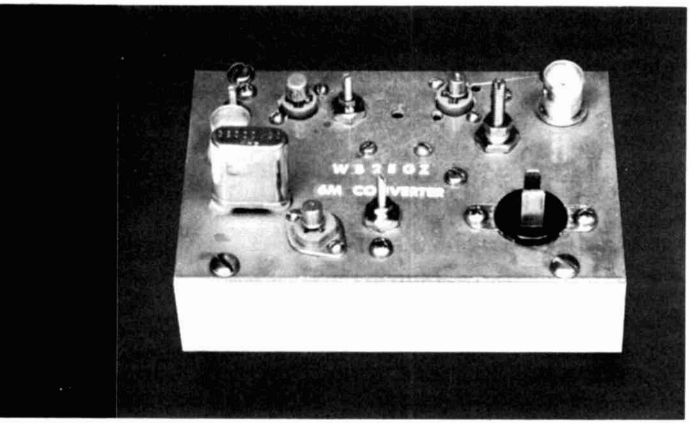
The present upswing in the sunspot number is expected to reach its peak in the very near future, probably at the end of 1968. It is commonly known that an increase in the number of sunspots will result in a marked over-all improvement of short-wave radio communications. However, during periods of increased solar activity, a prolonged deterioration of propagation conditions is very often experienced. This frequent disruption in radio communications is produced by gigantic flares that originate within the sun. Since the number of sunspots and the probability of solar flares are directly related, poor short-wave communications may be often expected during periods of high sunspot numbers.

A solar flare releases electromagnetic radiation, delayed by 8.3 minutes; cosmic-ray particles, delayed from 15 minutes to several hours; and magnetic-storm particles, delayed. between 20 and 40 hours. The electromagnetic radiation is in the form of visible light, radio waves, ultraviolet rays, and x-rays; some of the immediate effects are occasional F-layer increase, frequent E-layer increase, and D-layer increase. Cosmic-ray and magnetic-storm particles cause delayed effects, such as magnetic storms, ionospheric storms, auroras and a general increase in cosmic-ray radiation.

Besides the well-known "radio blackouts," solar storms produce an increase in earth currents which may cause drastic power-line voltage variations and over-heating of conductors. Power-line voltage fluctuations as much as 108 to 117 volts have been recorded. Corrosion and conductor vibration in areas of high ionization have also been correlated to magnetic storms. With the increase in system voltages and the increasing importance of system reliability, the nature of space plasmas and their associated phenomena on utility systems present a most important study for future utility system design.

As far as amateur radio is concerned, it must be borne in mind that a high sunspot period means "good DX conditions," but at the same time, you should anticipate severe and frequent short-term communications disruptions. The probability of an actual power-line failure is highly unlikely.

Joe Mikuckis, K3CHP



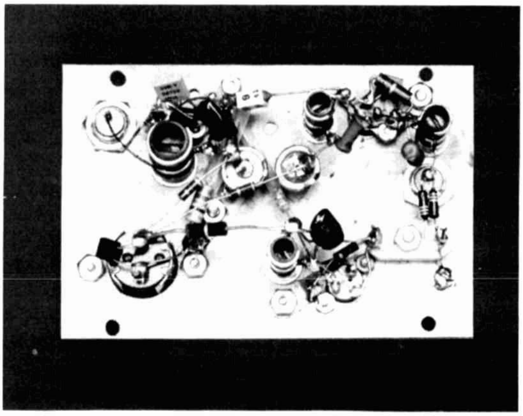
the WB2EGZ six-meter mosfet converter

Construction details for a low-noise, high-gain converter for six meters that out-performs designs using vacuum tubes or bipolar transistors

Don Nelson, WB2EGZ 9 Green Ridge Road, Ashland, New Jersey 08034

What's this we hear about MOSFET'S? Are they just a laboratory phenomenon? Do they burn-out with the flick of a finger? Are they priced for the idle-rich? Is their frequency response a limiting factor?

Never have I, as an experimenter, been so excited about a device as I am about the MOSFET (metal oxide semi-conductor, field-effect transistor). Immediate success with a simple six-meter converter has prompted my sharing a design which is almost certain to work for anyone with some talent for tinkering.



the MOSFET

Let me digress from the converter for a moment to talk about the MOSFET. Frequently, and possibly more correctly, the device is called an insulated-gate, field-effect transistor (IGFET). The reasoning behind this terminology is that there is no junction between the gate and the channel such as is the case with the junction field-effect transistor (JFET).

Most hams are more familiar with the JFET because of the vast number of magazine articles using the TIS34, MPF102, 2N3819, etc. The MOSFET's used here are similar to the JFET in that they operate in the depletion mode; that is, a negative voltage on the gate reduces the width of the channel, and increases the resistance between the source and drain.*

In the MOSFET, the semiconductor channel is isolated from the gate by a metal-oxide layer, and the change in channel width is controlled by an electrostatic voltage across that layer. The input resistance is considerably higher than a vacuum tube and also higher than the JFET. The feedback capacitance from output to input is considerably lower than that of a tube or JFET. As you will see later, this is a distinct advantage.

* There is a second kind of MOSFET where the channel width increases with positive gate voltage; this is an enhancement type.

The MOSFET's intrinsic input impedance may work to a disadvantage if the gate has no ground return; voltages which are high enough to destroy the transistor may develop without it. Static burnout, as the condition is called, may also occur by touching a charged body to the open gate. On the other hand, static burnout is almost impossible with any return path for the gate–even several hundred megohms.

A little extra care in handling the MOSFET will pay dividends; remember the following tips:

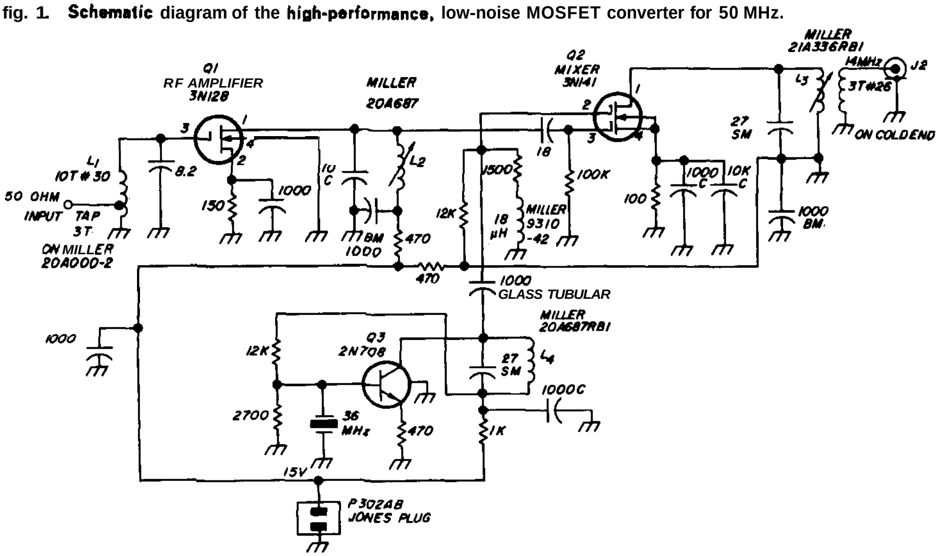
1. If the leads have to be cut, hold the lead and transistor case with one hand to reduce mechanical shock and possible static discharge.

2. Never insert or withdraw the transistor when power is applied to the circuit.

3. If the transistor is to be soldered in the circuit, the soldering iron must be grounded (three-wire system).

I have yet to burn out a MOSFET–touch wood!

fig. 1. **Schematic diagram of the high-performance, low-noise MOSFET converter for 50 MHz.**



converter design

Simplicity makes this converter an almost unbelievable circuit. Only three inexpensive transistors are used, and the number of tuned circuits has been minimized. Without a junk box you may pay as much as $25 for parts,

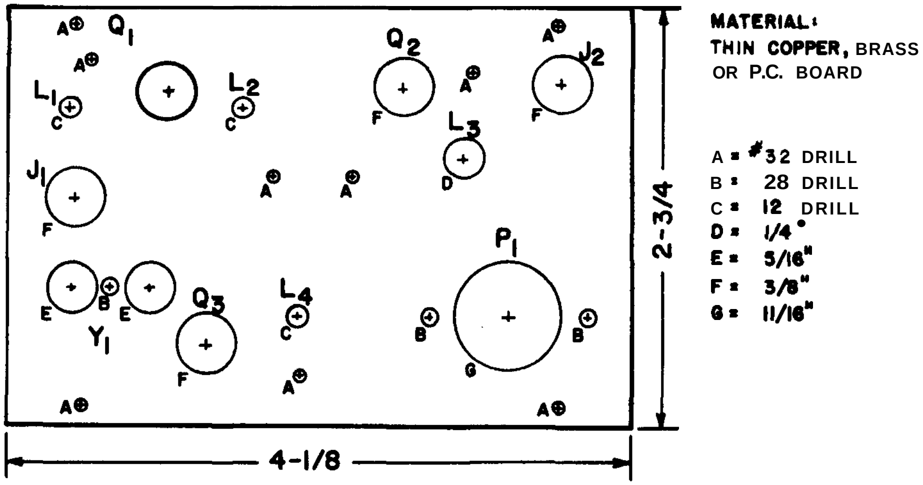


fig. 2. Chassis layout for the 50-MHz MOSFET converter. A full-size drill template is available from ham radio magazine for 25c and a srlf-addressed, stamped envelope.

but most hams will pay less.

The first stage, an rf amplifier, uses a 3N128 which is an N-channel triode **(fig. 1).** Both the input and output are tuned, but no neutralizing link is needed. The internal feedback capacitance is low enough so that the stage is stable as long as the source (antenna) and load (mixer stage) are properly matched. A trap may be needed to reject interference from Channel 2 if there is a station near you; the 58-MHz trap used in the ARRL **Handbook** Nuvistor converter would be suitable.

A typical noise figure for the 3N128 is 3.5 dB–this is in line with most good transistors, and superior to the Nuvistor. Evidently, the 3N128's I tried are even lower than the advertised figure. Because of the high gain of the 3N128 at 50 MHz, we can get away with only one rf stage. According to reports, the MOSFET is about ten times better than the bipolar transistor with respect to cross modulation and has 25 times the dynamic range. These characteristics compare favorably with tubes.

The mixer is an N-channel tetrode MOSFET, type 3N141. This is the most expensive transistor in the converter and costs $1.55. While tetrodes were originally developed to improve the cross-modulation characteristic of the MOSFET in front-end circuits, the mixer application is equally rewarding.

The construction of the device is similar to two MOSFET triodes in series. That is to say, the drain of one channel is internally connected to the source of the second channel. The gate of the second channel must be biased positive for proper operation by the resistor divider connected to pin 2. That gate also receives the output from the local oscillator, so a choke is necessary in the ground link of the dc divider to prevent rf attenuation.

Gate 1 (pin 3) accepts the output of the rf amplifier which is mixed with the local oscillator to provide the 14-MHz output. This mixer circuit is superior to any transistor circuit, either JFET or bipolar, that I have tried. The noise figure, conversion gain, and cross-modulation characteristics are excellent, challenging the best features of tubes and transistors.

There is nothing sacred about the oscillator circuit. The transistor shown in the photograph is actually a 2N2708, but the RCA40237 is less expensive and an equal performer.

construction

A simplified chassis hole-center layout **(fig. 2)** may be used to reduce construction time. Sockets were used and are recommended. In as much as the MOSFET is a high-impedance device, tanks may be dipped with the transistors in their sockets. Only slight touch-up will be required when the receiver is turned on.

Peaking the oscillator coil may cause hard starting. Should this happen, apply power, detune L_4 until the oscillator starts, and secure the slug at that setting.

testing performance

As you may have guessed by now, I have been impressed by the tube impersonator called a MOSFET. Test equipment was not available for sophisticated measurements of gain and noise figure on the converter, but a few observations were made:

1. The converter is noticeably quieter than the ARRL **Handbook** converter which uses a neutralized 6CW4 Nuvistor.

2. With the noise level set to S1, signals are about 6-dB greater (one S-unit) using the WB2EGZ converter.

3. No overloading or cross modulation has been detected at signal levels where bipolar transistors have failed miserably.

From the experience gained with MOSFET's on six-meters, I'm ready to try some experiments on my favorite band–two meters. If successful, I shall share the results with you.

ham radio

transistor-tube talk

If you're used to tube-terminology, you're probably often confused by transistor terms. Here is a little play on words that can be used to keep them straight. For those of you who are studying for the advanced or extra-class exam, this is a big help if you don't feel at home with transistors.

Let's review some tube terms first. A triode has three parts or elements: cathode, grid and plate. A transistor also has three parts—the emitter, base and collector. If you'll remember the function of each element, the comparison is simple.

The cathode of a tube **emits** electrons and the **emitter** of a transistor is the corresponding element. The key word is "emits." Carrying this a bit further, electrons are **collected** by the plate of a vacuum tube, so the transistor **collector** corresponds to the plate. The key word here is "collects."

This leaves the grid. As you all know, bias controls the action of the tube. The key word is "bias." Sounds like **base** doesn't it? So, that's it—the base is equivalent to the grid.

From these simple word similarities, it should be easy to remember that:

1. A grounded-cathode tube cricuit is the same as a grounded-emitter transistor circuit.
2. A grounded-plate or cathode-follower circuit is the same as a grounded-collector or emitter-follower transistor circuit.
3. A grounded-grid tube circuit is the same as a grounded-base transistor circuit.

just remember that the function of a vacuum-tube element is the key to its transistor equivalent. Score one for the advanced test.

George Haymans, W4NED

next month in ham radio magazine:

Bandswitching FET Converter

Solid-state Conversions

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Oscillators are not peculiar to single-sideband. They exist in all ham-radio equipment. Nor does single-sideband use oscillators that are any different, except that they should be extremely stable; the sideband relationships (to one another) have to be maintained consistently throughout a transmitter or receiver. Likewise, when the carrier is re-inserted by the bfo, that oscillator must be steady as a rock and right on frequency, or the recovered voice modulation won't sound much like the original.

There is another reason why you should understand oscillators in single-sideband. There are simply more of them. One fairly elaborate ssb receiver has five oscillators. In a typical exciter there may be that many, too, depending on how many frequency translations there are. At the very least, there will be two or three in a transmitter, and the same in a receiver.

The block diagram in **fig. 1** illustrates a transmitter that uses four different oscillators. Another, of similar design, also has an audio oscillator for A2 transmission and for testing. In **fig. 2** you see a receiver design that uses five different oscillators. In a transceiver, you may find some of the oscillators shown in **figs. 1** and **2** are combined, so that the overall transceiver may have only six or seven oscillators—maybe even fewer.

Somehow, oscillators have gained a reputation for being hard to understand. They are not, provided you are aware of certain principles. When you're trying to make one work that won't, you can simplify your troubleshooting by understanding what makes an oscillator tick.

Fig. 3 shows the four things it takes to make an oscillator. They are: amplification, dc power, feedback, and tuning. The differences among all the many oscillators that exist are in how each of these four jobs is accomplished. You can learn to classify the oscillator type by noticing how each function takes place. For example, a Colpitts oscillator, even though crystal controlled, derives feedback from a capacitive divider network—two capacitors in series, with a feedback tap-off between them. A Pierce crystal oscillator, on the other hand, has the crystal connected between plate (or screen grid) and control grid, providing feedback and tuning simultaneously. The Hartley uses a tapped coil for feedback.

Amplification is handled by a tube or a transistor. The **dc power** is merely to keep the transistor (or tube) working. The method by which these operating voltages are applied is usually the chief consideration whenever you think about or describe an oscillator.

The most important factor, so far as oscillation is concerned, is the **feedback.** Without

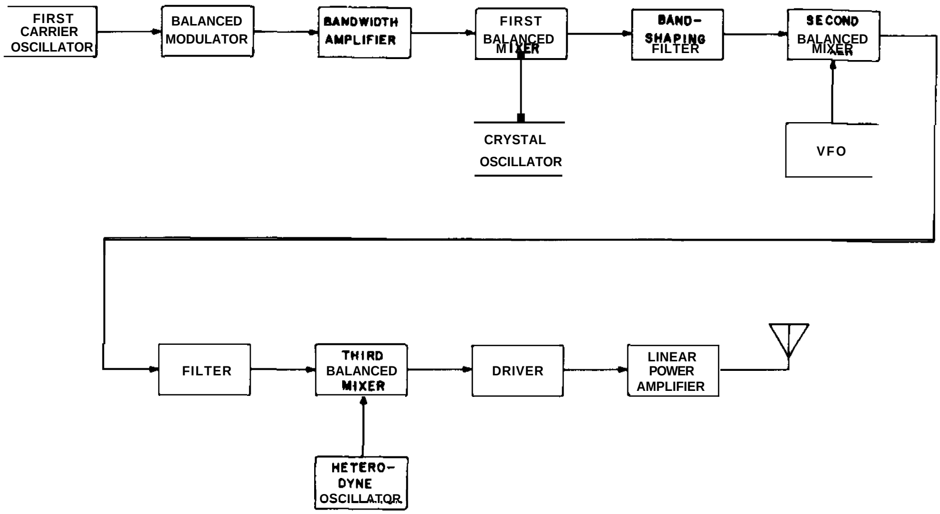


fig. 1. Block diagram of a ssb transmitter with four separate oscillator stages.

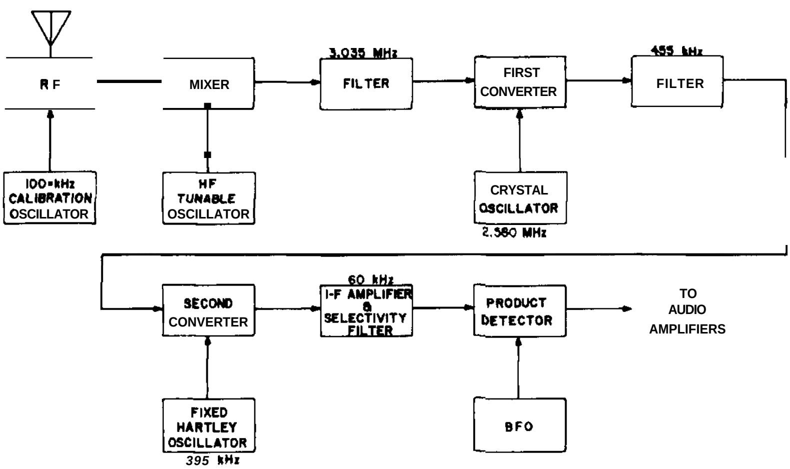
it, the tube and its dc operating voltages would form nothing more than another amplifier. The feedback takes some of the output **signal** voltage of the amplifier stage and feeds it back to the input in such phase that it is re-amplified. The signal is thus self-sustaining. You wouldn't want this action in a normal amplifier tube, but it is the essence of oscillator action.

Finally, the matter of **tuning.** It is of little value to have a circuit oscillate unless it is at some frequency you can use. The form of tuning in the oscillator often determines what kind of oscillator it is—what name it goes by. Tuning also affects how efficient or stable the oscillator is.

Keep in mind, then, that you can learn to recognize any oscillator by its characteristics in each of these four factors: the type of amplification, the method of applying dc operating voltages, the way feedback is developed and applied, and how the oscillator is tuned.

Rather than go into all the different possible combinations of these four requisites, it's more practical to examine typical circuits that use them. We'll begin with the most popular oscillator in all of single-sideband equipment—the Colpitts.

fig. 2. This communications receiver uses five different oscillators.



Colpitts—crystal and variable

The uses for this versatile oscillator are many. In different brands, you'll find it in one form or another as a vfo, as the carrier generator in a transmitter, and as a linear master oscillator (LMO) for transceivers. You'll find it both crystal-controlled and variable in frequency.

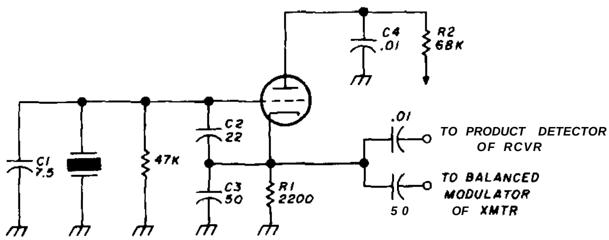
There are several reasons why the Colpitts oscillator is so popular. Mainly, it is stable over a wide range of frequencies. Because a capacitive divider is used, the ratio of feedback voltage remains approximately constant, since the reactance *ratio* between the two capacitors stays the same regardless of frequency change.

Fig. 4 shows the common crystal-controlled version. A triode tube is the amplifying device, although a pentode tube or a transistor could be used just as well. Feedback is developed in capacitive divider C2–C3, and fed to the cathode. The dc connection is typical. Voltage is applied to the plate through R2; the plate is grounded for rf by capacitor C4. If the tube is a pentode, a dc screen supply is provided.

The tube is grid-driven, and output is taken from the cathode. This offers lower output impedance than a plate-output arrangement. In a few models, particularly if a pentode is used, which offers better isolation between input and output circuits, you'll find conventional tuned-tank output arrangements. In the Heathkit linear master oscillator, for example, output is taken from a broadly tuned rf transformer. B+ is fed to the tube through the primary winding of the transformer.

As in all crystal-controlled oscillators, the tuning is accomplished by the crystal itself. The feedback arrangement can easily be designed to force the crystal into operation on an overtone (harmonic), which is desirable in some transmitters and receivers. Sometimes, where cathode bias isn't needed, an rf choke is used in place of the cathode resistor R1. This offers a high impedance to rf, and yet almost no resistance to dc plate current.

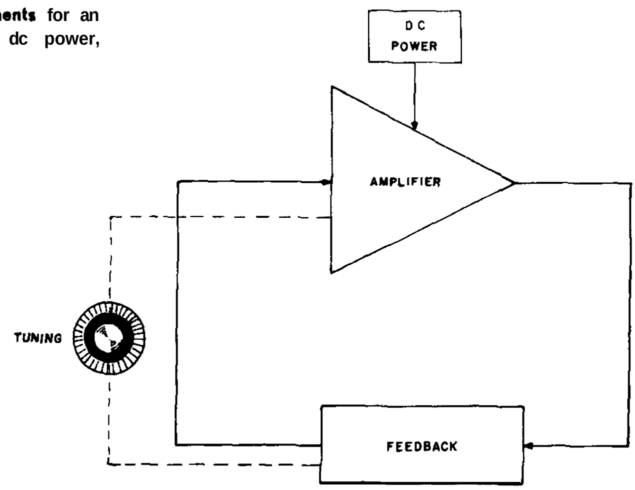
fig. 4. Crystal-controlled Colpitts oscillator.



In some transmitters, the frequency of the crystal is "warped" onto precise frequency by a capacitor—C1 in **fig. 4.** This is done only where frequency is critical, since the "raw" accuracy of a crystal is usually enough for ham work. The capacitor may even be adjustable.

Colpitts oscillators of the tunable variety

fig. 3. Tho four requirements for an oscillator: amplification, dc power, feedback and tuning.



generally use pentode tubes, which offer better input-output isolation. **Fig. 5** shows one of the most elaborate. Besides the basic tunable Colpitts oscillator, special innovations make this crcuit doubly interesting. The exceptional stability and linear operation of this particular circuit over a range of frequencies makes it particularly attractive for linear master oscillator (LMO) service in transceivers.

The pentode is generally a high-mu type with remote cutoff characteristics—the kind used frequently in television-set i-f strips. The 6CB6 and 6BZ6 are popular for this. In one version, the tube is operated as a tetrode, with positive voltage applied to both screen and suppressor grids.

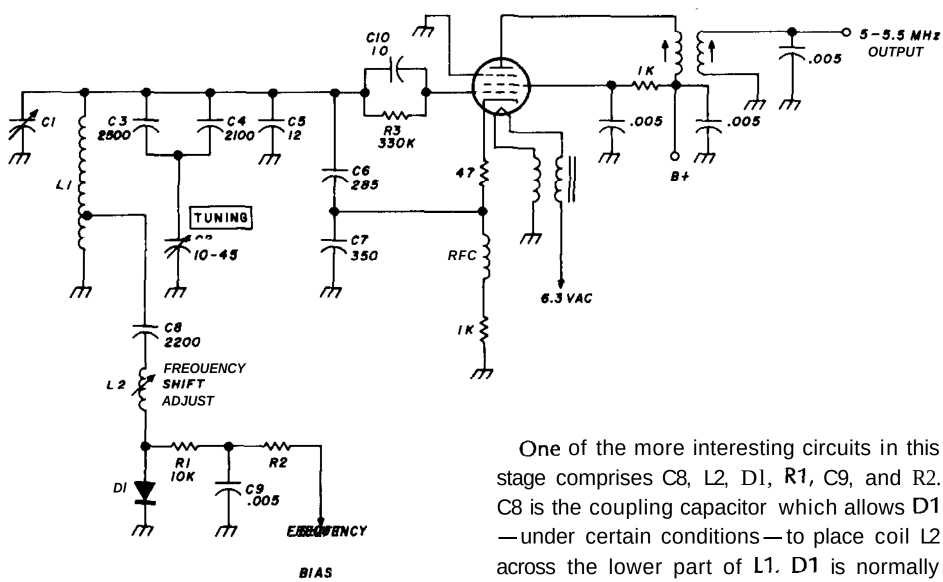
Despite all the elaborate devices for tuning, retuning, coupling, and decoupling, the basic Colpitts configuration is easy to recognize. Capacitors C6 and C7 between grid and ground develop the feedback voltage. The tap to the cathode is the giveaway. An rf choke keeps the cathode well above rf ground, so the feedback can be applied.

An unusual form of grid bias is used in this example, although not in most similar Colpitts circuits. Grid-leak bias is developed in RC network C10–R3.

Capacitor C2 is the main tuning capacitor, and C1 is the trimmer. C3, C4, and C5 are temperature-compensating capacitors; they make sure the oscillator stays at whatever frequency it's set for. To make sure the oscillator does not interact with other rf stages, elaborate decoupling is included. Besides the usual screen-grid and plate-supply decoupling, the filament lead has a bifilar-wound choke. Even when one side of all filaments is grounded, a choke is placed in series with the hot filament lead, with a capacitor to ground.

Output from this version is through a bandpass transformer in the plate circuit. Others use cathode-follower output, and one (in Collins equipment) uses a tapped plate coil and a coupling capacitor.

fig. 5. Variable-frequency oscillator using the Colpitts circuit.

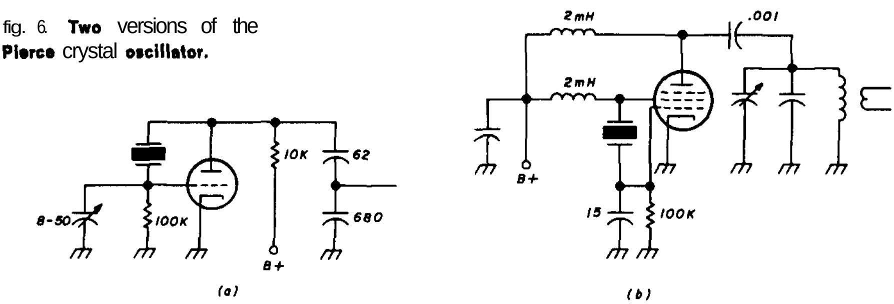


One of the more interesting circuits in this stage comprises C8, L2, D1, R1, C9, and R2. C8 is the coupling capacitor which allows D1—under certain conditions—to place coil L2 across the lower part of L1. D1 is normally reverse biased and therefore offers a high impedance. When frequency-shift bias is applied at the end of R2, however, the positive voltage makes D1 conduct. While it is conducting, it effectually grounds the lower end of coil L2, thus placing it across L1. This change in inductance shifts the oscillator frequency just enough to switch the receiver or transmitter to the other sideband. Normally, the oscillator runs at a frequency that produces upper-sideband operation. When the frequency-shift voltage is applied, oscillator frequency is lowered and operation switches to the lower sideband.

In practically all versions of the tunable Colpitts oscillator, frequency determination is in the grid circuit. The output is broadbanded. In most cases you'll find the frequency range covered by the oscillator is limited, particularly in ssb equipment. The way output frequencies are developed in single-sideband transmitters (by translation) makes it unnecessary for the carrier-generator oscillator to cover a very wide range.

Summarizing, then, you can see that factor No. 1, amplification, is provided by a triode tube in most crystal-controlled Colpitts oscillators and by a pentode tube in most variable-tuned Colpitts circuits. Factor No. *2,* dc power, is generally applied to the plate through a resistor or a transformer winding, even in stages that use cathode-follower outputs. Grid bias may be either by a cathode resistor or by grid-leak bias; in a few it is developed by natural grid current in a high grid resistance. Factor No. 3, feedback, is invariably developed in a Colpitts by a capacitive divider from grid to ground, with the cathode tapped in between the two capacitors. The cathode is kept above rf ground by a resistor or an rf choke. Factor No. 4, tuning, is either by a crystal or a tuned circuit from grid to ground. In the latter case, always keep in mind that the feedback capacitors are in parallel with, and form part of, the tuned circuit. The value of any replacement capacitor in the grid circuit of a Colpitts is quite critical.

fig. 6. **Two** versions of the **Pierce** crystal **oscillator.**



second most popular—the Pierce

This oscillator is popular because of its simplicity and stability. It uses a triode or a pentode, connected normally, with a crystal providing both feedback and tuning. The crystal is connected from plate to grid with a triode, and from screen grid to control grid with a pentode. Because of the accuracy and resonant efficiency of a crystal, a Pierce oscillator holds its frequency well over wide variations of dc input voltage. It isn't likely to drop out of oscillation unless plate or screen voltage becomes extremely low.

You'll find the Pierce in both transmitters and receivers; it's often used as a heterodyne oscillator for raising frequency in transmitters, and as a frequency-conversion oscillator in receivers. The simplest version, a triode, is shown in fig. 6A. The only elaboration is a frequency-warping capacitor connected between the grid end of the crystal and ground. It permits fine adjustments of the crystal's resonating frequency.

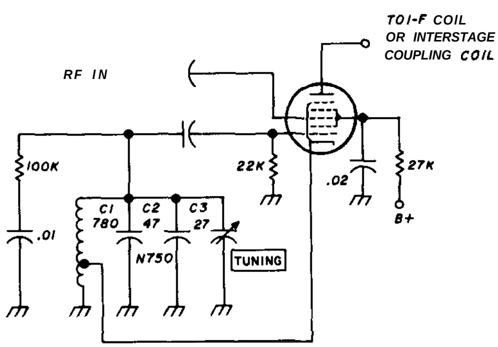
The output of a Pierce oscillator is usually rather strong. This is the reason for the capacitive-divider output network, which gives a ten-to-one reduction in rf voltage fed from this particular oscillator. This version is used in some receivers as a frequency-control oscillator, with output fed to the second mixer, and in at least one transmitter as a heterodyning (frequency-translation) oscillator.

Fig. 6B shows a pentode version of the Pierce oscillator. The dc supply is conventional; some versions use cathode bias while others have the cathode grounded. In either case, the cathode is always kept grounded for rf.

Output is from the plate. Since the tube is a pentode, the plate is isolated from the frequency-control network. The output arrangement shown is a little off-beat; it is called impedance coupling. The choke from plate to B+ is the untuned plate load. A coupling capacitor feeds the rf voltage to the tuned circuit, part of a transformer. Coupling to the next stage is inductive. Another version uses simple RC coupling—a resistor supplies dc plate voltage and acts as output load, with a capacitor coupling the rf signal to the next stage.

Both plate and screen in this stage are fed through rf chokes, which offer some load for rf developed in the plate-current stream of the tube. The screen grid may sometimes have a capacitor tying it to ground, but is

fig. 7. Although this Hartley oscillator is part of a pentagrid converter, the same circuit may be used with an electron-coupled oscillator.



very seldom completely grounded for rf; if it were, feedback couldn't take place from the screen grid to the control grid through the crystal. In the version shown, the screen is not bypassed at all; only a small stabilizing capacitor is connected between the control grid and ground.

There is one version that operates as a cathode follower, with the output tuned circuit in the cathode circuit. Output arrangements have no bearing on the "type" of oscillator. The Pierce gets its identification from the fact that it is controlled by a crystal between the plate and grid. In pentode versions, the screen grid is operating as a plate, not as a screen grid in the usual sense. (That's why it's not thoroughly bypassed for rf.)

the tapped-coil Hartley

This oscillator is distinctly recognizable because the tube's cathode always goes to ground through a tap on a coil (see **fig. 7).** The other end of the coil almost invariably is connected through a capacitor to the grid. The Hartley oscillator is uncomplicated and stable and is used extensively for tunable applications. There is a crystal-controlled version, but it is rarely used in modern ssb equipment.

The version in **fig.** 7 is part of a pentagrid converter; the same circuit can be used as an electron-coupled oscillator. The Hartley is found in both receivers and transmitters. In one transmitter, it is the first carrier generator, operating at 60 kHz; in another, the second conversion oscillator, operating at 395 kHz.

The oscillator plate in the tube of **fig. 7** is the double grid, grids 2 and 4. This oscillator plate doesn't have to be left ungrounded for rf, since the control grid modulates the entire electron stream. The rf from the previous stages—from a station or a transmitter stage—is fed in at grid 3. Grid 4 (part of the double grid) acts as a shield for the rf input grid, very much like the screen grid in an ordinary pentode.

The tapped coil that sets up the feedback is always a part of the tuning circuit. The tank capacitors include tuning capacitor C3, temperature-compensating capacitor C2, and main frequency-determining capacitor C1. All affect frequency. In some circuits, C1 may be a trimmer, and occasionally the temperature-compensating capacitor is omitted.

the rest of them

Another simple oscillator used in single-sideband equipment is the tuned-plate-crystal-grid oscillator, sometimes called simply **tpxg.** This one commonly appears as the first receiver oscillator in double- or triple-conversion receivers—named, in that application, the heterodyne oscillator.

Triodes are always used, because tpxg oscillators depend on interelectrode capacitance for feedback. The screen grid in a pentode would shield out this Miller effect and keep the tube from oscillating. The crystal is connected from grid to ground. A frequency-warping capacitor or coil can be used with the crystal, although it seldom is.

The dc voltages for the tube are conventional. The cathode may be either grounded or above ground (for cathode bias). In most tpxg oscillators, bias is in the form of grid-leak or "contact" bias across a fairly large-value grid resistor.

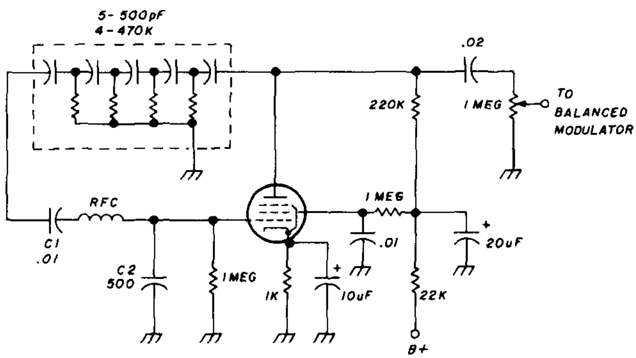
Output circuits vary just as much as with any other oscillator. Sometimes a broadband coil is used as the load, with a small capacitor coupling the oscillator output. Occasionally, inductive coupling is used, again with a broadband transformer.

This kind of oscillator is especially suitable for overtone operation — running the crystal at (usually) three times its normal frequency. This kind of operation may be necessary in the front end of a ham receiver, and that's why this circuit is not at all uncommon as the up-front heterodyne oscillator. Sometimes, particularly when cathode bias is used, the crystal is separated from the grid by a blocking capacitor. If grid-leak bias is used, there is no need for this.

Occasionally, there is reason to use a tone in a single-sideband transmitter — for testing or for A2 code transmission. **Fig. 8** shows an 800-Hz phase-shift oscillator. There are any number of other audio oscillators that can be used, but this one is exceptionally stable and its frequency is independent of output load. This latter characteristic makes it excellent for tone keying.

Feedback is via a printed component circuit (PEC) consisting of five capacttors and four resistors. As they are arranged, they give a phase shift that is almost adequate to sustain oscillation at 800 Hz. Capacitors C1 and C2 and the rf choke complete the job. The result is extremely stable audio oscillation. The output is RC-coupled to a buffer amplifier via a volume control that determines how much modulation will be developed.

fig. 8. An 800-Hz audio phase-shift oscillator.



what to look for next month

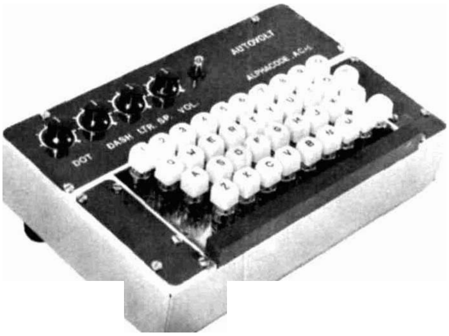
That's the story of oscillators in single-sideband equipment. There are other circuits just as interesting, and often more difficult to understand. We plan to cover most of them in this series on single-sideband.

The method of single-sideband generation in most of today's ssb transmitters is the **filter** method. A balanced modulator, which has already been explained, forms a double-sideband suppressed-carrier signal which is then pushed through a filter that removes one or the other of the sidebands. However, there's another way, called the **phase-shift** method.

Although the phase-shift method is almost never used in commercial ham gear, many readers have wondered about the principles behind it. It is sometimes less expensive than the filter method — particularly at high frequencies, because there is no need for stages of frequency translation. If the suppression ratios were improved, and the circuits made easier to adjust, phase-shift ssb might catch on. Next month, we'll explain how the method works, and give some pointers on making adjustments to this type of ssb generator.

ham radio

NEW!



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MODEL 22 . . . . . $25

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The new model 500C is the latest evolutionary development of a basic well proven design philosophy. It offers greater power and additional features for even more operator enjoyment. Using a pair of the new heavy duty RCA 6LQ6 tetrodes, the final amplifier operates with increased efficiency and power output on all bands. PEP input rating of the 500C is conservatively 520 watts. Actually an average pair of 6LQ6's reach a peak input of over 570 watts before flattopping!

The 500C retains the same superior selectivity for which Swan transceivers are noted. The filter is made especially for us by C-F Networks, and with a shape factor of 1.7 and ultimate rejection of more than 100 db, it is the finest filter being offered in any transceiver today.

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Voice quality, performance and reliability are in the Swan tradition of being second to none.

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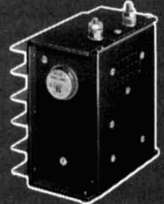

SWAN 14C DC CONVERTER

Converts the above 117XC A.C. power supply to 12 volt D.C. input for mobile, portable, or emergency operation.

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POWER

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Our improved standard 5 band model, now in production and still only. . .

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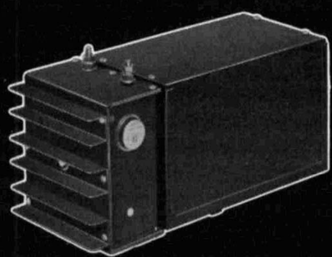

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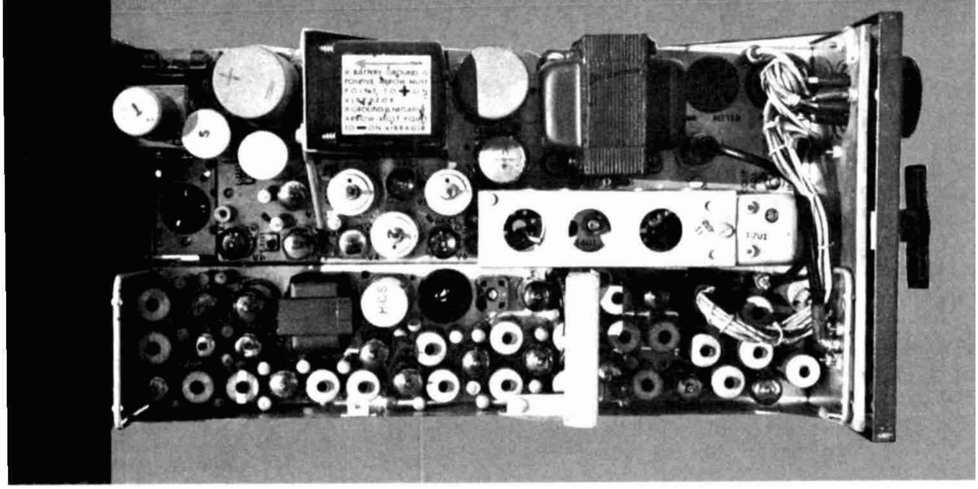


amateur vhf fm operation

An introduction to amateur fm operation on 144 and 432 MHz, including base stations, repeaters and accessories

John Connors, W6AYZ, 920 Sir Fran is Ave. Capi la, Cal ifrn'a 947 0

This article was written to introduce you to the wonderful world of amateur vhf fm communications. There are a number of hooks on the market which will explain all the technical aspects of the subject to vou; the purpose of this article is to consider fm seriously as a dependable, local communications mode with many advantages over a-m. You will meet some of the various types of equipment in common use and some of the fascinating mutations rleveloped by commercial and amateur users in their quest for reliability and convenience.

Amateur fm owes its very existence to the commercial fm two-way radio used by the mobile services such as police, fire and taxi. The first equipment that the commercial users used on a large scale was designed for wide-band fm. The transmitter was deviated plus or minus 15 kHz from the center frequency. Most of this early equipment was also designed for a six-volt supply.

Commercial two-way radio caught on big however, and the FCC was forced to alter their requirements to what is now called narrow-hand fm; the transmitter is only deviated plus or minus 5 kHz from the center frequency. This rules change provitlcd more channels through closer spacing and made all the old equipment obsolete. To help matters further, the automobile manufacturers decided that twelve-volt ignition systems were more practical than six-volt systems. Result: more obsolete commercial fm equipment. Now, vacuum-tube equipment is bring phaced out by solid-<fate units. The result is a lot of low-cost frn equipment for the interested amateur.

why use fm?

As you may have guessed, the first big advantage of frn for the ham is availability. Commercial users couldn't use much of the older equipment, and no one was willing to buy it. As a result, many equipment owners and manufacturers turned to the amateur market as a way of saving some of their original investment. The second advantage is price. For the price of a good a-rn transceiver, you can buy four or five fm transceivers with high-quality receivers and transmitters

The Link decoder. Although these units are stubborn at times, they a n often used as a command decoder and work quite wall as a selective-call decodar.

For mobile operation, frn units mount in the trunk and are remotely controlled froni the front seat For weak signals, fm is always more understandable than a-m Receivers won't drift off frequency since since they are crystal controlled Positive noise-sensing squelch circuits surpass anything found in a-m gear and completely silence the speaker until a signal is received.

Ignition and other forms of interference which cause a lot of trouble with a-ni operation don't affect fm; a typical mobile installation rarely needs any noise suppression. Another big advantage is reliability. Commercial equipment is built to run continuously and take a lot of abuse while doing it. Mobile units can bounce around in a trunk for years with no failures, and base stations require little maintenance.

how to get started

You can begin your search for fni equipment and information by writing to the Lynchburg Amateur Radio Club* for a list of active fm nets or groups in your area. The California Amateur Relay Council** will provide legal and technical advice on repeater operation. There are also a couple of books[1,2] and a magazine[3] which are helpful. After you're armed with this information, contact the people in your local area who use the equipment (both amateur and commercial) to find out what is available and where.

Speak to the man who runs the local fm two-way shop. hlany times he has a customer who wants to sell obsolete fm equipment; as a favor to his customer he will get you together. Talk to commercial users directly and leave your name and address around, especially at non-governmental businesses. They are not as involved in red tape as the police, fire and county facilities. You may even find someone willing to sell a group of five or ten units at one time. In this case, it's best to have a group of people go together and split the cost. Prices are generally better and this guarantees that you won't be the only fni station in the area. In any case, don't give up. Acquiring the first few units is the hardest part. Once the hall is rolling, and your group expands, you'll find availability is good.

equipment

Before discussing the establishment of your fm operation, a brief rundown on equipment is in order. Two-way fni equipment falls in three major catagories: low band (30-50 MHz), high hand (150-170 MHz) and uhf (450

* Lynchburg Amateur Radio Club, 1306 Grove Street, Lynchburg, Virginia

** California Amateur Relay Council, 51 Norwood Avenue Kensington California 94707

MHz). Low-band and high-band equipment is similar in design and appearance. Equipment for uhf is harder to come by and is generally used for repeater control and backbone links.

High-band equipment is the most popular for a number of reasons. First of all, antennas are less cumbersome and objectionable at the shorter wavelengths. Secondly, television interference has always plagued six-meter activity and two meters is relatively free of those problems. Thirdly, in most cases, two-meter equipment is ready to run as is; equipment that does require modification is taken care of with padding capacitors. Low-band equipment must be moved up beyond the 52.5-MHz band edge for wide-band operation. This is generally no big problem, but it can be quite involved.

In all fairness to six meters, it does provide greater range capabilities and less objectionable mobile flutter than two. This could be a major consideration for simplex operation. In addition, some low-band equipment can be converted to ten meters. This is becoming increasingly interesting for fm with the improved band conditions, and 29.6 MHz is a very active fm channel.

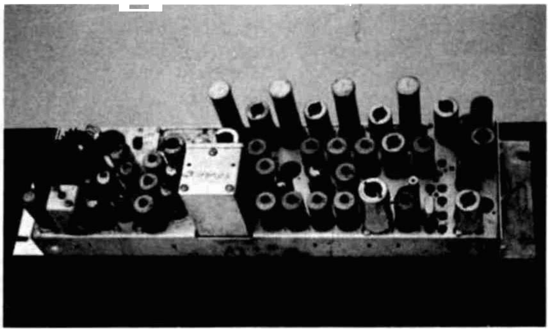
Motorola equipment is quite representative of the trends manufacturers followed over the years. It is also quite popular among hams, so it makes it a good example for discussion. Motorola started out with the **Deluxe** line, a separate transmitter and receiver housed in big, blue containers. They are old, bulky, inexpensive and ever so reliable. Some of these units were made with 110-volt supplies for base station use.

Next came the **Dispatcher** series, commonly called the 5V. These were small, single-case units which had six-volt supplies. Sometimes you'll find one which was converted to 12 volts; the transformers are still available for the 12-volt modification. The receivers in the **Deluxe** and **Dispatcher** lines are not well-suited to multiple-channel work since the same oscillator is used for both high and low i-f injection.

After the **Dispatcher** series, Motorola started its **Research** line which is still being expanded. The 41V became the most popular in this line with its six- or twelve-volt supply and multiple-frequency receiver and transmitter. Some units even had 110-volt supplies for base-station use. The basic unit is slightly larger than the 5V. Besides the 41V, the **Research** line contains a wide variety of transmitters, receivers and power supplies mounted on strip-type chassis. They can be rack mounted or put together in various combinations in a mobile housing. The best way to become familiar with these units is through the **FM Schematic Digest."**

When transistors came into service, Motorola's line of equipment was expanded with solid-state units of all kinds. Hybrid handie-talkies like the one shown in the photograph were partially transistorized at first. Many used rechargeable Nikad battery packs; this makes them very economical for amateur use. Pocket-sized transmitters and receivers followed, but they are still pretty scarce. Motorola's main line of mobile equipment went miniature with the advent of the **Motrac.** This is a hybrid unit with various transmitter powers, multiple-frequency kits and extra gadgets.

A 1-44 receiver strip. Excellent for the 450-MHz primary receiver of a repeater.



The other manufacturers have gone the same route over the years. General Electric had their **Preprogress, Progress** and **Master** lines. Link is also used quite a bit by amateurs as is RCA, Comco, Karr, and others. Generally speaking, most manufacturers produced models for both low- and high-band operation in a particular line, but this isn't always true.

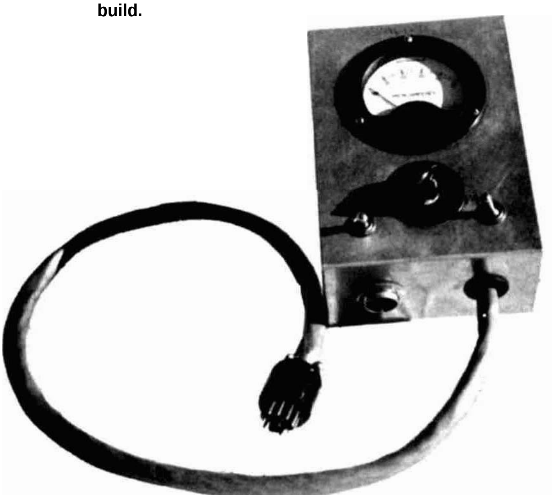
establishing an fm operation

There are several very important considerations you and your group must make before becoming deeply involved in fm. First is the problem of choosing a frequency. Before you start buying crystals, stop and consider these points:

Do any other amateur activities frequent the channel you intend to use?

Would it be advantageous for the group to use a frequency in the general-class portion of the band to minimize possible interference?This is especially important for repeater users since the repeater hears a lot more than a ground station.

A homemade tart aid for servicing almost any Motorola fm equipment. Adapters for the remaining units are easy to build.



Should you make use of some of the present national fm frequencies as your channels for emergency operation?

It should also be kept in mind that once a number of fm stations are set up on a particular channel, it is nearly impossible to change it. Consult the Lynchburg group regarding the channels being used in your area; you may want to join one of them. Monitor your choice of possibilities first and be sure nothing happens which could make you select another channel.

Once the frequency has been chosen, there is the problem of accurately measuring it. Although the transmitters are crystal controlled, they are adjustable over a few kHz and as Aagaard points out?, disagreements are sure to arise. Transmitters can he set to the receiver frequency by zeroing the discriminator in the receiver. All discriminators should be zeroed to a common rf carrier. When aligning the reference receiver, use the most accurate signal source available. An accurate frequency meter or counter is a big help in this area.

Point-to-point operation is called local channel or simplex operation and works quite well. For larger coverage, amateurs have gone to repeaters.[4] Repeaters require two methods of control which must be entirely independent. By using dialed tones transmitted to the repeater on the communications channel, secondary or member control is maintained. This is the method used by club members to turn the repeater on and off.

The trustee of the repeater must be able to turn the repeater completely off by using a remote-control system on 220 MHz or higher. His system is only used in emergencies such as equipment malfunctions. This is called primary control. Generally, a 450-MHz fm system is used for primary control because the equipment is more readily available than 220-MHz equipment and takes much less effort to convert.

450-MHz equipment may also be used as a backbone link between two repeaters. This is a two-way relay system in which the audio from one repeater is fed to the other via a 450-MHz link where it is retransmitted on vhf. This allows mobiles or base stations to communicate dependably over great distances. The 450-MHz receivers continuously monitor the channel, and the transmitters are turned on by command from the ground stations. This eliminates any need to monitor the other repeater continuously.

accessories and ideas

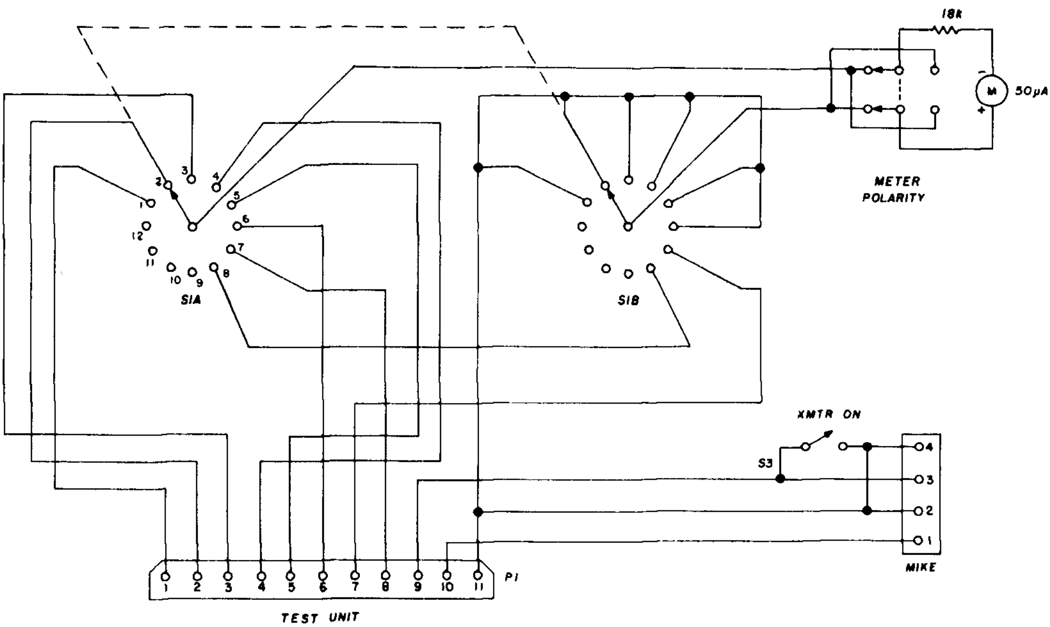
The following are some of the more popular accessories and ideas being used by fm amateurs around the country. This list is by no means complete in scope or content and is offered only as a stimulus to thought.

Tone oscillator. A tone oscillator can be the right arm of an fm amateur. By placing a telephone dial in series with the audio output, it can be used to control a repeater or selective-call system. Generally, provision is made for selecting a number of different preset frequencies with a rotary switch. Special oscillators are used for generating AFSK Teletype and facsimile signals as well as transpond and telemetry signals.

Selective-call decoder. For continuous channel monitoring with the speaker off, a de-

Private Line is a good example. This is somewhat similar to selective call except that it isn't as selective. To open the speaker on your receiver, two conditions must occur simultaneously. First, there must he a carrier present, and second, a short tone burst of a predetermined frequency must occur. The receiver speaker will then enable until the carrier disappears. A small encoder is used at the transmitter and a decoder at the receiver.

fig. 1. Schematic diagram of a homemade test aid for Motorola fm equipment.



coder is wired into the base station. The call is placed by dialing a series of digits which add up to ten. This completes a circuit which rings a bell and turns on an indicator lamp. Each channel user with this capability should choose a different four-digit number that adds up to ten such as 1-2-3-4, 1-1-7-1 or 3-4-2-1. Some amateurs transmit a short burst of tone back to the caller to tell him that the alarm did go off. This is called transpond and can even be used to indicate whether or not the party being called is home by transponding with different audio frequencies.

Tone access. Commercial manufacturers make tone-operated squelch systems. Motorola's

Autostart Teletype. For general information bulletins, passing large volumes of information and leaving messages for people not listening, autostart Teletype is very useful. This scheme was discussed in a recent magazine article" along with circuits for a complete transistorized system. The most popular machines in amateur service are the Teletype models 26 and 15. The model 26 is a small, quiet unit while the model 15 is a rugged, dependable machine. There are many sources for RTTY gear, and the prices vary, so shop around.

Facsimile. Repeater users soon discover that they are doing a lot of running back and forth exchanging manuals, schematics and so on. Because of the large coverage area of a repeater, one trip to get a schematic can cost a full day. By using facsimile equipment, pictorial information can be passed hundreds of miles in a matter of minutes. While facsimile equipment is less abundant than Teletype machines, these units can be obtained quite readily on the surplus market.

Telephone and desk remotes. Remote control of a base station from another room of the huilding is quite popular with commercial users. Many manufacturers make a desk-top base station, but it is senseless to run two stations at once. By rewiring an old telephone, a versatile remote can be made for practically nothing. A miniature push-button switch installed in the handset converts it to push to talk; coil cords with five conductors sell for under two dollars. The volume control can be hidden behind the handset cradle and other controls similarly placed. An alternative solution is to mount a speaker and

Typical of equipment available to the ham is this Motorola 5V. They are compact and inexpensive.

microphone in a small hox which can be holted to the wall in some convenient location. Most amateurs who use remote keep the base station in the shack or garage and use as many remotes as they want in the house. Don't forget to bring in a selective-call wire for the alarm so you will hear it when it goes off.

Receiving. Most active fm amateurs leave their receivers running continuously day and night. While this does consume some electricity, it makes operation much more pleasant and convenient.

When operating through a repeater, the receiver can be left running while transmitting. This lets you hear yourself coming back through the repeater. If another operator wishes to interrupt or inject a thought, he merely pushes his mike button and the beat note is heard in your receiver. This type of duplex operation makes communications just like talking on the telephone.

repeaters

Besides the normal repeat functions, the following additional features are quite popular with amateur repeaters of all kinds:

Frequency set. By dialing a command, a highly stable crystal-controlled oscillator is enabled. Any station placing a carrier on the input and duplexing can hear the beat note. He can place his transmitter on frequency by zero beating to the reference signal.

Signal-strength monitor. This device may be left on all the time or turned on by command. It delivers an audio beep sequence which indicates proportional signal strength of the carrier present at the repeater's receiver.

Secondary-control cycling timer. This is simply a timer which requires setting within a certain period of time or the repeater is automatically shut off. This is handy for mohiles that inadvertently drive out of range.

CW identifier. Elimination of repetitious voice identification of the repeater is accomplished through automatic CW identification. The usual procedure is to cycle the identifier once when the repeater comes on and once when it goes off as well as every ten minutes in between. Either mechanical or electronic means is employed to generate the MCW tones.

Directional antenna rotation. Some repeaters incorporate facilities for remotely directing antennas to assist in copying weak stations in distant locations. A simpler method is to use separate antennas pointed in the desired direction; they are selected by command.

Emergency repeater or portable power. Some groups will prefer to install an automatic generator at the repeater site for ac power during commercial power failures. Others choose to install a battery-operated repeater which comes on during power failures. The generator has the advantage that normal repeater operation is not affected. However, it requires continuous service as well as maintenance and fuel. The battery-operated repeater gets away from all that, but power capability is limited, and only skeleton operation is possible.

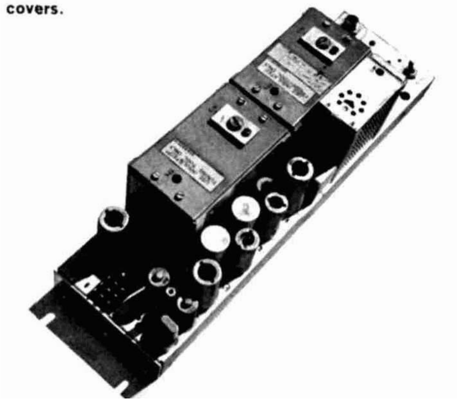
Multifunction decoder. The ability to perform many different remote-control functions upon command is a must for some repeater groups. Some have only two commands, repeater on and off, while others have facilities which can handle up to 1000 different commands. Diodes, relays, transistors and tubes all play an important part in units such as these, and each group must assess its own

A Motorola 41-V. Slightly larger than the 5V and with better features which make it worth owning. Many of these units a n coming out of commercial service now.

needs and potentials. For the group just getting started, a simple command decoder is a modified Link or Secode. These can provide enough commands for any club's beginning needs, but you'll soon discover you're only getting by and are actually quite limited.

Alternate receivers and transmitters. Endless combinations of frequencies and modes can be used to avoid conflict with local amateurs who unknowingly fall on the main input channel or for other special purposes. Most popular is a second-channel monitor which listens to another fm channel and "pipes" the audio out over the repeater at reduced level. Many repeater groups are installing facilities for monitoring and transmitting on one or more of the national fm calling frequencies such as 52.525 or 146.94 MHz. These frequencies can be used for traveling mobiles or local co-ordination activities such as interstate traffic handling.

A 450-MHz transmitter from the T-44. Power plug is on left and finals and drivers are under the covers.



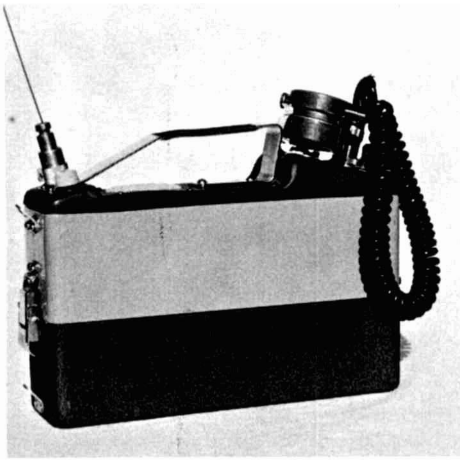
Telemetering. When many commands are available for use, it's difficult to remember which ones are on and which are off. Remote metering via telemetry overcomes this, and there are a number of different approaches. Slow, repeating "beeps" of three different audio frequencies can be read by ear and decoded by counting their position with respect to a sync pulse of a different frequency. Another method generates a dot pattern on an oscilloscope, and intensity modulates the dots for those commands which are on. Amateurs who are interested in television might consider transmitting a television picture of a lamp panel at the repeater site down to the base station.

Tape logging. Repeater logging can be eliminated through the use of recording tapes. A short burst can be made to start the recorder for a predetermined interval of time, or the recorder can run continuously. The advantages and disadvantages of each system are obvious.

Tape-message secretary. A tape recorder will record anything on the repeater input for a predetermined time on command. When another command is given, the message can be played back.

I have deliberately ignored the details of

A battery-operatad **Handie-Talkie. One** watt **transmit** with **recharge-able battery,** speaker and micro-phone.



these ideas since most fm groups will have their own ideas and will want a system engineered to their own needs. This is the best part of fm radio. Transmitters and receivers, the major concern of the low-band ham, become only secondary to the vhf fm operator. His major interest lies in the development of a sophisticated communications system, and fm radio is merely a stepping stone to that goal.

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ham radio

solid-state transmitter switching

If you're tired of holding that PTT switch down, try one of these push-to-talk, push-to-listen circuits described by W2EEY

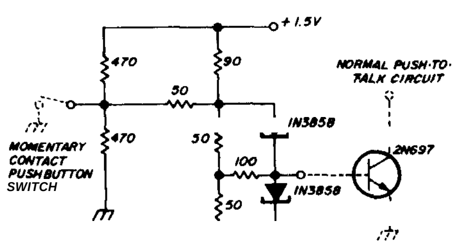
John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 06355

Sometimes the simplest gadgets contribute the most to your operating convenience. For example, if you replace the push-to-talk (PTT) mode of operation with a push-to-talk **and** push-to-listen (PTT/L) circuit, you don't have to continually push the mike button during transmit periods.

Amateur operators used to have a simple toggle switch for send-receive switching; push-to-talk was confined to mobile installations. Today, however, transceivers are in the vogue, and push-to-talk operation is used by many stations. The only disadvantage of PTT is that you have to depress the PTT button continuously during transmissions. Your hand can get pretty tired during long operating periods. VOX operation is very convenient, but not all transceivers have VOX built in.

In this article I'll describe some simple ways to convert your rig to push-to-talk and push-to-listen. With this setup, the mike button is pushed momentarily to transmit and then released. When it's depressed again, the rig is switched to receive. The advantages

fig. 1. A tunnel-diode circuit which is the solid-state equivalent of a latching relay.



may not appear very great, but try to imagine how much easier many QSO's would be if you didn't have to use a death-grip on the mike continuously!

Although you can use relays or special switches to obtain push-to-talk/-listen (PTT/L) operation,[1] the solid-state circuits described here are a lot more reliable. The use of integrated circuits to replace a relay should interest you even if you don't want a PTT/L circuit in your rig. These switching functions can also be adapted to other uses, such as switching filters or crystals.

solid-state switching conversions

For PTT/L, you have to have a solid-state circuit which will duplicate the action of a "double-push" switch or a latching relay. This can be done electronically with a triggered bistable multivibrator. This term may be confusing, but the basic multivibrator circuit has been used by many amateurs as a CW monitor. The triggered circuit will change "states" (that is, have an output or no output) only after a triggering pulse has been momentarily applied to the input. Each successive triggering pulse changes the state of the multivibrator.

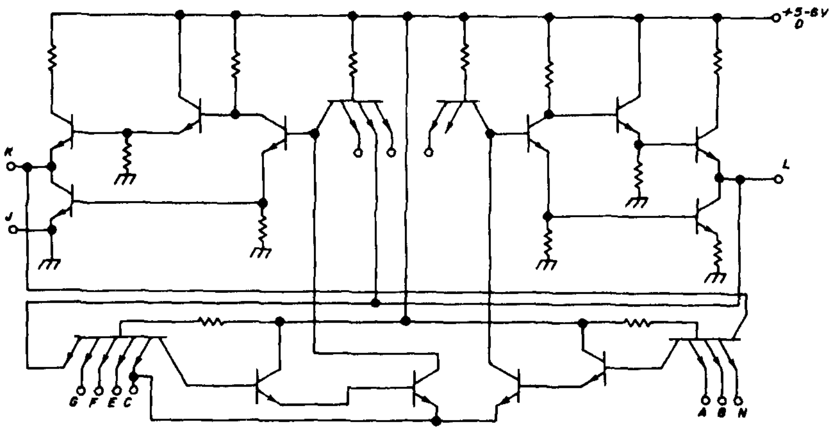
It is fairly easy to convert a free-running multivibrator into a triggered multivibrator. However, succeeding pulse inputs to trigger the circuit have to be opposite in polarity. You can add circuitry so that successive pulse inputs of the same polarity will trigger the multivibrator, but the added components result in some rather complicated circuitry for the simple switching function you need.

tunnel-diode latching circuit

The two tunnel diodes shown in the circuit of **fig. 1** provide the equivalent circuit in a relatively simple form. Each time the input terminal is grounded, the output terminal alternates between a positive 1.5-volt output and a zero-volt output. The output of this circuit cannot be loaded appreciably if it is to function properly. It is better to drive a switching transistor which activates the actual PTT circuit in the transmitter. It can also be used to pickup a SPST relay which then controls PTT action.

There is nothing complicated about the operation or construction of this circuit. If you can buy the diodes reasonably, this circuit is about the simplest electronic latching circuit you can use. For more information on tunnel-diode switching circuits, consult references 2 and 3.

fig. 2. Internal circuit of en integrated-circuit J-K flip-flowthis circuit can be used in place of a mechanical latching relay as described in the text.



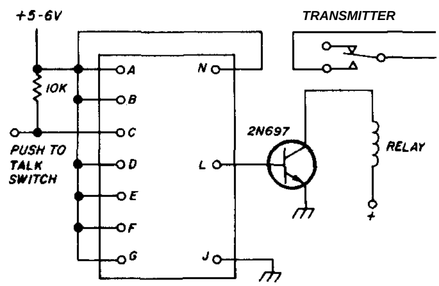
integrated-circuit switch

Another approach to inexpensive solid-state switching involves the use of an integrated-circuit J-K flip flop. The internal circuitry of the IC is shown in fig. 2. With this device, negative-going input pulses of the same polarity will produce alternate "on" or "off" states at the output. The integrated circuit I used was a surplus unit available from Solid-State Sales* for $1.20.

Rather than becoming too concerned with the internal circuitry of the IC, let's examine the switching condition that will provide PTT/L operation. If a positive 3 volts is present at all six terminals EFC and ABN, every time the input voltage at C goes from 3 volts to 0 volts (but **not** from 0 to 3 volts), the voltage at K and L will reverse. For example, if K is at 3 volts and L is at zero, momentarily grounding C will change K to zero and L to 3 volts.

Fig. 3 shows the J-K flip flop in a PTT/L circuit. The integrated circuit cannot be loaded down too much, so a switching transistor

fig. 3. Connecting an IC and 2N697 relay driver to provide push-to-talk and push-to-listen operation.



is used to control the relay circuit. An NPN transistor is shown here because a positive control voltage is usually used to activate the send-receive relay. If your transmitter happens to use a negative control voltage (very unusual), simply reverse the emitter and collector terminals of the switching transistor. For added flexibility, you can add a bypass switch so that either PTT/L or conventional PTT operation is possible.

There are only two precautions when using this circuit: don't exceed the recommended supply voltage, and make sure the PTT switch provides a good, clean contact. The IC responds extremely fast, and a dirty switch will cause it to change states unexpectedly. Otherwise, its switching action is far more reliable than any mechanical mechanism.

* Solid-State Sales, p. O. Box 74, Sornrnerville, Massachusetts 02143; order FF-1 J-K flip-flop integrated circuit.

switch replacement

If you feel you have to use a mechanical PTT/L circuit, sometimes the mike switch can be replaced with a so-called "push-push" or "double-push" action switch. These switches will alternately open or close a circuit with alternate depressions of the switch. A double-push switch that will fit in the base of home station microphones is the Cutler-Hammar model 7208-K4. Arrow-Hart and other manufacturers produce similar types. A miniature double-push switch which lends itself to hand-held microphones is the Alco-switch MSP-105D. It is less than 1/2-inch square and costs under $2.00.

For those situations where a momentary action switch can be replaced with a double-push switch, this may be the easiest conversion scheme. However, in most cases, solid-state methods of switching are usually preferable. Not only do they result in push-to-talk and push-to-listen operation with any PTT microphone, with a simple bypass circuit you can have simple push-to-talk operation too.

summary

Push-to-talk and push-to-listen operation is certainly not a major technical innovation. However, it has proved to me to be one of those small operating conveniences that becomes invaluable after it has been used for awhile. Try using push-to-talk and push-to-listen operation for an hour or two, and then go back to plain push-to-talk; the inconvenience you've been putting up with will be quite apparent.

references

1. D Collins, VE3GLX, "Push to Talk or Speak to Talk," 73, October, 1967, p. 98.
2. RCA, Technical Manual TD-30, "Tunnel Diodes," 1963, pp. 33-54.
3. "GE Transistor Manual," Seventh Edition, 1964, pp. 366-371

ham radio

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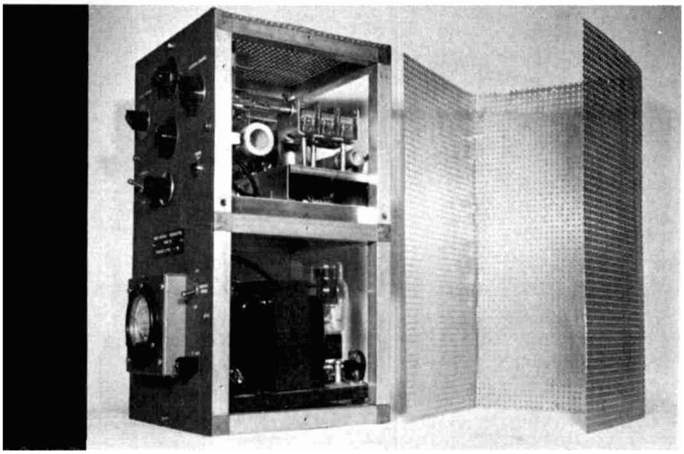


A three-band 80-watt transmitter using modem rack and panel construction.

return to rack-and-panel construction

Here's a construction technique that has all the advantages of the old rack and panel but results in neat, compact designs

Howard S. Pyle, W7OE, m :44 74th Avenue, S. E. Mercer Island, Washington 98040

Grizzled old timers will nostalgically recall the pioneer days of arnateur radio when a station was a scattered collection of motley components arranged as fancy dictated. A tuning coil or loose coupler, a crystal detector, perhaps a variable capacitor of sliding metal plates in a grooved wooden frame arranged on a table top constituted the receiver. A massive helix of copper tubing, a huge capacitor of discarded tin-foil-coated, photo plates, a heavy transformer arid an awesome rotary gap somehow found space on the floor, the table or a shelf—wherever they would fit—made up the transmitter. Odds and ends of hell wire, lamp cord and copper tubing tied the scattered pieces together electrically and you were on-the-air!

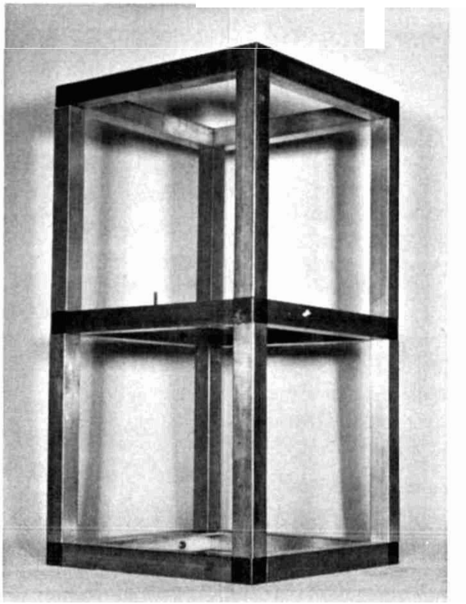
Because of the inconvenience and space-hogging of such arrangements, the receiving elements were soon combined in a more compact group. Generally this took the form of a breadboard, with all the components screwed down. Sometimes a wooden cabinet was used. The tuning coil or loose-coupler was mounted on the cabinet top, rotary variable capacitors were mounted in the box with their shafts sticking out and fitted with homemade scales and typewriter knobs.

Transmitters were another story. A box the size of a coffin was too small to house the massive items making up the more ambitious spark transmitters. Commercial companies began to group their components for ship-hoard transmitters on slate or hard-rubber panels—one manufacturer even used marble! These panels were mounted on a vertical metal framework of angle iron.

The dawn of the vacuum-tube era ushered in a new day. Receivers rapidly progressed to cabinet-style housings with controls on the front panel. Panels progressed from wood through hard rubber to bakelite, often backed with a metal sheet nr tin foil to eliminate body-capacity effects, and finally to all-metal panels. Cabinets weren't far behind, and metal cabinets with matching panels began to appear on the market.

How about transmitters? In the 200-meter

fig. 1. The basic rack framework is built from snap-together aluminum extrusions and castings.



days, the components were rather massive because of the necessity for large spacing to avoid the possibility of flash-over with the high voltages involved. A single cabinet was simply impractical. Vacuum-tube transmission significantly reduced the size of transformers, inductors and capacitors. Again the rack-and-panel scheme looked good. Except for the kilowatt stations, the average low-powered ham gear no longer required a floor-mounterl rack and panel the size of a small closet.

A simple, easily constructed wooden or metal frame occupying about a square foot of table space and standing only a couple of feet high could be used The idea rapidly caught on; manufacturers started offering metal racks of various heights. tlrilled for matching panels. Commercial companies adopted a standard panel width of 19 inches and these were available in heights ranging from 1-3/4 to 24 inches in increments of 1-3/4 inch. Later, cabinet racks became available in various heights up to six feet or so.

People have a strong tendency to follow the leader; if an electronics manufacturer encloses his equipment in an attractive metal cabinet and the design catches on, other designers follow the same general pattern. As a result, today's receivers, transmitters and transceivers are almost invariably housed in handsome, compact metal enclosures. This is fine for the ham who is content to place a nice-looking piece of gear on his hand-rubbed, hardwood operating desk in tasteful surroundings, plug it in, twirl a few dials and get on the air. Most of this equipment is pretty reliable; it will zit there and perform for a long time with a minimum of maintenance. But what about the ham who doesn't want to be a push-button operator? There are a lot of them and thev like to experiment—try this and that, change a switch, a coil—maybe increase power. They hesitate to butcher a piece of factory gear; modifications generally destrov its resale or trade-in value.

The answer? Homebrew of course! Rut why stick to convention? Hams have always been known to be ingenious, resourceful and quick to come up with new circuits, new ideas. Then why develop a novel electrical design only to become a conformist in the final stage—the housing? Almost invariably, homebrew equipment winds up with the components buried in the bowels of a shallow metal chassis or mounted on a permanently attached panel.

You have to he a contortionist to wire such an arrangement. Poking a hot soldering iron down in the recesses of a crowded chassis generally produces burned insulation, scorched fingers and a string of cuss words!

Changes and modifications which occur from time to time are worse. Can't things be made so that they present a pleasing appearance, are convenient to assemble, wire and modify? Of course they can.

modern rack-and-panel construction

Let's turn back the hands of the clock a·bit —back to the rack-and-panel days of ham radio! However, we are going to do it in the modern mode so that we retain the neat, compact appearance of present-day equipment along with accessibility.

While I was developing a three-band 80-watt transmitter recently, I decided to dip into the past and produce a rack-and-panel job that would compare favorably with the current conception of how ham gear should appear. The result is shown in the accompanying photos.

Your basic unit is the mounting rack. This should be metal. Don't get a sinking feeling envisioning a lot of mitering, hack-sawing, filing, drilling and other strenuous, time-consuming and temper-shortening operations. We've come a long way since we had to be metalsmiths to produce a neat-appearing, substantial job of ham construction.

Take the basic frame shown in **fig. 1;** "Would you believe. . ." (to paraphrase Maxwell Smart) that cutting the members and assembling this framework consumed just eighteen minutes from the time the raw material was placed in the vise? That's just what it took by the clock! Cutting the aluminum extrusions was like slicing bread; another minute and a few strokes with a file removed the burrs, and the frame members were ready for assembly. With small aluminum castings made especially for use with the extrusions and supplied with springy, toothed locking devices, assembling the final rigid framework was like fastening the snaps down the back of your wife's dress—it was that simple!

Where do you get this kind of material? A good question. There may be others making these aluminum extrusions and castings, but to date I've relied solely on the Ameco Engineering Company.* Their catalog of small aluminum extrusions and castings is available on postcard request. If you're interested in this type of construction, by all means get the catalog. You'll be surprised at the wide variety of assemblies which you can make up; the catalog shows you how as well. For my project, I used 314-inch extrusion for the frame with the castings at the corners and mid-sections.

* Arneco Engineering Company, 7333 W. Ainslie Street, Chicago, Illinois 60656.

construction

The frame was made 8-3/4 inches wide by 16-inches high. For the front panel, I cut the notched ends off a standard 8-314-inch high rack panel to reduce the height to 16 inches. These panels should be available at your electronic supplier. If not, most of the mail-order electronics houses catalog them. With the exception of toggle switches, a key jack, an indicator light and a fuse holder, **nothing** was permanently attached to the front panel. The variable capacitors, crystal selector and band switches were mounted on a subpanel on the rf baseboard. This arrangement permits removal of either the power supply or the rf deck **without** unsoldering a single wire. By removing a few knobs and mounting nuts, either of the two decks can be taken out of the frame from the back.

What about the rf and power-supply subassemblies themselves? These are also a throw-back—breadboards. I used pieces of 112-inch plywood cut to fit inside the frame members as shown in the photographs. The tops were covered with thin sheet aluminum and the exposed plywood edges painted black. Two screws through the boards and tapped into the extrusions hold them in position on the horizontal side members of the frame. Before putting them in the frame, the components were wired and mounted on the breadboards—on the workbench where everything was in plain sight and readily accessible. Terminal strips are used for interconnection between decks with a small wiring harness.

After all the assembly and wiring is completed, the decks are put in position with the control shafts through the front panel, fastened in place, knobs installed, and, after connecting the wiring harness, the rig is ready to go on the air. If there are any changes, adjustments, or modifications required, you slide the rack out of the frame, put it on the bench, and go to work on it in comfort.

While the rf deck may require frequent removal as your ideas change, the power supply deck will ordinarily remain pretty much in place. If you want to completely rebuild the rf deck, simply slide it out, remove the components, throw away the breadboard and make up a new one—what could be simpler?

This was not intended to be a construction article. If the return to rack-and-panel construction appeals to you, work out your own design to accommodate the components you intend to use. Slightly different dimensions and proportions would be used to accommodate an additional deck for a modulator unit; my transmitter was designed for CW only.

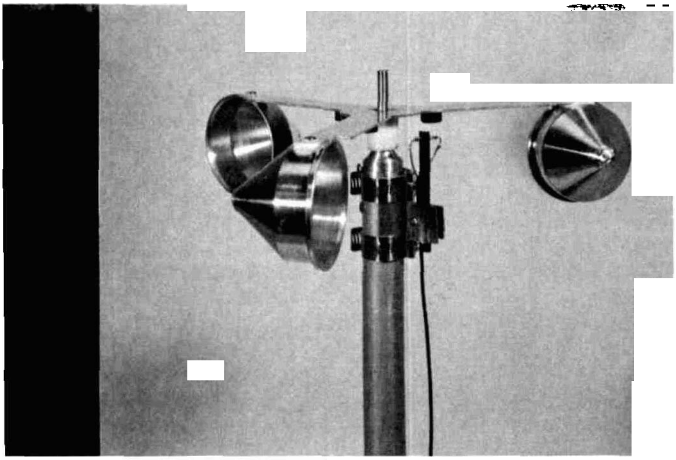
Devise as you plan; for example, the only milliammeter I had available had a long body and required more space behind the main panel than I wanted to spare from the power-supply deck. I simply cut a 4 x 4 x2-inch utility cabinet in two; this gave me an extension ring one-inch deep which moved the meter forward.

The subpanel on the rf deck is a 6-inch square of 16-gauge aluminum, mounted vertically on the front edge of the breadboard. The tuning capacitors, band switch and crystal selector are mounted to this subpanel and permanently wired into the circuits. When the rf deck is placed in position in the rack, the subpanel fits up against the back of the main panel with all shafts protruding through matching holes in the main panel.

As a final step, to suppress TVI and for safety, I enclosed the open areas of the frame in a piece of perforated sheet aluminum. It is available at local hardware and building-supply stores. It is easily cut with small tin snips and easily bent by hand to fit the frame. A few strategically placed 6-32 machine screws tapped into the frame members hold the ventilated shield in place.

Frankly, I'm rather proud of this little piece of rack-and-panel fabrication dug from the distant past. I have a nice-appearing rig, well ventilated, handy to work on and it adds a distinctly professional flavor to my shack—give you any ideas?

ham radio



The anemometer wind head. This unit features a teflon bearing and magnetic pickup for long life.

an amateur anemometer

Here's
a simple gadget
that will tell you
when to
crank your tower down

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

***Say,* beam owner,** when the winds are abroad, and you're warmly tucked in the sack, does sleep come easily? The haunting question, "Will I find my antenna farm in a big pile in the back yard?" gnaws at your brain, and sleep is elusive.

The question of just what to do about protecting your antennas against winds has as many answers as there are antenna types and weather variations. One thing is sure, however; if you are to batten-down for a wind storm, you must know that it's coming. The weather bureau furnishes storm warnings, and by heeding them you won't be caught with your beam down. However, if the antenna is hauled down every time the small craft come in, you can look forward to a busy year of battening down.

Having your own anemometer is another way of determining, for yourself and in your own exact location, whether the wind warrants the work. By keeping an eye on wind velocity, and adopting some criterion of unsafe levels, you're all set.

You can buy a simple, remote-reading anemometer for as little as $25.00, but the types that are most commonly found in ham shacks cost about $100.00. The anemometers I have had experience with have worked quite well, but have had mechanically weak points about them. Because I'm electronically biased anyway and don't like repairing anemometer wind-heads "in situ," I looked for a simpler type of anemometer.

The unit described here has no sliding electrical contacts or rotary pipe joints; its most complicated mechanical part is the bearing. This mechanical simplicity is complimented by electronic complexity, at least in concept. However, relatively complex electronic systems can now be built easily and inexpensively with modern semiconductors, as this anemometer circuit will demonstrate.

I decided upon a three-cup anemometer configuration; each of the spokes has a small permanent magnet mounted on it. As the magnets rotate past a magnetic reed switch, the switch closes and the associated circuitry puts out three pulses of constant level per revolution. By filtering out the dc value of this pulse generator output (integrating), a voltage is obtained which is proportional to the number of pulses per second. Obviously, this voltage is also proportional to the number of revolutions per second of the three-cup-rotor, since the rotor caused the three pulses per revolution to be generated.

construction

The wind-head is constructed from an aluminum spider 10-1/2 inches in diameter as shown in the photograph; three modified aluminum household funnels are mounted on the ends of the spikes. The funnels, 2-1/2-inch spun aluminum, 39c each at a local variety store, are mounted with 6-32 screws and nuts.

Internal layout of the remote windspeed indicator.

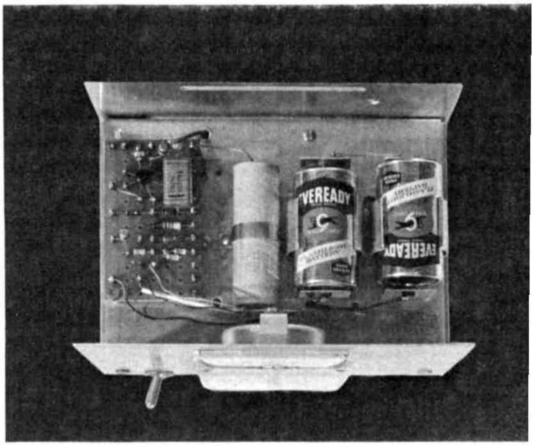
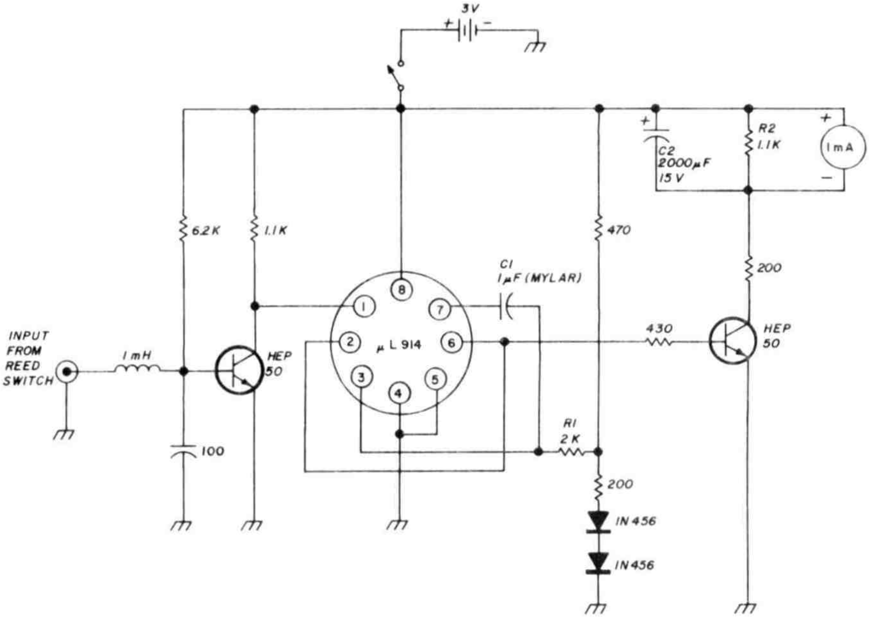
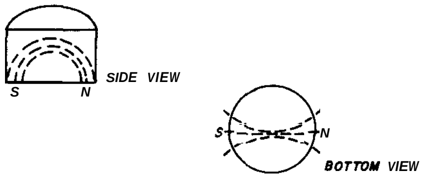


fig. 1. Schematic of the remote windspeed indicator.



The funnels originally had handles; the hole left after removing the handle was used for mounting to the spider tip. A dab of General Electric RTV (Silastic), widely sold for bath-

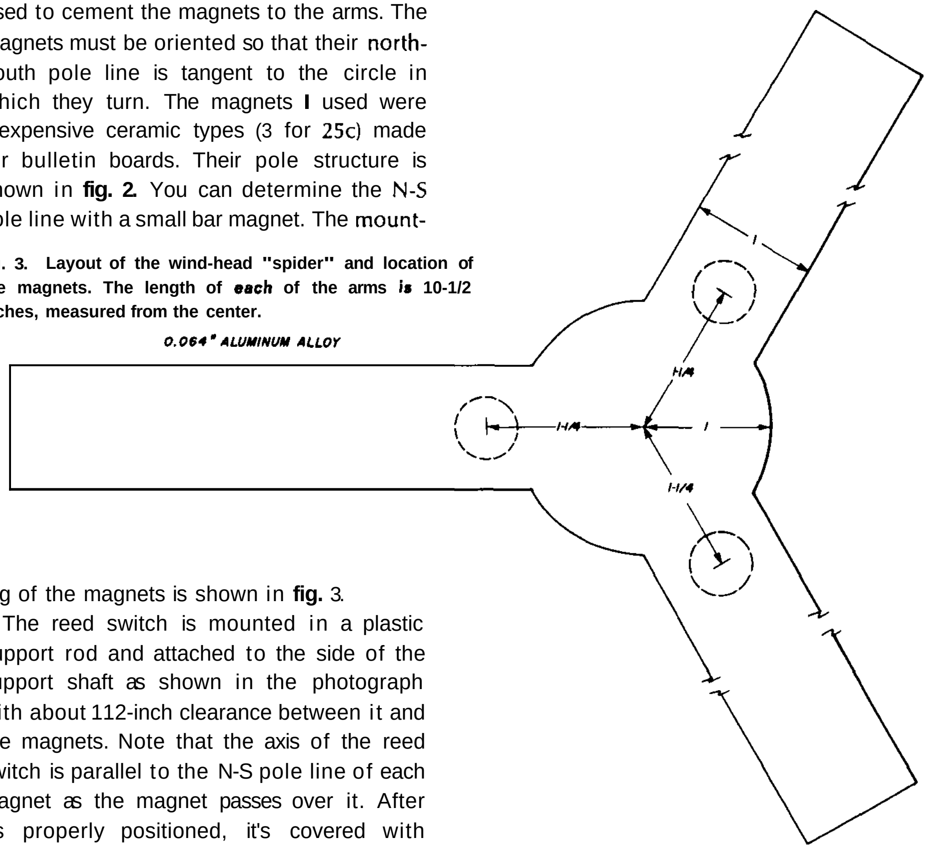
fig. 2. **Pole structure of the miniature bulletin-board magnets used in the anemometer.**



tub and shower caulking, was used to secure the funnel-spider union. The stem of each funnel was also removed and some of the excess metal bent in to close the cone; RTV was used to finish the sealing job.

Three magnets are attached to the spider arms, 1-1/4 inches from the center. RTV is used to cement the magnets to the arms. The magnets must be oriented so that their north-south pole line is tangent to the circle in which they turn. The magnets I used were inexpensive ceramic types (3 for 25c) made for bulletin boards. Their pole structure is shown in **fig. 2.** You can determine the N-S pole line with a small bar magnet. The mounting of the magnets is shown in **fig.** 3.

fig. 3. Layout of the wind-head "spider" and location of the magnets. The length of each of the arms is 10-1/2 inches, measured from the center.



The reed switch is mounted in a plastic support rod and attached to the side of the support shaft as shown in the photograph with about 112-inch clearance between it and the magnets. Note that the axis of the reed switch is parallel to the N-S pole line of each magnet as the magnet passes over it. After it's properly positioned, it's covered with bathtub caulk.

The main bearing was made of Teflon because it is self-lubricating and should stand up in the weather. I felt that a brass sleeve or ball bearing would lose its lubricant in a driving rain. Probably nylon would do as well as Teflon here, but I didn't try it.

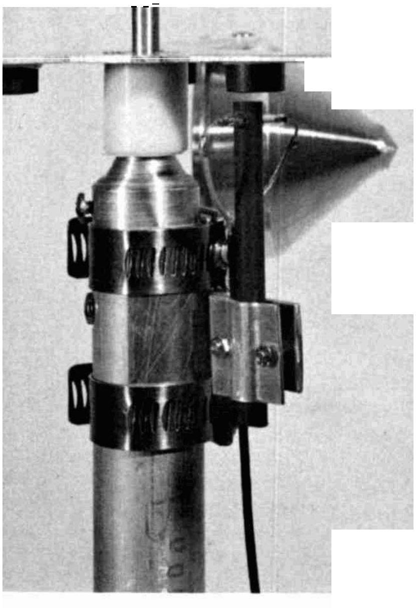
The reed switch is connected to the electronics package through a suitable length of coax—RG-58/U or RG-174/U. The shielded cable was used to keep stray local rf out of the electronics.

The circuit of the indicator is shown in **fig. 1.** It was built in an LMB-WO1A cabinet with self-contained batteries. The pulse length is set by the resistor R, producing a pulse length of about two milliseconds.

calibration

The anemometer sytem may be calibrated fairly accurately by mounting it on your car and making various runs up and down a straight road on a still day. Early Sunday morning seems best for these tests in my area. For best accuracy, your speedometer

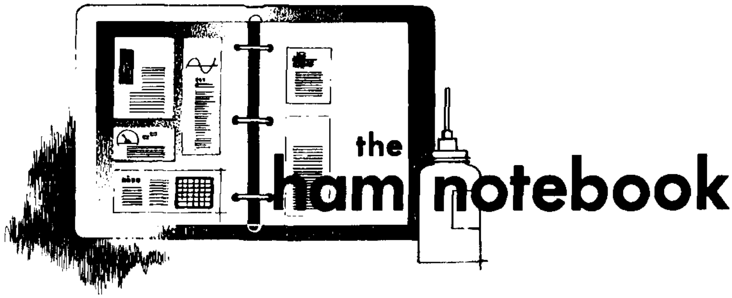
Closeup of the wind head **showing the reed** switch support and **teflon** bearing.



should be checked at the AAA or similar facility and the wind head mounted on a 2 x 4 that puts it out **ahead** of the automobile.

ham radio

"**Now** do you **understand what a squelch** circuit is, dear?"



converting wavelength to inches

It is always exasperating when you have to convert free-space wavelength from meters to inches when you're building an antenna or making a balun. With the help of a simple jingle and a little long division, three-place accuracy is easy:

"Eleven point eight-o-three inches per k-M-z."

Here's how it works for 432 MHz (3/4 meters):

$$\frac{11.803 \text{ in/kMz}}{.432 \text{ kMz}} = 27.321 \text{ inch}$$

For coax it's:

$$\frac{11.803 \text{ (in/kMz)}}{\text{Velocity constant x frequency (kMz)}} = \text{coax (inches)}$$

WA6SXC

Motorola MPS transistors

Plastic transistors carrying MPS numbers below MPS6500 are made by Motorola and are similar to 2N transistors carrying the same number. MPS stands for Motorola Plastic Silicon, and numbers over 6500 are special transistor types.

W2DXH

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

s-meter readings

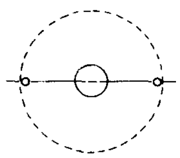
If you have ever wondered what the S-meter readings actually indicate, the following chart may be of some help. This chart is based on an input signal of 50 microvolts at S-9 and 6-dB steps between each S-unit.

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.

Jim Fisk, W1DTY

punching aluminum panels



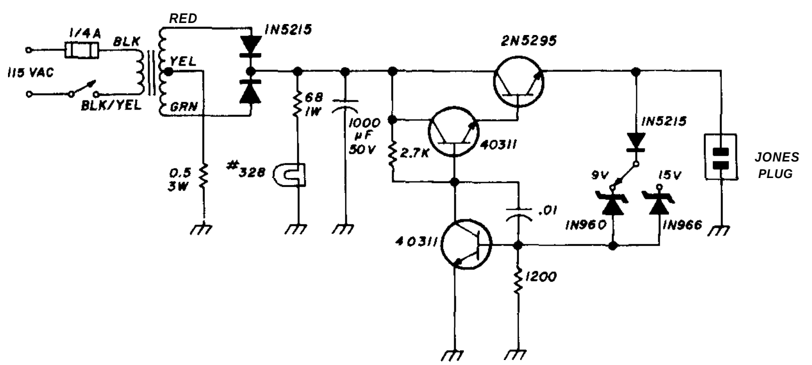
After breaking my share of Greenlee chassis punches, I have found a way of using them on 1/8" aluminum panels. First, drill your starting hole; insert the drive screw; screw on the cutter; and rotate the cutter by hand to scribe a circle. Then scribe a line through the circle; drill two small holes just **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

soldering fluxes

There may be some confusion regarding the use of "soldering paste" mentioned by VE3GFN in his article ort page 74 of the March 1968 issue of **ham radio.** Although Mike used the term "soldering paste" once modified with the term acid-free, I'm sure he doesn't refer to No-Korode and similar fluxes. 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 hydroscopic. 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



WB2EGZ's low-voltage power supply. The transformer is **a Triad 91X** or equivalent.

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.

Any small chassis may be used, but my unit was mounted in a 4 x 2-1/4 x 2-114-inch Minibox. The power transistor is heat-sunk with insulating washers and thermal compound to the minibox. Wiring is not critical. You can obtain other fixed voltages by using a 400-mW zener diode of the desired voltage.

Don Nelson, WB2EGZ

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. Cost? 30c!

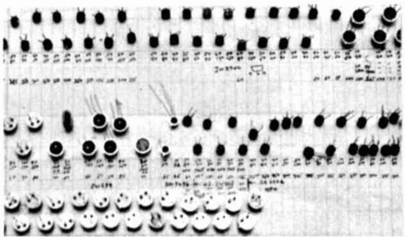
George Haymans, WA4NED

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.

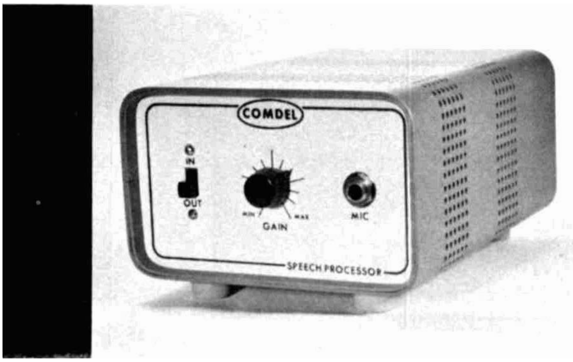
Don Farrell, W2GA

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 the photograph. 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



some considerations about speech processing in amateur equipment

J. R. Fisk, W1DTY, RFD 1, Box 138, Rindge, New Hampshire 03461

Since the inception of voice-modulated redio communications, engineers have recognized that speech processing was absolutely essential so you could get optimum use of the system without exceeding its peak-power limitations. The schemes that have been used to accomplish this restriction of peak excursions have fallen into two general categories: agc or alc control, or clipping. The agc system uses feedback to control the gain of one or more amplifier stagec, thereby keeping them within their linear operating range.

Rather complicated agc systems are used by every broadcast station in the United States to maintain the average modulation level within presrrihed limits. In addition, most modern amplifiers which are designed for amateur use incorporate a feedback system to prevent overmodulation, and keep the modulation level up.

Unfortunately, the "clipping" system of cpeech processing has earned a poor reputation because of misuse and a general misunderstanding of its advantages and limitations. The important differences between the two approaches is time. The agc system uses feedback with its associated time constants; the clipping method has no time constants other than the attack time of the device that does the actual clipping.

Each system has definite advantages and disadvantages. In general, with a properly adjusted and operating agc-alc system, you can never overmodulate your transmitter. In addition, the modulation level will he maintained within the range of the agc circuit. Essentially then, the agc circuit generally keeps the system operating within its linear range and acts as an emergency brake in case of overmodulation.

audio clipping

Clipping systems have advantages and disadvantages not inherent in the agc system. Due to the absence of time constants or long attack times, the amplitude clipper can effectively change the peak-to-average power content in the speech waveform. This is the primary benefit we're all trying to achieve. A linear system modulated with a speech waveform whose average power content has been improved will show the same improvement in average output power. However, there are definite disadvantages to the simple clipper technique.

Clipped audio frequencies generate a host of harmonic products which fall within the audio band. In normal a-m transmission, 4 to 6 dB of straight audio clipping is generally considered the most you can use before the disadvantages of high distortion outweigh the improvement in averaged radiated power. Under poor operating conditions, however, this can be a very acceptable technique.

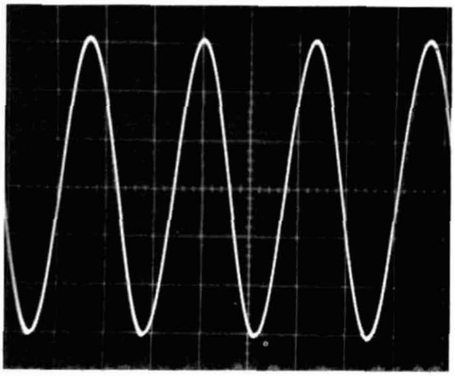
Another disadvantage of audio clipping is that audio processed in this way should not be used to modulate a ssb transmitter. Neither phasing nor filter-type exciters provide the phase linearity required to prevent the audio harmonics from recombining with the fundamental signals. This results in amplitudes that are higher than those precent before clipping. A good ssb transmitter with properly operating alc will decrease its average power output in order to accommodate the abnormal peaks generated in the exciter by harmonic-rich audio. The system not only suffers from increased distortion due to clipping, it does not provide any benefits of an improved average-to-peak modulation ratio.

Elaborate filter techniques and clipping at rf (which eases the filter problem) have been used to improve this situation, hut these systems are clipping three signals at once: the lower sideband, the upper sideband and the carrier. These signals are spaced only by the value of the audio frequency at any given instant; hence their intermodulation products still generate unwanted signals within the up-converted audio band.

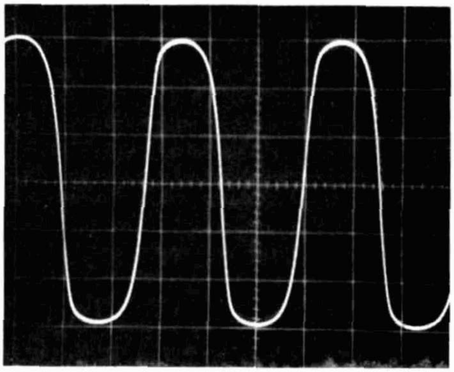
With these considerations in mind the next step should be obvious; a speech-processing technique that provides the benefits of instantaneous clipping without the disadvantages of harmonic distortion.

rf clipping

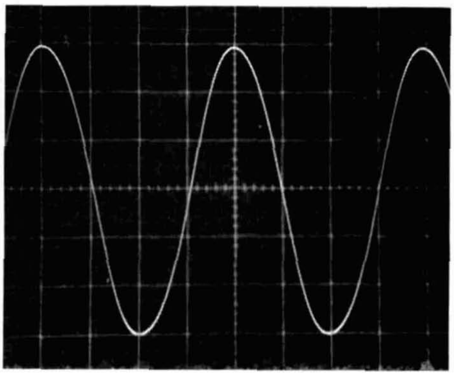
One answer is rf clipping. This is essentially nothing more than clipping the rf or i-f ssh signal, not the audio. If the single sideband signal is clipped, harmonic products are generated as with audio, hut the closest



Pure sine wave at 700 Hz.



Output from a commercial speech clipper with 15 dB of clipping. Note the flattening of the audio peaks.



Output from the Comdel CSP-11 with 15-dB clipping. Note that the output of this speech processor closely resembles the input waveform.

fig. 1. Input and output audio waveforms using a commercial speech clipper and the Comdel CSP-11.

one is removed by at least the value of the i-f frequency. With this approach, the harmonics fall outside the audio band and are easily removed by filtering. The presence of intermodulation products tends to broaden the bandwidth of the signal, so an additional ssb filter is required.

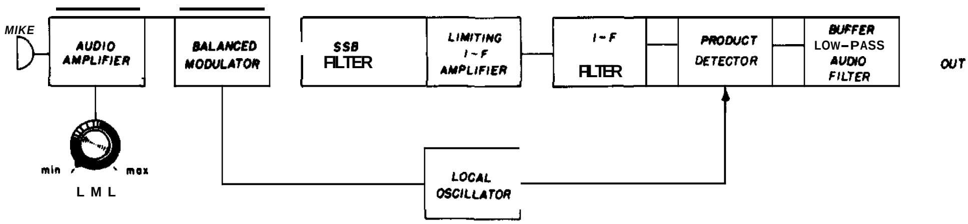
The output signal contains all the original audio information and has an improved average-to-peak power content as much as 15 dB (30 times). This obviously changes the quality of the transmission, but very little distortion is noticed. There is no loss of intelligibility "in the clear" as with a straight audio-clipping circuit, so the rf-clipping circuit may be left in the system permanently.

Modifying existing equipment to incorporate rf clipping is a little bit tricky. For one thing, the additional filter has to track the existing ssb filter very closely, and in practice, this is difficult to achieve. Also, additional gain has to be provided in the right place, since just turning up the audio gain control will cause overloading somewhere—usually in the first balanced modulator.

The Comdel CSP-11, illustrated in block form in **fig. 1,** is effectively a closed-circuit ssb system incorporating i-f clipping and avoids these difficulties. You can see that there is no frequency error because the same oscillator is used for both generation and demodulation of the ssb signal. The audio output of the system is instantaneously peak limited, yet free from harmonic distortion. For best operation, 75- to *20-dB* clipping is recommended. The apparent power gain, as confirmed in numerous tests, is often as high as 10 dB at the receiving end.

The circuitry required to accomplish this type of speech processing is quite complex. But power ratio, lack of harmonic content, a fixed known peak audio-input level, and a specifically tailored 3-kHz audio response are well worth the effort.

fig. 2. Block diagram of the Comdel CSP-11 speech processor.



Comdel CSP-11 Specifications

Frequency response:	**500-2500 Hz**
Signal-to-noise ratio:	**36 dB minimum**
Input impedance:	**0.5 megohms**
Output load:	**not less than 6000 ohms**
Input level at limiting point:	**10 mV peak**
Output level at limiting point:	**40 mV peak**
Power requirements:	**9 volts at 18 mA**
Battery life:	**300-500 hours**
Size and weight:	**5-1/2 x 3-1/4 x 7-1/2 inches; 32 ounces**
Price:	**$120.00 postpaid**

* **Comdel, Inc, Beverly Airport, Beverly, Massachusetts 01915**

references

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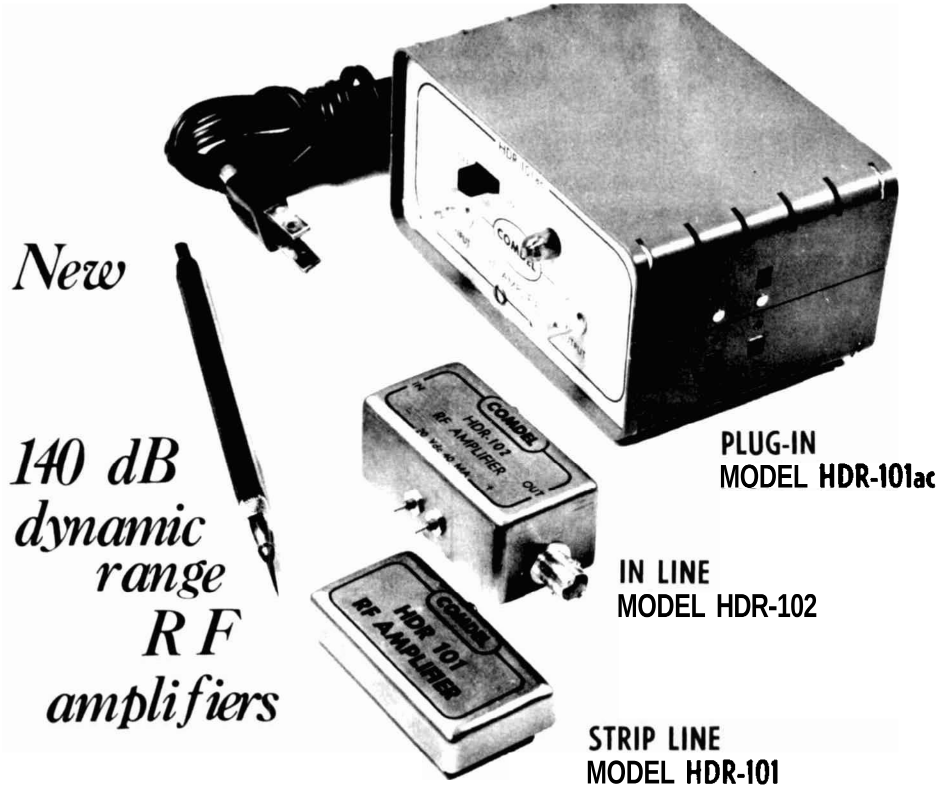
ham radio



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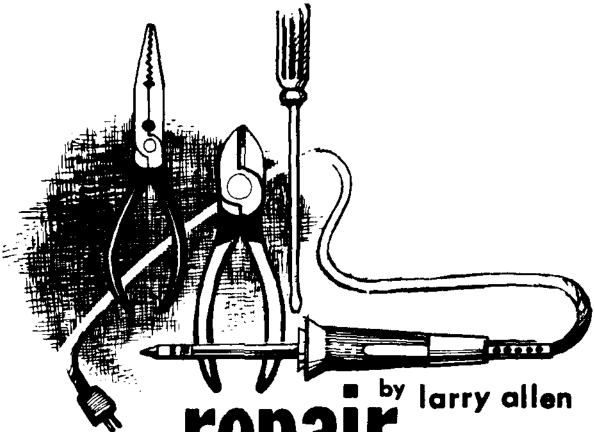


by larry allen

the repair bench

ham receiver alignment

Just about a month ago, a young college student from down the block telephoned to ask if I could help him. It seems Jeff had built an inexpensive shortwave receiver kit and couldn't get much out of it. Oh, he got some noise, but it was a far cry from the listening he was planning to do.

He brought it over, and we put it on my bench and turned it on. It was a general-coverage receiver, and we could get a few stations in the broadcast band, but very little anywhere else. The top band, which went on up to 30 MHz, was completely dead, even with a beam I have.

I ran through it real quick with the signal tracer and found the i-f gain poor. Usually these kits come with i-f cans prealigned, so I asked him if he had messed with the alignment screws. He had. The instruction book gave some adjustment instructions, and he had tried to go through them. There wasn't a screw in the i-f or rf section he hadn't turned. I could see I had my work cut out for me.

My young friend, who aspires to become a novice, had committed a sin that is often committed by adults who should know better. He had attempted to align a sensitive, selective receiver without either knowing how or having the equipment to do it with.

In the first place, you need a small screwdriver and an alignment tool that will fit the i-f slug, an rf/af signal generator, similar to the one I talked about in this column last month, and a meter. For good alignment, that generator needs to be fairly accurate. Even better, a heterodyne frequency meter can be used as a source of highly accurate signals. A lot of hams have the old surplus BC-221—I still have mine sitting around. A meter like the Lampkin MFM freq meter is okay for this, too. Anyway, make sure your signal source is accurate. If you have to, check its dial calibration at each of the several WWV or WWVH frequencies, using any receiver that is working. Later, I'll tell you how.

The signal generator output cable is usually terminated in its characteristic output impedance. If not, you can add a resistor of that value across the output leads. In case you don't know the output impedance of your generator, use a 50-ohm resistor. Use a .001-μF blocking capacitor, in series with the output lead between the resistor and the set, to make sure no dc reaches the resistor. The sketch shows how.

For an indicating meter, I prefer a vtvm, which I connect to the avc line of the receiver. A few hams and service technicians prefer to use a VOM, set to its "output" function, which puts a capacitor in series with the test lead. They use this as an indicator by hooking it across the speaker or headphone output to monitor the audio developed. The effects of alignment tuning are much broader visualized on this kind of indicator, but the system works.

Alignment has two purposes. One is to adjust all the rf and i-f tuned circuits so that they pass the correct frequencies as efficiently as possible. In elaborate ham receivers, alignment may even involve some traps to make sure that certain frequencies—such as an upper or lower sideband—are rejected; trap circuits also help steepen the skirts of a band-pass response curve.

The second purpose of alignment is to make sure all the oscillators are operating at the correct frequency, to assure correct calibration of the receiver's dial. Dial calibration is pretty important for both shortwave listening and ham reception. In a simple receiver like the one young Jeff brought by, this part of the alignment can become quite difficult. However, there are ways to overcome the problems, and I'll mention them later.

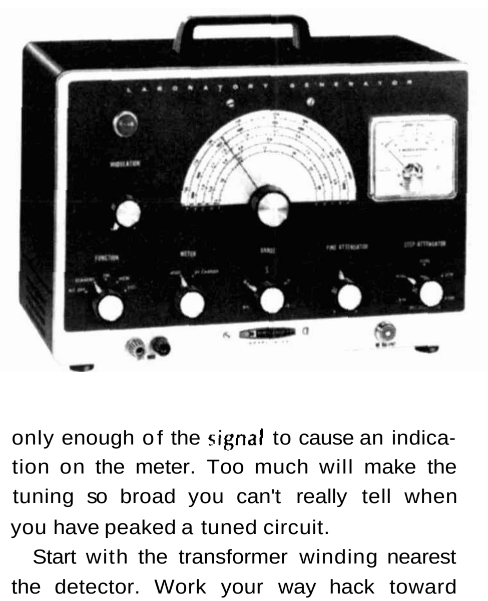
lining up the back section

Before you start aligning any shortwave receiver, let it and your test instruments warm up at least a half-hour to stabilize—an hour is better. Stability is important.

The first step is to "calibrate" the detector and the final i-f transformer. The accuracy of the signal used for this step is critical; the effectiveness of succeeding steps depends on this adjustment being made correctly. If the receiver is a multiple-function rig, set the switch for a-m reception. The whole set will be easier to align in this mode. You can clip the vtvm dc—probe to the output of the a-m detector—common lead to chassis, of course.

Connect the signal source at the input (grid or base) of the very first stage of the last or low i-f. Set its frequency very precisely. If the receiver uses a nontunable selectivity filter in conjunction with the low i-f, getting the right frequency will be simplified if you feed the signal in before the filter. Feed in

Popular kit-type signal generator suitable for receiver alignment.



only enough of the signal to cause an indication on the meter. Too much will make the tuning so broad you can't really tell when you have peaked a tuned circuit.

Start with the transformer winding nearest the detector. Work your way hack toward the input of the i-f section. When you get back to the input transformer of the section, peak it as best you can. It may be a little broad, because of the connection from the signal source. Later, when you have the signal source connected further toward the front of the receiver, be sure to come back and repeat this one adjustment; you'll usually find the true peak slightly different.

Go through the i-f strip twice. You want to be very sure it's aligned correctly. If this is the last i-f of a double- or triple-conversion receiver, the frequency may be as low as 50 or 60 kHz; on the other hand, it could be as high as 2 or 3 MHz. In Jeff's receiver, it was 455 kHz. Transformers for that frequency are inexpensive and plentiful.

the middle i-f

This part of the receiver differs considerably from model to model. If the receiver is double-conversion, it may be called the high i-f, being the only one other than the low i-f. It may be a broad-banded stage, if the tuning is done **after** it. (Examples are the Heathkit SB301 and the Collins 75S-1. Both use a crystal-controlled heterodyne oscillator for first conversion and a tunable **linear master oscillator** (LMO) for the second.) In expensive receivers, a bandwidth filter may be used for this i-f instead of tuned circuits; it is not adjustable.

Generally speaking, the **best** place to feed in the test signal is at the grid of the mixer preceding the i-f strip you're about to align. This islolates the output cable of the signal generator, and eliminates the need to repeak the first input slug.

the low-cost front end

One problem with an inexpensive receiver is poor image rejection. An image, as you may know, is a signal on the other side of the local oscillator frequency from the desired signal, exactly as far from the local oscillator as the desired signal is. In other words, it is a signal that, when mixed with the local oscillator, also produces the i-f of the receiver. The image response is always separated from the desired signal on the dial by exactly **twice** the i-f frequency. With a 455-kHz i-f, for example, the image is 910 kHz away from the desired signal. When you are receiving rf frequencies in the neighborhood of 20 and 25 MHz, an image is comparatively close, percentagewise —less than 1 MHz away. It is not easy for the preselector tuned circuit to block out the unwanted image. In fact, unless the *Q* of the preselector circuit is unusually high, good image rejection in a low-cost receiver is virtually impossible.

This means, then, that the oscillator in such a receiver is hard to get calibrated correctly, because you can't tell if you're picking up the signal at the image or at the fundamental on the receiver dial. If you work very carefully, however, you can surmount this obstacle and do a really good alignment job.

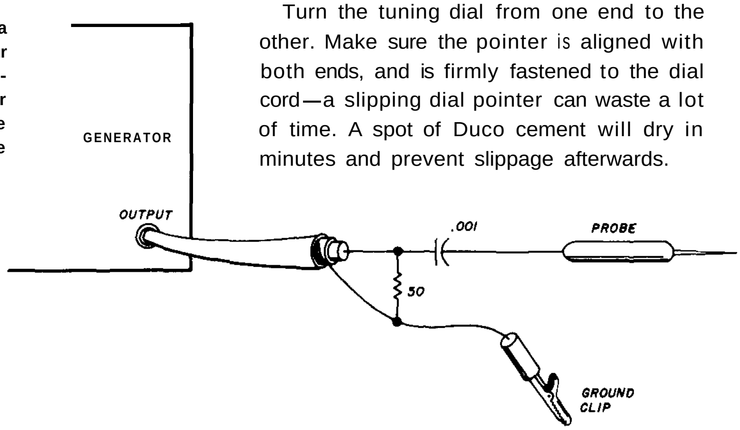
Start with a long warmup — both generator and receiver. You want as much stability as is possible. The vtvm is the best indicator for this step; the VOM across the speaker isn't really sensitive enough. You are going to depend on the meter to help you separate fundamental reception from image reception. As a general rule, use only enough signal — fed to the antenna terminal of the receiver — to cause an indication on the meter. Any more will distort the avc action of the set, and make accurate alignment difficult.

Make it a rule always to start with the highest band. In some receivers, adjusting the high bands affects low-band adjustments. In any case, it's a good idea to start with the high band — get the tough part done first.

There are three steps to each phase of front-end alignment, once you have all the i-f's precisely adjusted. The first step is calibrating the high-end dial frequency, with the oscillator trimmer capacitor. The second is adjusting the low-end dial frequency, usually by adjusting a slug in the oscillator coil. There is interaction between these two adjustments; you have to go back and forth several times. It is very important that you make sure you do not align either end with an **image** signal. The third step is alignment of the *rf* tuned circuits. There may be more than one rf adjustment. The instruction manual that comes with your receiver will tell.

In preparation for the actual adjustments, make yourself a chart of their locations. Figure them all out **before** you start sticking a screwdriver into them. During the alignment your attention should be on the frequencies and on peaking the adjustments; you have no time to be figuring out where each one is. Label them with a pencil if you can. Do anything to avoid touching the wrong trimmer, because you might then have to begin all over.

Turn the tuning dial from one end to the other. Make sure the pointer is aligned with both ends, and is firmly fastened to the dial cord — a slipping dial pointer can waste a lot of time. A spot of Duco cement will dry in minutes and prevent slippage afterwards.



This is how you connect a probe to the output of your signal generator. The resistor terminates the generator cable in a 50-ohm load; the capacitor isolates it from the dc supply in the receiver.

Now for the first adjustment: the high-end oscillator adjustment for the top-most band. Set the rf generator, modulated, to 30 MHz. Your first job is to identify the image so not to confuse it with the main signal. Tune back and forth in the neighborhood of the high end of the band, and you will hear signals at two points. Tune in one with the receiver dial, very cautiously and carefully, tuning for a maximum meter indication. Then do the same for the other. Notice which dial position produces the higher meter reading. That one is the fundamental.

See where the dial pointer is. If it is not indicating 30 MHz exactly, you'll have to readjust the oscillator trimmer capacitor for that band. Be sure you have the right one. If the dial is off very far, there is only one way to bring the dial reading into proper position without confusion: Turn the dial just the tiniest bit in the direction to correct its reading, but not so far you lose the sound or the meter indication. Then adjust the trimmer to make the reading strong again. Next, move the dial a little closer to the correct setting, until you almost lose signal again. Then again peak the trimmer. In this manner, you can slowly "walk" the trimmer and dial right onto the correct frequency setting without losing the proper signal and inadvertently zeroing in on the image frequency.

Next, you have the same job to do at the low end of the same band. This might be around 15 kHz, for example. Set the generator to precisely that frequency. Tune the receiver around the low end until you find the fundamental frequency; again, use your meter indications to make sure you don't have the image tuned in. Then, adjusting the slug in the oscillator coil for that band, slowly "walk" the signal onto the correct 15-MHz spot on the dial.

When you tune back to the high end of the dial, you'll find the frequency has shifted slightly, especially if you had to do much adjusting at the low end. Again, tune in the fundamental 30-MHz generator frequency, being sure it's not the image, and "walk" the trimmer setting until the dial is accurate. Repeat at the low end and repeat at the high end. Final adjustment should be at the high end, because the oscillator slug has less effect on the over-all frequency.

Before you move to the oscillator adjustment for the next band, make the rf adjustments for this one. There may be a high-end and a low-end adjustment in the rf section, in the more elaborate receivers. Otherwise, there will only be a high-end trimmer or two to adjust. Again, be sure the signal you're feeding in and tuned to is the fundamental one. You don't have to "walk" the adjustments in the rf section. Just tune the dial for maximum indication on the meter. Then tune the trimmers—and slugs, if there are any for low-end adjustment—for maximum.

Next, go to the band next below in frequency. Go through the same procedure: First, the high-end oscillator adjustment, "walking" the dial setting into proper position. Then the same for the low-end—and then back and forth between the two until both ends are accurate. Finally, peak the rf adjustments. You'll find that separating the fundamental from the image becomes much easier as you get down in the lower bands, because the 910-kHz (or whatever it is) signal has a wider spread on the dial. In the lowest band, usually the broadcast band, you will have no trouble at all; the image is half the dial length away.

Finish by aligning each band all the way down to the lowest, repeating the procedure already outlined. When you are finished, if you really care about calibration, accuracy, and sensitivity, start with the high band and double-check the high- and low-end frequency settings.

You may want to check the **true** calibration of the receiver, to overcome any inaccuracies the signal generator may have, especially in this highest band. A good test is to tune in the National Bureau of Standards station, WWV, at 15 MHz. You'll have to do this at a time of day when the band is open. If you can, tune also at 20 MHz and 25 MHz. You may not be in the right locality for best reception at those frequencies, but you can check them whenever propagation permits. See if the receiver dial is correct. If you find the dial is not accurate, make the correction only with the oscillator **trimmer.**

Make the same WWV test in the lower bands, using the 10-MHz and 5-MHz WWV signals. Remember to make corrections only with the oscillator trimmer; do **not** change the oscillator slug. Some night, you can also check at 2.5 MHz, using WWV or WWVH.

The broadcast band is easy to check out, because broadcast stations must be very close to correct frequency. Tune one at the high end and correct the dial calibration with the oscillator trimmer. Pick one at the low end and correct it with the oscillator slug. As you did with the generator, go back and forth to get each precise.

verifying generator accuracy

Probably the handiest way to do this is with a freq meter. The simplest way is just to feed the outputs of the generator and the freq meter into a receiver. It doesn't really matter whether the receiver is in alignment or not, since all you are actually doing is using the. receiver as a detector to detect the beat between the two.

Set the frequency meter precisely for whatever frequency you want to check—say, 20 MHz. Set the generator there, too. Tune the receiver until you can hear the signals beating together. Don't use an antenna. Swing the generator dial until you hear the whistle caused by the beat between the generator and the frequency meter signal. Swing it slowly back and forth until you find the exact zero beat between the two. See whether the generator dial actually rests at 20 MHz. You can plot any inaccuracies at intervals of about 5 MHz, and use the graph as a correction factor when you're calibrating a receiver dial.

Another way to check accuracy of the generator is with the receiver and WWV (or WWVH). just tune in at the 2.5-, 5-, 10-, 15-, 20-, and 25-MHz signals. Then check to see if those corresponding signals from the generator are accurate on the generator dial. A zero beat is your guide to precise tuning.

ham radio

simple untuned crystal mount

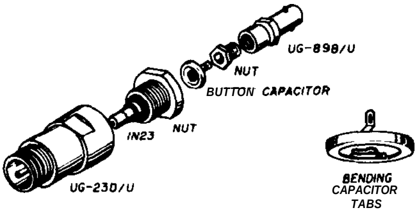
Many times in amateur UHF and microwave work you need an untuned crystal mount for detection of small levels of rf energy. Usually a small diode and bypass capacitor are simply haywired into the circuit, but this method usually causes some problems. First of all, if the frequency of operation is high 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.

This crystal mount 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 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 type-N 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 feed-through capacitor. Make sure that the N nut has been reamed out sufficiently so the BNC nut sits flush and level. The easiest way to solder the parts together is on an electric stove or hot plate.

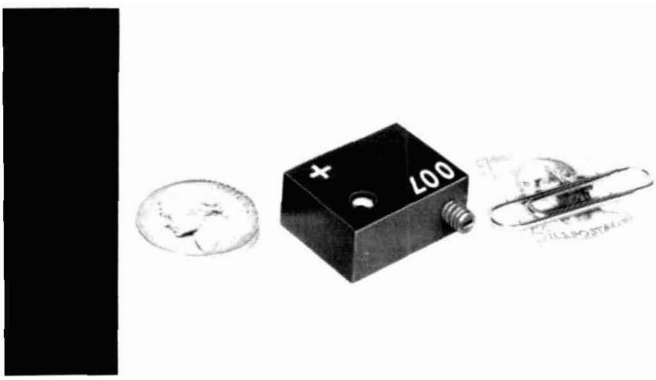
After this assembly has cooled down, solder a short piece of wire about 5/8 of an inch long to the upright tab on the button capacitor. Place the female pin from the BNC connector



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



One of the more highly publicized bugs is this "007" sugar-cube mike. It is completely self-contained and will transmit up to several hundred feet.

electronic bugging and the ham

The true story behind the 'electronic bugging" headlines

Robert M. Brown, K2ZSQ/W9HBF, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

Surprising though it may seem, very few surveillance devices are more complex than those shown in this article. For the most part, these simple devices reflect the vast majority of gadgets employed today. Exceptions are custom-built designs used by the CIA and a few other federal branches who have the funds available to develop more advanced (and generally higher-powered) bugs for specific applications. If you can ignore the mass-media spy syndrome ("Get Smart," "U.N.C.L.E.," etc.), you are well on your way.

Basically, the bugs break down into two major types: the audio-microphone group using connecting wires, and self-contained rf transmitters. Countless variations exist, yet these are the exception rather than the rule. Quite naturally, the microphone-types are the least expensive since they aren't nearly as complex as the transmitters and require a bit more installation time. One company that specializes in low-cost mike configurations is Consolidated Acoustics of Hoboken, New jersey. Perhaps the largest direct-mail solicitor, Consolidated offers mike kits to put together everything from wall-listening devices to parabolic-dish long-range snoopers and also offers inexpensive telephone induction systems. These normally make use of a Philmore-type pickup coil placed near the phone sidetone coil. Output is coupled into a nearby audio amplifier, recorder, or rf transmitter tuned to the fm music band. While a big outfit in terms of sales, Consolidated's offering is pretty simple in concept. Audio amplifiers, crystal mikes, cabling, and recorders comprise the bulk of equipment sold, and this is fairly illustrative of their competitors, too.

The more recent outpouring of tunable fm band transmitters, however, poses a more serious threat to the potential victim. While pretty much short in range capability (few travel more than 300 yards at 100 MHz), countless forms are taken, since the tiny one-way devices can be hidden practically anywhere. Although phone bugging is very big right now, these transmitters have turned up behind pictures, inside ashtrays, disguised as diaries, in pen desk sets, and in table lamps. While not elaborate in design circuitry, they're assuredly strong in imaginative application.

Common bugs. To the left is the model 003; a parallel-connected fm phone bug with a 1/4-mile range. In the middle is a model 001 series-connected phone bug with a 1/2-mile range powered by the telephone. To the right is a model 003A, a room transmitter similar to the 003 with a built-in microphone.

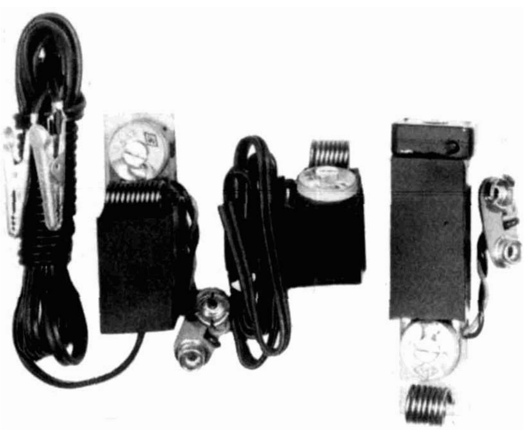


Photo courtesy L. N Schneider

Feeding from both sides (microphone-wire, and transmitting) the electronic tailing business emerges. The assortment of car bugs and bumper beepers now saturating the marketplace seems to be a combination of several techniques, with heavy emphasis on rf transmission. Interestingly, these devices are found as voice-bugs, strategically located under the dash and powered by the ignition system with an antenna coupled to the car radio whip, and as pulsed-tone beepers, emitting a constant a-m or fm signal interspersed with a 400-Hz note, antenna trailing just above the road surface. Regardless of method, the tracker assumes the identity of a ham or CB'er, using a vehicle well-equipped with a directional loop antenna, or phased whips coupled to a receiver under the dash. Complex triangulating systems are frequently employed with the aid of other vehicular operatives linked together with a CB set and zeroing in on the target. Yet, even the single agent can approximate distance and direction by looking for a null between the two phased antennas and noting S-meter readings.

While it has been said that telephone bugs have been the biggest boon to the surveillance industry, it is the tailing devices which have had the greatest application to the largest number of U.S. private investigative firms, since they cut required manpower enormously and enable safer tails (ones in which the target car does not have to be in direct sight of the operative).

Even this grouping is not exotic in terms of equipment function or design. From an amateur standpoint, in fact, the majority of devices now in use are rather behind the times. Very few employ IC's, and even fewer make use of tunnel diodes or FET's, although basic solid state long ago replaced vacuum-tube types. The exception to this is the 117L7GT a-m room bug made in Florida and designed to be permanently installed in the wall of a freshly-built apartment or home. This type is constantly on, drawing power from the household currents and transmitting on a frequency just above the tuning range of conventional a-m radios. It is good for the life of the tube, normally several years, since plate current is resistively dropped below the amateur norm. Most of the circuits employed in transmitting devices are straight from an old **Popular Electronics** or Radio-TV **Experimenter,** with certain improvements and changes made in the tanks and power supplies.

transmitting eavesdrop "plants"

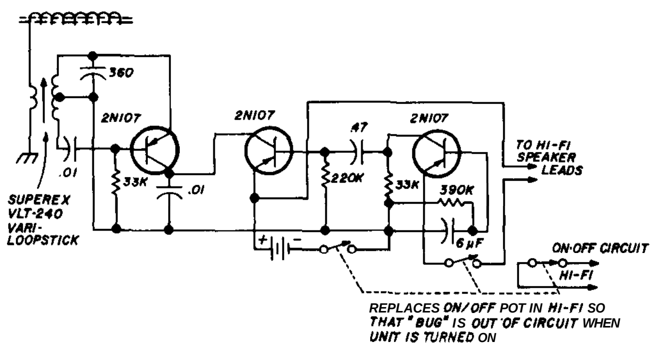
I think we can safely assume that the experienced ham already has a good deal of working knowledge on audio amplification, types of miniature microphones, etc., normally used in direct-line taps and other forms of wired bugs, so for this reason we can skip into the more interesting field of rf transmitting types. As can be seen in some of the circuits shown here, these mini-transmitters are far from complex. The vast majority transmit with outputs ranging from 50 to about 250 milliwatts and seldom contain more than three semiconductors. Yet, determinate upon frequency of operation, the range expectancy can vary significantly.

Most room bugs and a good number of wired-in phone transmitters are powered by Mallory Duracells and the like, require little maintenance, are relatively drift-free, and offer a life expectancy of two to three weeks of constant operation. Often, tiny hearing-aid variety mercury cells will be used too, although the power input to the final transistor (and resultant DX) is proportionately less. Few circuits employ more than 12 Vdc, with most in the 6- to 9-volt area.

Frequency of operation is the key to the questionable success of the modern eavesdropper. Obviously, if he uses standard a-m or fm frequencies, he runs the risk of accidental interception by the very party he's trying to overhear. A great number of private individuals though, dabble in snooping with inexpensive mail-order devices use the 88- to 108-MHz band, centering their transmitters at 89 MHz. For the most part, however, these are small-time operators, and, judging by the places the ads appear, curious youngsters. Although a professional may begin with such a device, he invariably turns later to a rig capable of transmitting either just outside one of these bands, or on a different frequency range altogether. Consequently, anyone even slightly interested in this aspect of electronic eavesdropping must be prepared for just about anything.

A group of nine operative bands are used in 99% of the transmitting bugs, breaking down pretty much like this:

fig. 1. This a-m room bug can be concealed in a hi-fi, tv or radio cabinet with the speaker as the mike and the lina cord as the antenna. Signal is tuned in between 500 and 700 kHz on a standard broadcast receiver.

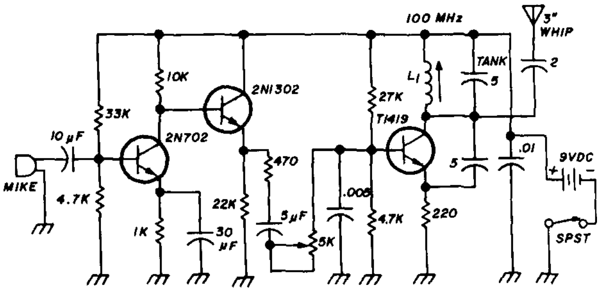


1. Just below the a-m broadcast band,
2. Just above the a-m broadcast band but below 160 meters,
3. An unused channel on the 27-MHz Citizens' Band,
4. 30 to 36 MHz fm,
5. 42 to 49 MHz fm,
6. 70 to 74 MHz fm,
7. Just below the fm music band (below 88 MHz),
8. Just above the fm music band (above 110 MHz),
9. 150 to 174 MHz fm.

By and large, the most popular range of frequencies falls in category 7, about 1.5 MHz below the bottom of the fm broadcast band. The reason for this is twofold: one, any transistor portable FM radio can be detuned to capture 86.5-MHz signals and, two, the antenna-length requirement is quite minimal (usually on the order of 3 inches and in the form of #28 wire). Additionally, fm operation at these frequencies is generally interference-free and not troubled appreciably by ignition pulses. It has been estimated that 50% of the professional (law-enforcement, detective agencies, etc.) market for such equipment falls into this band, and more than 80% of the gear sold over the counter or by direct mail to the general public.

To a somewhat lesser extent, category 8 can also be grouped with these bugs, because the same equipment is generally tunable over a 25-35 MHz range (easily hitting above 108 MHz), although it is not as frequently employed. One of the reasons for this is that commercial-aircraft frequencies have been twice "hit" unintentionally by eavesdroppers and monitored by the FAA. Subsequently, vague warning letters arrived from the FCC, and many of the surveillance houses promptly dropped all references to above-the-fm-band operation in their promotional literature. One company, Jones Equipment, of southern California, discontinued their product entirely.

fig. 2. This circuit is the backbone of the commercial spy equipment industry. With a tunable output. this set is generally pocket-sized and will transmit well over 50 feet with a whip antenna cut to the 88- to 100-MHz fm band. L1 is 3-1/2 turns number 26 enameled on a 0.3" diameter ferrite core, 1/4" long.



The result has been a recent trend away from anything above 89 MHz.

While a-m broadcasters have been around for years, and new transistor circuits are popping up at an average of one every three months in electronics publications, they are nowhere near as popular as you might expect. Except for permanent-type installations in furniture, such as hi-fi's, TV's, and the like and rugged designs built into the home during construction, the tendency is to pass this one up altogether in favor of hf and vhf. In spite of this, however, it would be foolhardy to rule this possibility out during a debugging investigation, since you can never be sure. Generally, though, the transmitter is capacitively coupled to the household wiring or to the telephone lines to achieve the required antenna length, and can be rapidly located with this in mind.

Professional "kits," as they are called by federal purchasers (such as the Bureau of Narcotics, Internal Revenue, Customs Bureau, FBI, etc.), more often than not consist of larger transmitters in the 1/2- to 6-watt category which are crystal-controlled in the 30- to 50- and 150- to 174-MHz bands. These bugs often take the form of the "left-behind" attache case, although at least one is known to have been secreted into a picture frame.

Another band popular with law-enforcement and professional security people is the 70-MHz area, although custom-built fm receivers required for monitoring are available from only three suppliers. Generally, this equipment has a greater transmit range (frequently to 15 miles) and is retrieved as insurance against ultimate detection and for re-use at a later date.

The great proportion of bumper beepers and other tailing aids also falls into this three-band frequency range, although an increasing number are cropping up on the 27-MHz CB band due to the inexpensive availability of pre-wired transmitter printed-circuit boards from such sources as Radio Shack, Round Hill Associates, Lafayette Radio, etc. These transmitters, of course, range anywhere from 100 milliwatts to 5 watts input, although the greatest majority are in the 250 mW area.

Although it has been known to happen (a 1965 case in Los Angeles), the likelihood of an eavesdropper using a ham radio band is nearly nil. The reason is the poor availability of commercial flea-power oscillators for the amateur vhf bands, at least when contrasted with the burgeoning CB marketplace. Additionally, the chances of accidental interception by a monitoring operator might trigger an unofficial investigation, particularly since eavesdropping transmitters are usually at work over extended periods of time.

Stability is another factor that varies considerably from one device to another. For the most part, L/C tanks control output frequency with very few crystal-controlled bugs. For this reason, the eavesdropper selects a spot on the receiver with ample room both sides of the carrier to allow for drift, however slight. Over extended periods of time, battery drain will cause frequency "jumps" in addition to gradual drift which must be contended with. This can work, though, to the benefit of the countermeasure people, since accidental jarring of the unearthed bug (jogging frequency at the same time) will probably be read at the other end as a normal transmitter characteristic. Too frequent occurrences like this, however, are dead giveaways. It should be noted that diode rectifying fm oscillators powered by phone-line currents seldom follow this trend, and, therefore, should be treated with care. Aside from minor frequency shifting, these configurations are relatively stable.

more exotic bugging devices

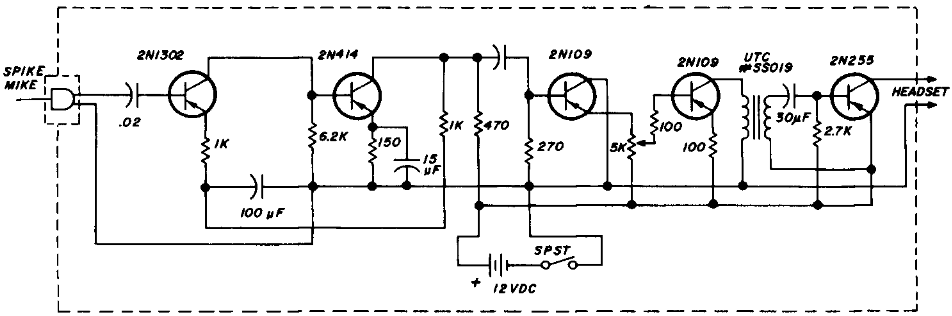
While it is true that the following devices have received more publicity than those previously described, it should also be noted that their use is nowhere near as extensive in the field. One such animal is the tunnel-diode transmitter, discussed at length in a recent issue of **Electronics World.'** This configuration has been developed by R&D shops filling in during slack periods with "custom-engineered" bugs for law-enforcement customers concerned with microminiaturization and increased range. One type produces fm modulation by using the audio signal to change the anode bias of the tunnel. This affects the resonant frequency sufficiently to create an fm signal in step with voice frequencies. Best of all, it takes only 10 tiny components in addition to the L/C circuit to do it.

Another TD application not yet beyond the drawing boards is the "free-power" transmitting bug, which uses the tunnel to rectify rf derived from a local high-power broadcast station. In operation, the tunnel follows a tuned circuit and is in turn followed by a hefty storage capacitor. In practice, however, not enough power has been generated in this fashion to power most fm transmitter circuits used by the eavesdroppers.

The advent of transistorized VOX, however, has been a boon to the private detective agencies plagued with high manpower costs and low-budget clients. NYC's Manny Mittleman has designed a device which is preset to respond to only normal room sounds such as footsteps, doorknob turning, voices, etc. In the presence of such sounds, the VOX completes the mercury cell circuits and actuates the transmitter. This item is just the ticket for motel-type plants, since it can be placed in the room days ahead of schedule, "in advance of whatever action he is interested in," according to Mittleman.

Carrying this further, some agencies make use of a "double-VOX" system that extends the concept to the monitoring post. Using a broad, constantly running receiver (in case the remote transmitter shifts a bit in frequency), the VOX mechanism is placed between the receiver output (usually squelched by preset) and a Concord-variety tape recorder

fig. 3. The much-publicized spike mike can look like this, or have the mike element and audio amplifier separated by a short cord. Since this bug works on a conductance basis, it can be driven into a door or a plaster wall. Transformer is a standard miniature output type, 500 ohms center tapped to 8 ohms; only one-half of the primary winding is used.



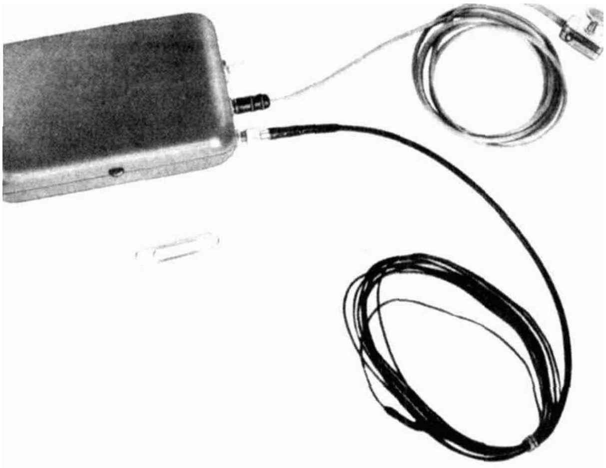


Photo courtesy C. H. Stoelting Co.

This transmitter has a range up to two miles and may be completely concealed in the eavesdropper's clothing. The cuff-link mike is in the upper right.

with long recording-time capability. When the receiver "hears" the room noise, it starts the recorder. In addition to permitting a system unattended by agents, it furnishes the client with an unedited, near-continuous tape of what's been happening.

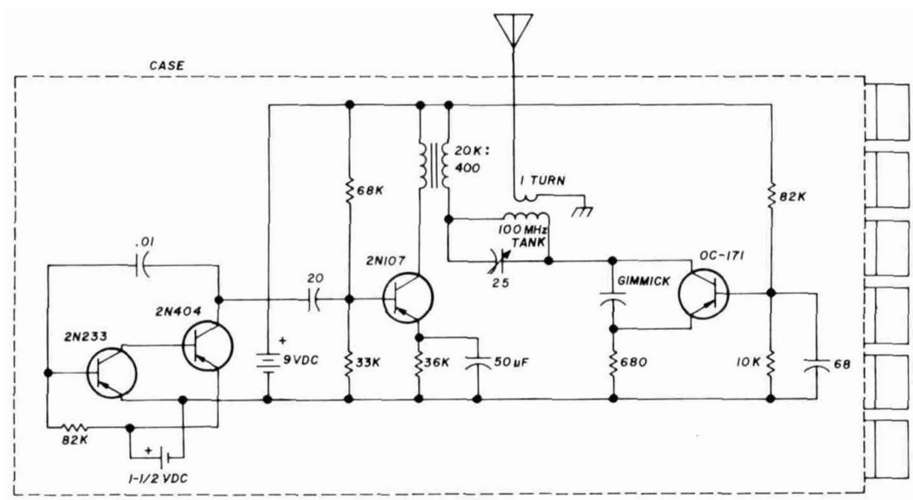
Perhaps the most widely publicized, however, has been the "harmonica bug," a unique circuit that doesn't use rf transmission at all. Designed to be planted in the telephone

table 1. This milliampere-hour chart can be extremely helpful in determining the life of an unearthed "bug". To ascertain maximum power life, divide "mA hours" by the current drain of the bug. Example: a 2-mA bug powered by a 9-volt #146 battery would have 175 hours bug life.

mercury cell life chart

battery type	mA hours	at drain of (mA)
1	1000	35
3R	2200	42
4R	3400	63
9	2400	50
12	3600	62
42R	14000	250
133	1000	35
146	350	3
163	500	10
164	500	10
165	500	10
177	160	5
233	2200	42
312	36	2
400	75	2
401	800	25
450R	350	3
5MR	2400	50
520	130	5
601R	1800	30
625R	250	5
630	350	3
640	500	10
675R	160	5

fig. 4. Here's a popular tracking device which is often used with a second operative at a different location for triangulation purposes. It's a screeching fm bumper beeper which emits a shrill audio tone near the top of the 88- to 100-MHz fm band. Antenna is a 3" whip. Transmitter is attached inside the bumper or under the gas tank with magnets. L1 is 5 turns number 14, spaced 3/4" on a 3/8" diameter form.



where room conversation is to be overheard, the eavesdropper merely dials the number of the bugged phone from any telephone in the Direct Dialing System, blows a 500-hertz harmonica note into the phone just before the bell at the other end rings, and presto! Instant eavesdrop.

Consisting of a subminiaturized Bramco-Controls-type resonant-reed decoder relay and a miniature single-stage audio amplifier, Generally installed in the base of the phone itself, this eavesdrop "ultimate" sells anywhere from $699 to $1000, depending upon supplier.

Although microminiaturization has given us the sugar-cube mike, fountain-pen transmitter, etc., it is well to remember that these tiny transmitters are limited in terms of range capability. The famed martini-olive, disclosed in a Senate Judiciary Subcommittee on Ad-

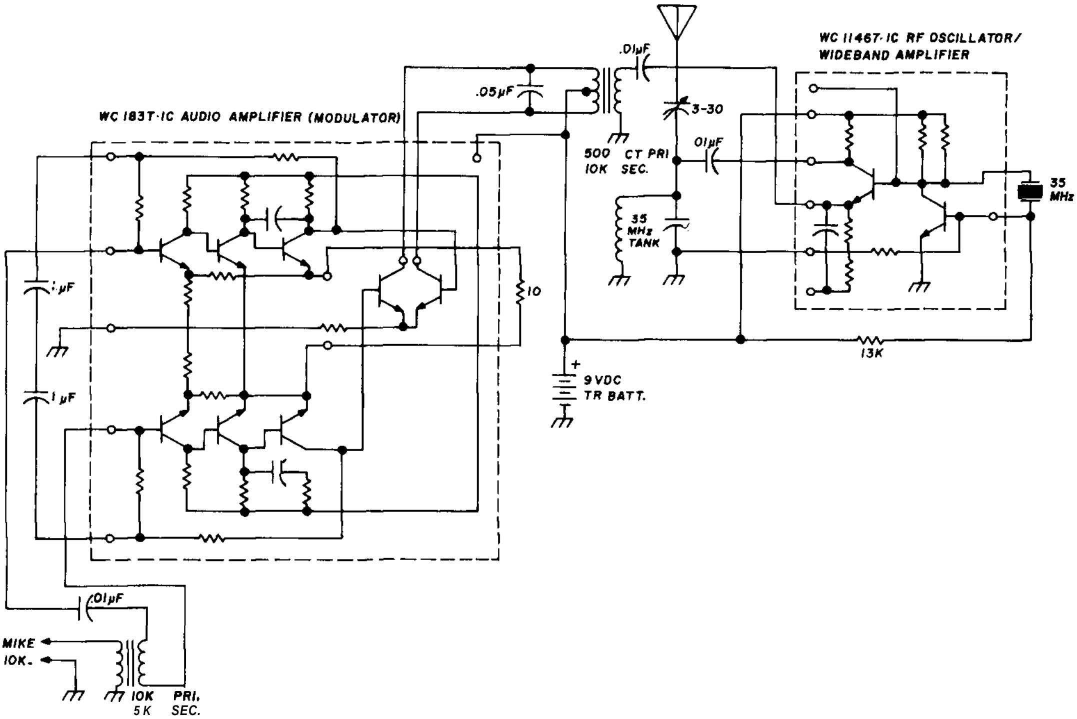


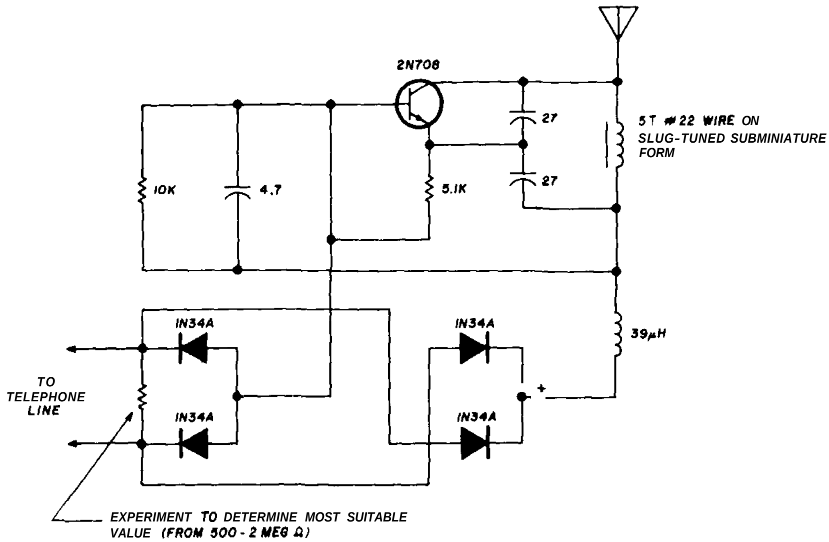
fig. 5. **Professional eavesdroppers pay upwards of $200 for this bug which uses IC's for ultra-compactness. With an output of 50 milliwatts at 35 MHz, the listener only needs a 30- to SO-MHz fm monitor receiver and a dipole.**

this sophisticated bug is permanently wired across the phone lines. When the decoder picks up the 500-hertz trigger note, the relay is actuated, simultaneously deactivating the bell-ringing circuit and transferring the mike line to the audio amplifier, which in turns feeds information back to the eavesdropper.

Should someone else dial the number during this period, he merely gets a busy signal. Should the victim desire to use the telephone himself, the eavesdropper simply hangs up just before the victim picks up the phone receiver. If timing is good, no one is the wiser.

ministrative Procedures and Practice session under the chairmanship of Senator Edward V. Long, turned out to have a range of only 20 feet. In spite of the sensation it created in the press, the bug was a custom job not suitable for actual use since submersion in a dry martini would quickly put it out of commission. Other microminiature designs are similarly referred to by professionals as "toys" geared for sale to the general public on their gimmick value. The smallest, known as the "007," measures 1 x 3/4 x 1/4 inches.

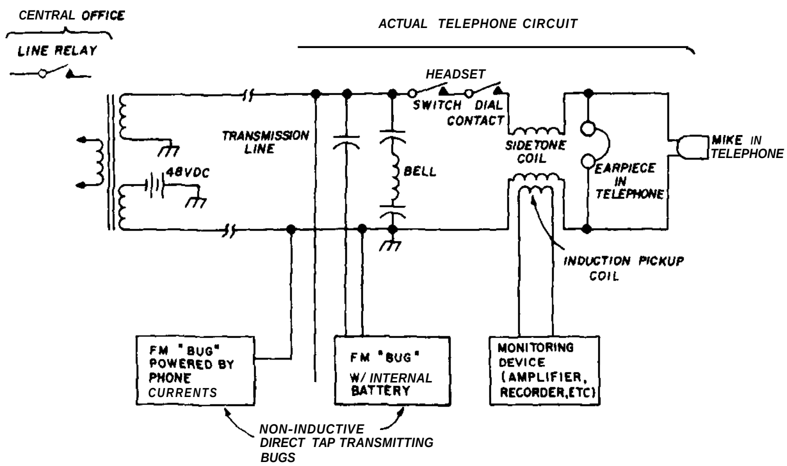
From this point on, discussions of exotic fig. 6. Known as the drop-in transmitter, this item is one of the hottest circuits around. It's built onto the back of a standard telephone carbon mouthpiece with subminiature components. The eavesdropper merely unscrews the existing carbon button and replacer it with this baby. Hot wax is poured over the circuitry to hold it in place.



bugs take on an air of the future. Long-range parabolic microphones, while effective in certain circumstances, are unwieldy and easily spotted. The Electro-Voice tubular shotgun mikes **(not** a la Les Crane) are similar monstrosities not well suited to practical snooping. Additionally, audio-filtering systems are not fully developed to the extent necessary to extract passing automobile noises, etc., from the resultant pickup.

This brings us to the much-discussed laser beam eavesdropping, perhaps the furthest of all from reality at this point. Experts in laser research feel that it will take a minimum of 20 years before any tangible results are realized, although such media as the TV spy shows and **Esquire** magazine would lead you to believe quite the opposite. While it is true that micro-reflectors consisting of optically invisible Angstrom-thick reflective paint have been developed, its use as a window reflector for laser snooping is pure conjecture right now. And as indicated earlier, audio filtering techniques are far from the ultimate,

fig. 7. Telephone tapping is popular because **it's simple.** This diagram shows three ways a bug can be hooked **up.**



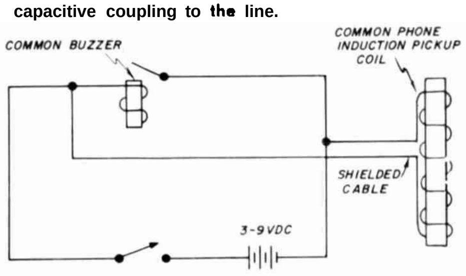
although Miles Wireless Intercom, Ltd., of NYC is said to have perfected a system (primarily for plastic "Sonaband" recording purposes) guaranteed to separate voices from typewriter clatter, cars, air conditioners and overhead airplanes.

electronic debugging equipment

In addition to having a working knowledge of what he's up against, an amateur reader concerned with detecting and/or defeating electronic bug plants should know something about the so-called professional bug locators often selling for prices many times their component cost.

By and large, the biggest selling countermeasure item is a variation on the tried-and-true field-strengh meter. Often going for prices up to $250, bandswitching FSM's offer the debugging man a means of observing the signal on a meter. Thev also permit

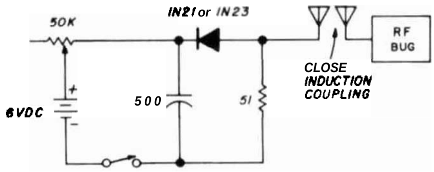
fig. 8. Cheap and dirty jamming device induces battery-powered buzzer hash into the telephone through induction, although many users prefer direct capacitive coupling to the line.



earphone monitoring to insure that the tuned-in "blip" is indeed a bug. Most FSM's are sold as "hug locators" (R.B. Clifton's $98 model is called the "Hound Dog") and contain at least one stage of amplification. Unfortunately, however, this procedure is frequently beyond the scope of the typical non-electronics type and few manufacturers, who often simply re-label Lafayette instruments, offer clear instructions for the user.

Another instrument uses a TD as a broad-band rf detector that operates as an audio oscillator when triggered into operation by the presence of a low-level rf signal. The user only has to adjust the potentiometer until the device is just on the verge of breaking into oscillation; it can also be used as a sensitivity control lo insure against triggering from strong local broadcast stations. Similar in appearance to the Radar Sentry gadgets that clip to your auto visor, these signallers also come in a beep-light configuration for mounting over doors or on desks to catch the

fig. 9. Simple vhf noise generator is an affective bug-defeater if placed close enough to an unearthed fm bug. Simply wrapping the antenna wires creates enough hash to disrupt the bug's oscillator.



attache-case types. A specialty of the Dee Company, about $250 will buy one.

The CPO/keying monitor selling in ham circles for under $15 is frequently sold to countermeasure people at prices up to 10 times that much. Essentially operating on the same triggering-concept as the TD instrument, this one uses a telescoping whip antenna for pickup and rectifies the signal with a conventional diode, filtered, and fed to an audio oscillator, amplifier, and loudspeaker.

By far the most effective device now employed in countermeasures, however, is the

With one stage of transistor amplification and a sensitivity to 600 MHz, this field-strength meter is guaranteed to "get the goods."

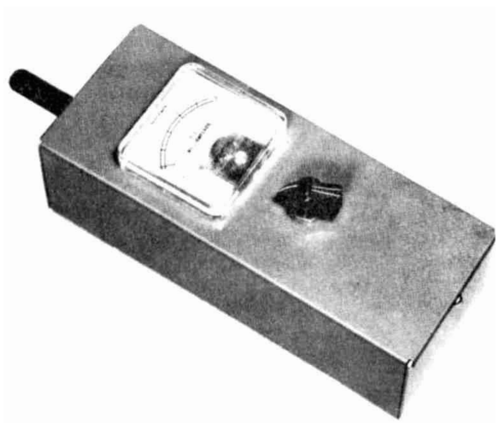


Photo courtesy R. B. Clifton

feedback detector. Although it isn't available through many commercial outlets, it appears to be a "specialty" of Security Electronics and R&S Research, hoth leaders in good lines of debugging kits. While new to the eavesdropping marketplace, the feedback detector is old hat to hams. Basically, it is a broadband (nearly untuned) super regenerative circuit that functions somewhat like a Heath Sixer or Twoer receiver section. The debugging man only has to tune slowly across the band–which may run 35 to 120 MHz–and listen for the bug's signal.

While such a device will also pick up TV, commercial fm broadcasts and the like, in the presence of a nearby transmitting bug it will screech in the true feedback tradition. Once the bug is revealed, the operator can simply switch from the loud-speaker to an S-meter position to close in on the instrument and pinpoint its location. Good feedback detectors run about $150 and come with an rf gain control, earphone jack and meter switch.

fig. 10. Block diagram of an effective **defeater/jammer.** A 2-MHz sweep generator circuit sweeps 1 MHz eech side of center. This is amplified to drive a harmonic generator, thereby sweeping through the spectrum from 1 MHz **to** beyond 400 MHz. Result is complete blanketing over so wide **a** frequency range that few bugs escape obliteration.

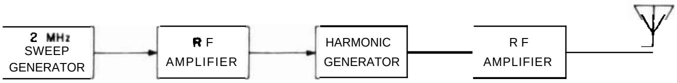
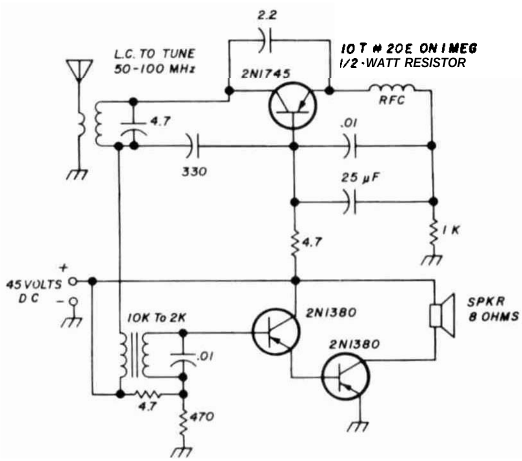


fig. 11. This portable broadband superregen is ideal for defeating hidden fm bugs since the regeneration is generally much more intense than the bug's own signal. In addition to wiping out the center **carrier** frequency, it will obliterate several **MHz** on **each** side.



It is claimed that the **Antibug Mark III** "effectively jams all commonly-used electronic eavesdropping **transmitters—including** the telephone tap." It is basically a souped-up while-noise generator.

Photo courtesy **Dectron** Industries, **Inc.**

illegal **antibug** jammers

While it's ethical to hire a countermeasure man to unearth the suspected bug, it is frequently much less expensive to simply buy a single jamming device guaranteed to render "all kinds of electronic surveillance devices virtually harmless." While illegal from an FCC point of view, these instruments can be turned on and off at will and do wonders for the paranoid corporate executive.

Known as "wiretap traps" and "antibugs," these gadgets are generally souped-up white noise generators that wreak havoc on bugging equipment, to say nothing of nearby radios and TV sets. Since they eliminate the need for

Once a bug has been found, the victim should consider the pros and cons before deciding to take one of the following moves:

Feed the bug wrong information purposely designed to make the eavesdropper show his hand.

Defeat its effectiveness by jamming; this often also forces the eavesdropper's hand.

Conduct a search to locate the monitoring post and/or eavesdropper.

Turn the matter over to a private detective agency (and lawyer) before tampering with the device.

Call the local police department now that evidence is at hand.

Carefully drain batteries without affecting bug's operation to accelerate the "death" of the device and move up the replacement time.

Defeat effectiveness by turning on running water (shower) or placing a radio in front of bug.

Notify FCC local office to reveal monitoring post without bringing in local police. FCC may take initiative in contacting authorities, but is discreet.

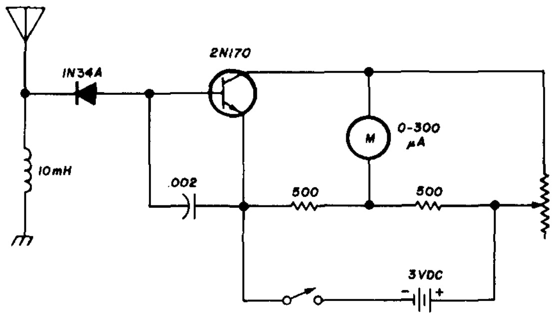
Remove device entirely.

visual and electronic detection techniques, however, they command good prices, usually from $225 to $350.

Other types include ac-powered versions of the super-regenerative feed-back detector. A single tube variety I experimented with used a single tube with about 100 Vdc on the plate. This little gadget put forth sufficient regeneration on the 88- to 108-MHz fm band to completely knock out a commercial 20,000-watt broadcast station at a distance of only 1/2 mile from the jammer. And this was accomplished using only a 3-inch link of hook-up wire as an antenna. Similar designs employed as bug-killers could totally wipe out a low-powered fm listening device at the listening end.

Crude spark-gap transmitters, the ultimate in hash generation, are occasionally em-

fig. 12. The basic field-strength meter is the backbone of tha bug-detection industry, although it is frequently souped up for added sensitivity.

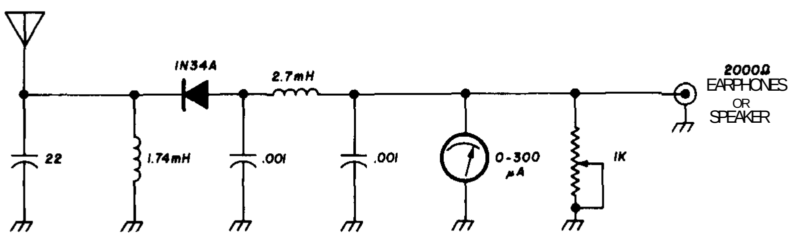


ployed today, but they are not mass-marketed for obvious reasons. In use, the spark-gap is placed fairly close to a source of high-level corporate conferences and powered by dry-cell batteries. By remote-control switching from any convenient point in the building, the spark-gap is actuated during periods of sensitive speech.

what the law says about transmitting bugs

On September 19,1966, Congressmen Moss and Gallagher introduced legislation aimed at the ultimate destruction of the surveillance

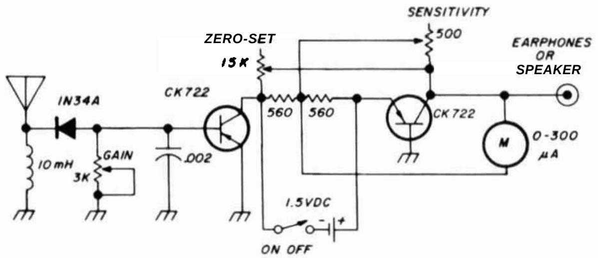
fig. 13. Circuit used most often in $100 bug detectors. An amateur with a well-stocked junk box can put one together for less than five dollars.



equipment industry (H.R. 17826 and 17827): "A bill to prohibit the shipment in commerce of electronic eavesdropping and wiretapping devices."

The legislation, if enacted, would give exception to "any department, agency, or in-

fig. 14. This simple two-stage field-strength meter is about as sophisticated as you'd care to go, since further amplification is unnecessary. In fact, with the circuit you may pick up local radio and tv broadcasts unless the gain is reduced



strumrntality . . . authorized to use such devices by Federal statute." This would take effect 180 days after passage of the bill. These bills were referred to the Committee on Interstate and Foreign Commerce, where they remained as the 89th Congress closed.

Earlier, the FCC adopted certain amendments to its Rules on April 8, 1966, designed to prohibit eavesdropping. Though entered as a specific regulation (Part 2, Subpart H, Section 2.701), the FCC spelled out the same ruling in Part 15:

"No person shall use, either directly or indirectly, a device required to be licensed . . . for the purpose of overhearing or recording the private conversations of others unless such use is authorized by all persons or parties engaged in the conversalions."

Further, this exception was noted:

"...This section shall not apply to operations of any law-enforcement officers conducted under lawful authority."

In Commission Chairman Rosel H. Hyde's 1966 year-end wrap-up statement, it was noted that since this rule change was adopted, the FCC investigated eight cases of alleged electronic invasions of privacy. A query I made to Commission Secretary Ben F. Waple, however, succeeded only in learning that such matters are "internal" FCC affairs and regarded "as confidential and not for release." Obviously, however, the FCC cannot apprehend all violators since the vast majority of bugging equipment is used only for short periods of time and is of an extremely low-power nature.

what the law says about telephone devices

Although no legislation is currently pending, it is hoped by proponents of anti-bugging that Senate investigations may permit some proposal to be worked up in 1968. The only agency that can now become involved is the FCC, which holds as a rule violation of Section 605 the "unauthorized interception" of phone conversations by direct wiretaps, or induction-coupling. Yet, for the Commission to take action, it must have evidence on hand that the

This bandswitching field-strength meter tunes from 25 to 250 MHz but reaches up to 450 MHz untuned. It will function over 400 hours with the internal mercury cell.

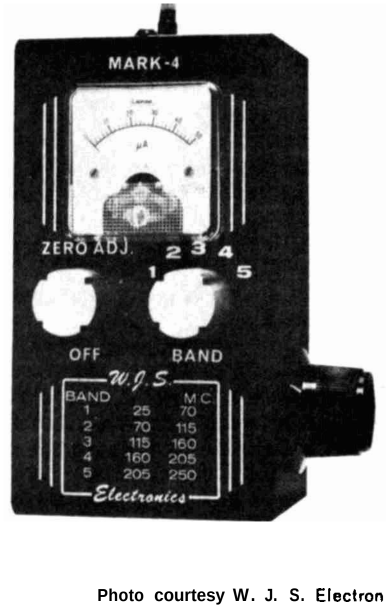


Photo courtesy W. J. S. Electroni

gleaned informationhas been clearly divulged or in some way "beneficial."

As phone-patch enthusiasts are well aware, many telephone companies have policies prohibiting attachment of any foreign device to its property. An illustration can be provided by this Wisconsin Telephone Company policy:

> "No equipment, apparatus, circuits, or device not furnished by the telephone company shall be connected with the facilities furnished by the telephone company, whether physically, by induction, or otherwise, except as provided in this tariff."

The tariff permits connection only of radio equipment of the Armed Forces, mobile telephone systems, and the major U.S. railroads.

on your own

Since hams are generally more interested in circuit technology of new breeds of devices, I have attempted to emphasize this aspect of eavesdropping rather than illustrative

Master bugger Fred Gluckman of Security Electronics demonstrates how an fm phone bug can be detuned to confuse the counter-measures investigator.

The following list covers **all** known major and minor **manufacturers** and distributors of **eaves**dropping equipment. Complete details on equipment types, prices. and street addresses can be found in "The Electronic Invasion" published by John F. Rider Publisher. **Inc.**, New York.

Baker Electronics Co.. Greencastle, Indiana

Britton Enterprises. Don, California and Hawaii

George F. **Cake** Co., Berkeley. California

R. B. Clifton Co., Miami, Florida

Consolidated Acoustics, **Hoboken,** New Jersey

Continental Telephone Co.. **Inc.,** New York, New York

Criminal Research Products, Inc., **Conshohocken**, Pennsylvania

Dectron Industries, Inc.. Van **Nuys,** California

Dee Co.. Houston, Texas

Dehart Electronics. Sarasota, Florida

Ekkotronics Co., Milwaukee, Wisconsin

Fargo Co., San Francisco, California

Fudelia **&** Associates Electronic **Surveillance** Devices, Toronto

Kel Corporation. Belmont. Massachusetts

Martel Electronics Sales. Inc.. New York, New York

Micro Communications Corp., Los Angeles, California

Miles Wireless Intercom, Ltd., New York, New York

Mittleman. Manny, New York, New York

Mosler Research Products, **Inc., Danbury, Connecticut**

R **& S** Research, **Inc.,** Houston, Texas

S.A.C. **Electronics,** Los Angeles, California

Saber Laboratories. San Francisco, California

Security Electronics, New York, New York

Sierra-Tronics, Nevada

Silmar Electronics. Inc., Miami, Florida

Spindel, **Bernard** B.. Holmes. New York

Steckler Sales Co.. New York. New York

Telephone Dynamics **Corp.,** North Bellmore. New York

Telmar, New York, New York

Tri-Tron. Inc., Dallas, Texas

Tron-X Publications. Hollywood, California

Wireless Guitar Co., New York, New York

W.J.S. Electronics. Hollywood. California

As the author has not had the opportunity lo visit the places of business or purchase equipment **from** all tha companies **represented** he cannot vouch for authenticity of product claims or equipment reliability.

This feedback detector sweeps over the frequency range from 6 kHz to 10,000 MHz with a broadband fm receiver configuration.

Photo courtesy R&S Research, Inc.

case histories showing where bugs have been unearthed, who has been indicted, and what corporations are involved. For much the same reason, this report has taken on an objective, factual approach and not one interjected with moral overtones.

Electronic invasions of privacy will continue to be part of the American way of life in the years to come, despite federal, state, and local efforts to suppress its growth. It is too well-entrenched now in U.S. business to do much more than go underground.

ham radio

references

1. "Electronic Bugging and the Ham," *CQ*, December 1966, page 34.
2. "Electronic Eavesdropping," *Electronics World*, April 1967, page 23.
3. "The Electronic Invasion," John F. Rider Publishers, Inc., 1967.
4. "The Intruders," published by Frederick A. Praeger, 1967.
5. "The Eavesdroppers: Fallout irnm R&D," *Electronic Design*, June 21, 1966

sporadic-E propagation predictions

Some of the best vhf propagation conditions oftheyear occur in early summer. Here are some forecasts of sporadic-E openings in June

Vic Frank, WB6KAP, 2450 Sylin e Boulevard, Woodside, California 94062

Although high-frequency propagation over paths greater than 2500 miles will be the primary emphasis in the months ahead, I am kicking off the propagation column with sporadic-E predictions for June, 1968. The curve in **fig. 1** shows the percentage occurance of sporadic-E measured at Point Arguello, California, during the first half of June, 1967, and is indicative of the sporadic-E propagation you can expect this year.

These curves show critical-frequency sporadic-E propogation versus time of day. The critical frequency is the highest frequency at which sporadic-E reflections were obtained with a **vertical** sounder.

Four critical-frequency contours are shown: 7, 9, 11 and 13 MHz. The critical frequency contour of 7 MHz corresponds to a range of 625 miles on 28 MHz; the 9-MHz critical frequency indicates a range of 470 miles on 28 MHz or 1400 miles on 50 MHz. The critical frequency contour of 11 MHz corresponds to a range of 800 miles on six meters; and a 73-MHz critical frequency corresponds to 625 miles on 50 MHz. For a sporadic-E opening on two meters, the critical frequency must go up to 24 MHz. The corresponding range on six meters would be 250 miles.

In summary, six meters will probably be open to distances between 1000 and 1400 miles more than half the days of the month, but you shouldn't miss many openings by sleeping between 0200 and 0600.

Two meters should open up to distances of 1250 to 1350 miles at least once during the month. The reflection point will probably be south of 38° north latitude between 0900 and 1500 local time at the path midpoint. Unfortunately, two-meter openings may not be noticed because of lack of activity at the proper places. More beacon transmissions and noon-time schedules over distances between 1200 and 1500 miles would help.

summary of high-frequency propagation

80 and 40 meters. On these two bands, summer-time noise levels and absorption will limit propagation to darkness paths. Since the north pole is in continuous daylight, no propagation is expected over the pole. This is not without merit—foreign broadcast stations will cause less interference than they do during fall and winter.

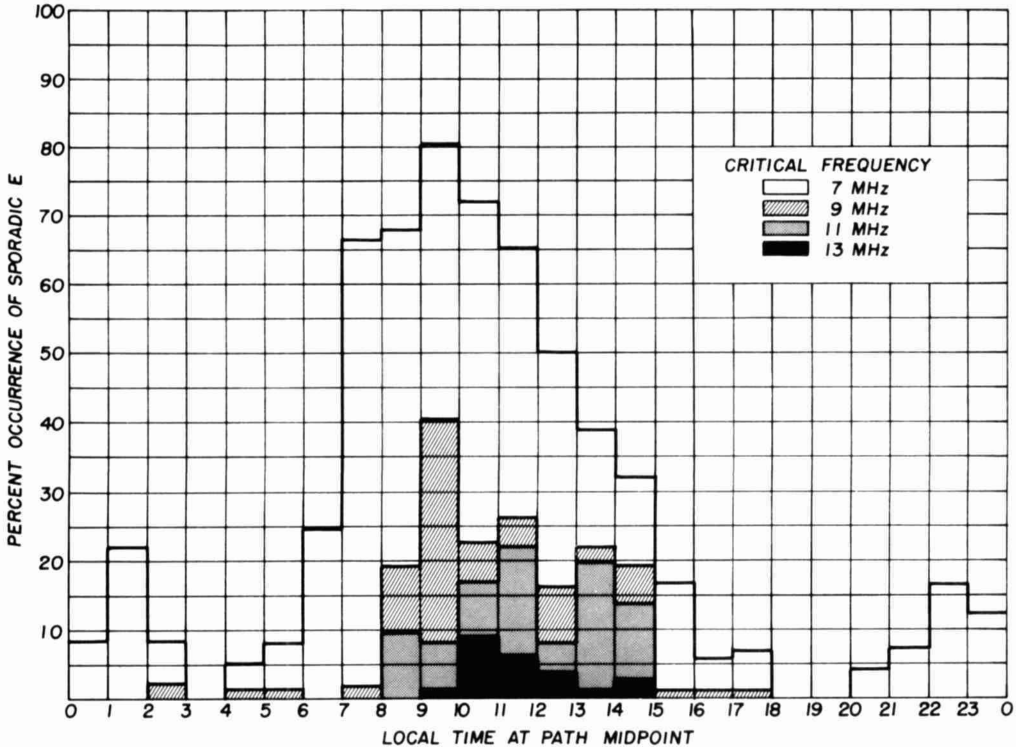
20 meters. Twenty will be primarliy a night-time band. The maximum usable frequency should be above 14 MHz for all but polar paths during the pre-dawn hours.

15 meters. Fifteen should be open during day-light and evening hours to directions south of east and west. Short skip may be prevalent because of sporadic-E.

10 meters. Ten should be open up from the southern half of the United States toward the south. Short skip sporadic-E propagation will be prevalent. Double-hop sporadic-E will permit occasional communications up to 2500 miles.

ham radio

fig. 1. Percentage occurrence of sporadic-E measured with a vertical sounder at Point Arguello, California during the first half of June. 1967.



new products

Main cw audio filter

If you are irritated by the QRM level on the CW bands these days, this new device will be of interest to you—the new Main Electronics high-selectivity CW audio filter. This unit offers very high selectivity for CW reception on all transceivers and receivers. To use the CWF-1, you merely plug into the 2- to 4-ohm audio output of your receiver and plug your headphones into the CWF-1. A switch is provided for taking the filter in or out of the circuit as interference dictates.

The passband of the filter is 120 Hz wide at the 6-dB points and 200 Hz wide at the 10 dB points. This is achieved by the use of high-Q toroidal inductors in a four-pole filter circuit. The output is designed to match 2000-ohm headphones. The filter not only separates the wanted signals out of the QRM, it also improves the signal-to-noise ratio when receiving weak CW signals which are close to the noise level such as in vhf DX work.

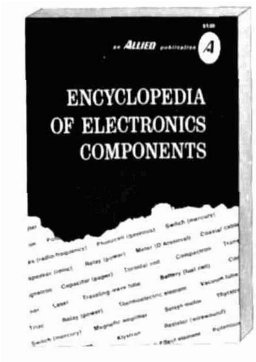
The CWF-1 Audio Filter is 2-718-inches wide, I-518-inches high and 4-inches deep. A descriptive brochure is available upon request; $19.95 from Main Electronics, Inc., 353 Pattie, Wichita, Kansas 67211.

Raytrack 50-MHz Converter

Raytrack Company has introduced a new six-meter converter using FET's in both the rf and mixer stages. These transistors give high immunity to cross modulation and a 3-dB typical noise figure. A 15-dB nominal gain is claimed. Features of the Horizon *VI* include a trap for TV Channel 2 and a built-in power supply. Crystal switching from the front panel permits a choice of expanded coverage to operating band segments. An output jack is also provided in the crystal oscillator circuit to permit transceive operation with your transmitter.

This converter has both a 150-ohm input and output impedance. It is designed to operate over the frequency range from 50 to 54 MHz with an output from 14 to 18 MHz. It is built on a fiberglass chassis and weighs only 18 ounces. The Horizon *VI* is priced at $59.95 by the Raytrack Company, 211 Springhill Drive, Columbus, Ohio 43221.

electronic components encyclopedia



A new encyclopedia of electronic components recently put out by Allied Radio alphabetically lists, describes and illustrates the basic electronic components currently in use. This book is virtually an electronics text that provides an understanding of individual units used in electronic devices and systems in one reading. Descriptions are completely non-technical. Each component is identified, and its use is carefully explained. In addition, any special handling or installation requirements are covered.

This is a handy reference book for anyone in electronics and is an interesting and useful aid to amateur radio operators, experimenters and students. One dollar postpaid in the USA. For more information, write to Allied Radio Corporation, 100 North Western Avenue, Chicago, Illinois 60680.

new Lafayette catalog

Lafayette has just announced its new 1968 Summer Catalog #684. This catalog is available free upon request and has the latest electronics and home entertainment equipment. It includes Lafayette's equipment plus many other major manufacturers, plus values in power tools for the home workshop, marine accessories and a complete line of amateur radio equipment, test equipment and citizens band two-way radio. It may he obtained free by writing to Lafayette Radio Electronics Corporation, P. O. Box 10, Department HR, Syosset, Long Island, New York 11791.

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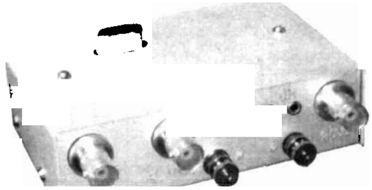

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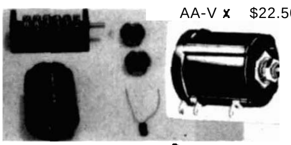

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T5	"	6.6 V @ 10 amps, c.t., 6.6 V @ 1.0 amps c.t.	$3.50
T6	"	6.6 V @ 18 amps. c.t.	54.75
T7	115 V, 60 Hz	44, 22 or 11 V c.t. @ 2.0 amps	$4.00
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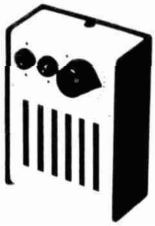

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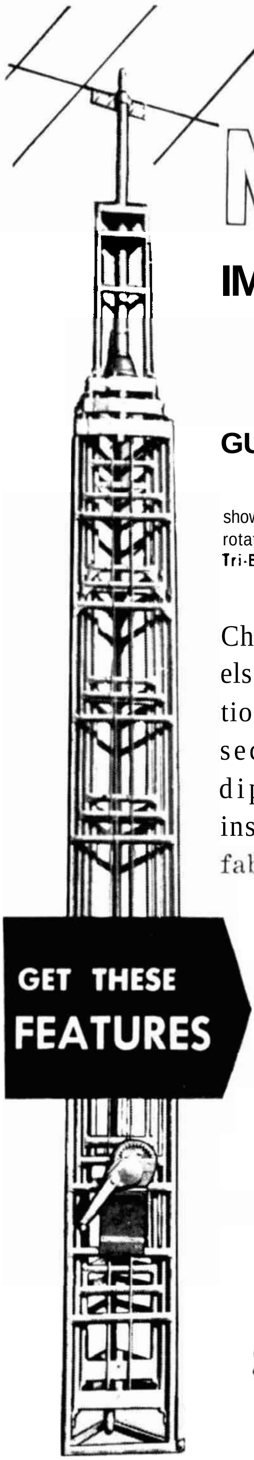

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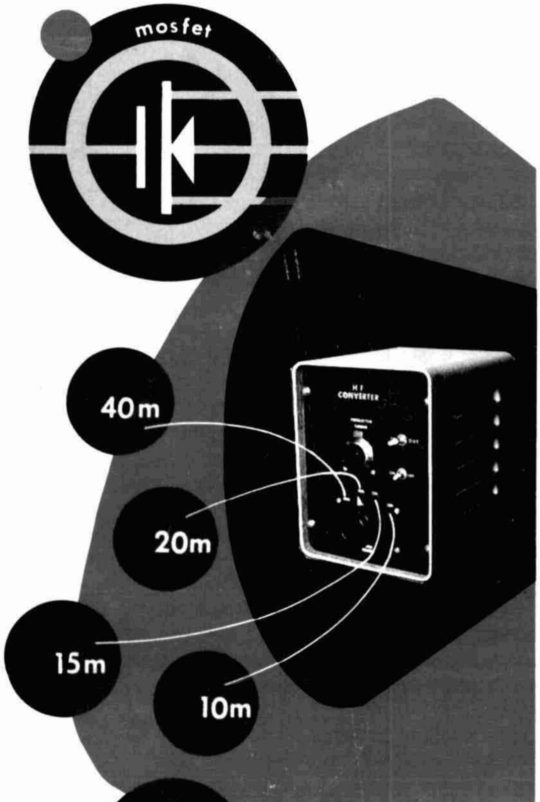

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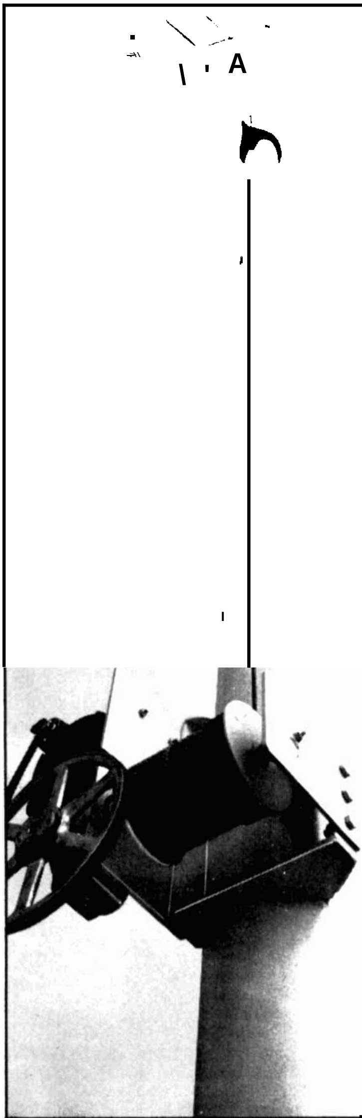

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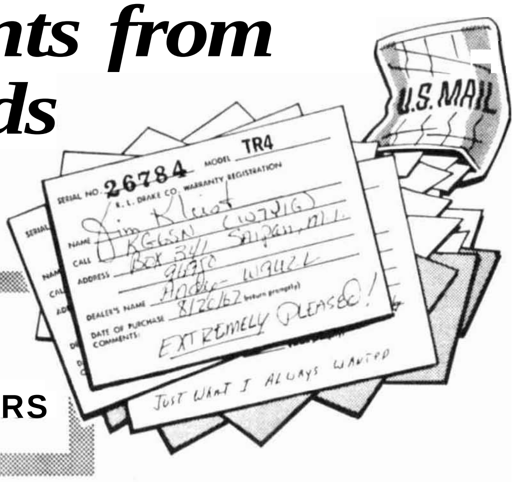

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"A beautiful **piece** of equipment. **My second piece of Drake. The first was a 2-B and this sold one friend an R-4** receiver and another a TR.4. We are Drake-minded here in town. Many thanks."

Charles E. Boschen Jr., WA4WXR
Ashland, Va.

"**I'm** sure this, **like** the other Drake equipment **I** have, is the finest money can buy. YOU MAY **QUOTE** ME ON THAT."

C. E. (Ed) Duncan, WA4BRU
Greenville, S. C.

"**I'm** a real happy man with **it. Does** a real good job of getting **thru.**"

Jerome D. Lasher, W2RHL
Hamburg, N. Y.

"**Replaces** my **TR-3.**"

D. G. Reekie, VE 6 AFS
Calgary Alberta Canada

"Finest performing gear **I** have ever had the pleasure of operating."

Milton C. Carter, W2TRF
Lakewood. N. J.

"**PS** Several months have passed . . . **I** now employ TR-4 as mobile unit and base station. **I** have logged more than **1000** contacts, many being rare **DX. I** am looking forward to owning a second unit to be used strictly for mobile. To date **TR-4** has been trouble-free."

Milton C. Carter, W2TRF
Lakewoob, N. J.

"**Well** pleased."

Rev. James Mohn, W3CKD
Lititz, Pa.

"**I** am delighted with Drake **gear. This** is the **second of** your **transceivers** for **me. I have used a TR-3 in my car for** about 2½ years—only trouble: replacing a fuse!"

Guy N. Woods, WA4KCN
Memphis, Tenn.

"Ask the ham who owns a Drake TR-4"

. . . or write for details . . .

Dept. 458 **R. L. DRAKE COMPANY** 540 Richard St., Miamisburg, Ohio 45342

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contents

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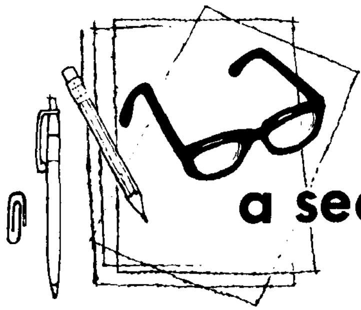


a second look

by jim fisk

I have been appalled by the number of bad operating practices that have been cropping up on our bands during the past few months. Evidently other people have been troubled too, because I have received several letters on the subject. None of these practices is new, they're just more prevalent. Deliberate interference, tuning up on net frequencies, playing music, calling CQ without listening first, obscene language, incorrect identification or no identification at all, using a kilowatt when 100 watts will do, talking crosstown on 20 meters instead of using vhf—the list could go on and on.

Our high-frequency bands are crowded, but deliberate and malicious interference, and discourteous operating tactics aren't going to relieve the situation. Everything is more crowded today; the population has exploded, the expressways and turnpikes are jammed, homes are being built on smaller and smaller pieces of land, and practically everywhere you go, you find a mass of humanity. It follows that we'll have congestion on the amateur bands—but congestion doesn't necessarily mean bedlam. Zeroing your kilowatt in on a QSO or local net isn't going to make them move. Why not join them? They'd probably be glad to have you.

Today, there is a net for almost every range of interest—they aren't restricted to handling traffic. Some of the groups that congregate on the bands are not really nets at all, but simply groups of hams who get together for a common purpose. There are DX nets, such as the International SSB'ers (who also handle traffic), the county hunters, the Cracker Barrel and Breakfast Club nets on 75 meters in the morning and various single-frequency gabfests. There are technical nets and the VHF Nut Net, for the vhf addicts, and of course, a multitude of local and intercontinental traffic nets, if traffic handling is your forte.

If you don't like net-type operation, fine; there are many amateurs who don't. On the other hand, if there weren't any nets, just imagine what the QRM would be like. There are thousands of amateurs who congregate on particular net frequencies; since they're a member of a net, they just "read the mail" a good deal of the time. If they didn't have the net, they would be calling CQ, fishing for a new county or active in one of the horrendous DX pileups. So, when you hear a net in operation, don't use it for a tuneup frequency. Whether you know it or not, the most hedonistic of them will stand by to handle traffic if asked to do so. They **all** do a service to the amateur fraternity by minimizing interference with channelized communications.

Six months ago, the FCC legalized "tail-ending." This is a big boon to the traffic handlers and the DX and contest operators because it allows them to transmit only their own callsign when calling another station. It also minimizes QRM because it lessens the amount of information that has to be transmitted. However, even with legalized tail-ending, you must still send both your call and the call of the station you are working at the **end** of the exchange.

The law is very explicit in this respect. Although you can send only your own call at the beginning of an exchange, or at intervals not greater than ten minutes, you must give the callsign of the station you are working or one of a group of stations that you are working at the **end** of an exchange.

During a recent DX test, it was remarkable to sit back and listen to the number of stations who never indicated the call of the station they were working. They simply sent their call, a signal report and contest number; the DX station came back with similar information. You could sit there for minutes on end waiting to hear the call of the DX station. When you finally gave up and asked him for his call, you'd probably find out that you'd worked him the day before! Interestingly enough, the sharp operators, the fellows who win the contests, were the ones who were the exceptions—they gave both callsigns at the end of each exchange.

Deliberate interference and incorrect identification are only two of the bad operating practices that you can find on any band you listen to. You can hear any number of stations working cross town on 15 or *20* meters when they should be on 75 or vhf. I have copied distant W/K stations on 20 meters, running well over S9 in New Hampshire, working their neighbors. With modern linears, it's a simple matter to turn the big box off when you don't need it.

Why all the penchant for S9 signal reports when you can maintain perfectly adequate QSO's with S6 or S7? You may need the linear for a long-haul DX QSO or for making initial contact, but once communication has been established, in 95% of the cases you can turn the linear off with no detriment to the QSO. In some cases, a kilowatt is necessary, but just because you own one doesn't mean you have to use it all the time. It isn't necessary and generates unnecessary interference.

I've heard a lot of stations go QRT because of interference and poor operating practices. This is not the answer. If you hear a station who has a bad signal, is not identifying properly, is causing unnecessary interference or being generally obnoxious, tactfully tell him about it. Most amateurs are gentlemen and will accept your suggestions with grace.

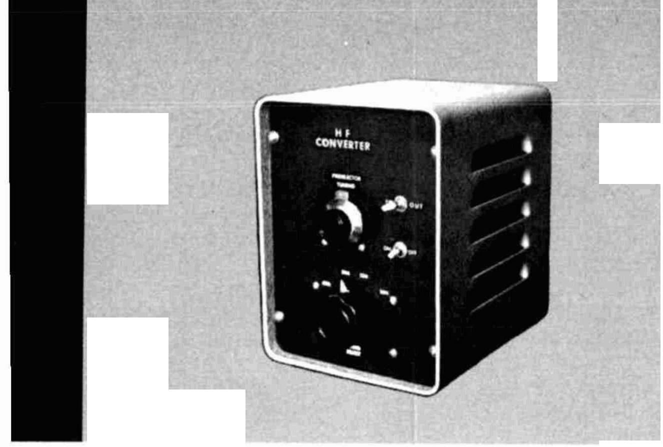
The next time you sit down at the operating desk, take a quick look at the rules of the ARRL A-1 Operator's Club before you turn on the transmitter. Try to follow their basic precepts for general keying and voice techniques, procedure, judgement and courtesy. Strive to be a first class operator; use operating finesse instead of brute force. If you're an A-1 Op, nominate the good operators you hear; if you're not a member of the club, make every effort to qualify. Let's promote good operating on our bands—discourtesy breeds pandemonium.

attention authors

I have a rather good short article in my files describing a unit that uses an integrated circuit—unfortunately, I don't know who wrote it! If you sent in something like this, write and identify it and I'll send you a check forthwith.

When you send in an article for consideration, please make sure your name and address are on the manuscript. It's helpful if you put your name or callsign on the back of each sheet, but that's not absolutely necessary. Same for photos. We haven't lost a manuscript yet, but occasionally an author gets mislaid because the only identification is the return address on an envelope.

Jim Fisk, W1DTY
Editor



bandswitching fet converter

A high-performance **four-band converter for 10, 15, 20 and 40 meters**

Mike Idstein, V3G, 22 Kingswood Road, Toronto 3, Ontario, Canada

The converter described here was originally designed as the front end of an 80-meter tuner[1] I recently built, the Ethersniffer, Mark II. Great care went into the design of the 80-meter tuner to obtain freedom from cross-modulation and overload, and to ensure good stability and sensitivity. The front end had to perform equally as well, so it wouldn't defeat the work that had gone into the tuner.

It didn't take much to convince me that the FET was the solution to many problems. It had all the advantages of solid state without the problems of the bipolar transistor; the literature shouted its virtue as a mixer, and it was easy to design around. Furthermore, no ham literature had come up with a bandswitching FET converter—a chance for a scientific breakthrough was at hand. Slide rule in hand, the foe was engaged.

design

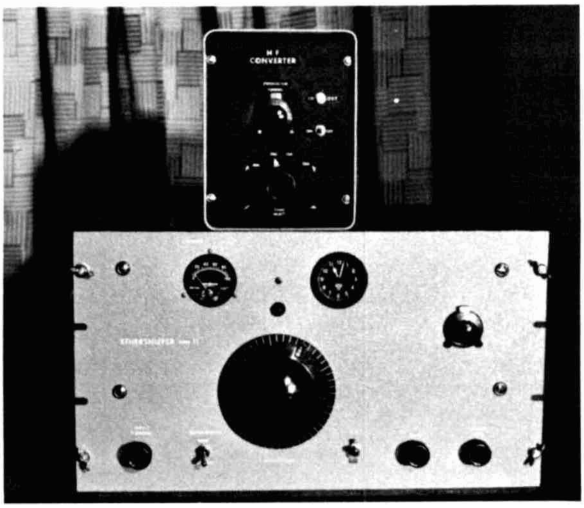
Two problems were presented. First of all, since the converter was all solid-state, I wanted to limit the size of the unit; the size of such circuits seems to be determined by the bandswitches. Related to this problem was the desire not to get involved in bandswitching a neutralizing circuit for each band. Since the converter was to cover 500 kHz on each range, it would be wise to be able to peak the rf and mixer circuits. This would provide maximum sensitivity and image-rejection across the band, but demanded a very stable circuit for smooth tuning.

The rf and mixer circuits are basically tuned to 20 meters, with shunt capacitance switched in to cover 40 meters, and shunt inductance switched in to cover 15 and 10 meters. Careful choice of circuit constants permits one shunt-connected inductor to raise the resonant frequency of the basic tuned circuit such that both higher bands can be tuned, while allowing complete coverage of each lower band (plus a little overlap) with the capacitance range available. The tuned circuits are sufficiently broadband so that tuning is only slightly critical on the upper two bands.

The design of such circuits is reasonably easy, particularly if you use a reactance slide rule. The only pitfall is when you neglect to consider the input capacitance of the FETs plus stray circuit capacitance. Murphy's Law states that you'll always fall into any pits presented, and I conformed.

I should mention here that you should check the operation of these tuned circuits with a grid-dip meter after wiring them. However, since we're using FETs, it will be necessary to remove the FET's from the circuits and substitute a fixed capacitor to represent the FET (about 6 pF should do) before using the grid-dip meter. Connect this capacitor between ground and the point where the FET gate is connected. Be sure to remove the capacitor before putting the FET back in the circuit.

The MOSFET **converter and 80-meter tuner—a complete high-frequency receiving set-up.**



I didn't feel it was necessary to bandswitch a tuned circuit in the oscillator output. Its only purpose would be to present sufficient output load impedance for plenty of oscillator output. Previous experience with a prototype indicated that the problem was to reduce oscillator injection to the mixer, not the reverse, so an rf choke was used as a broad-band load for the oscillator transistor.

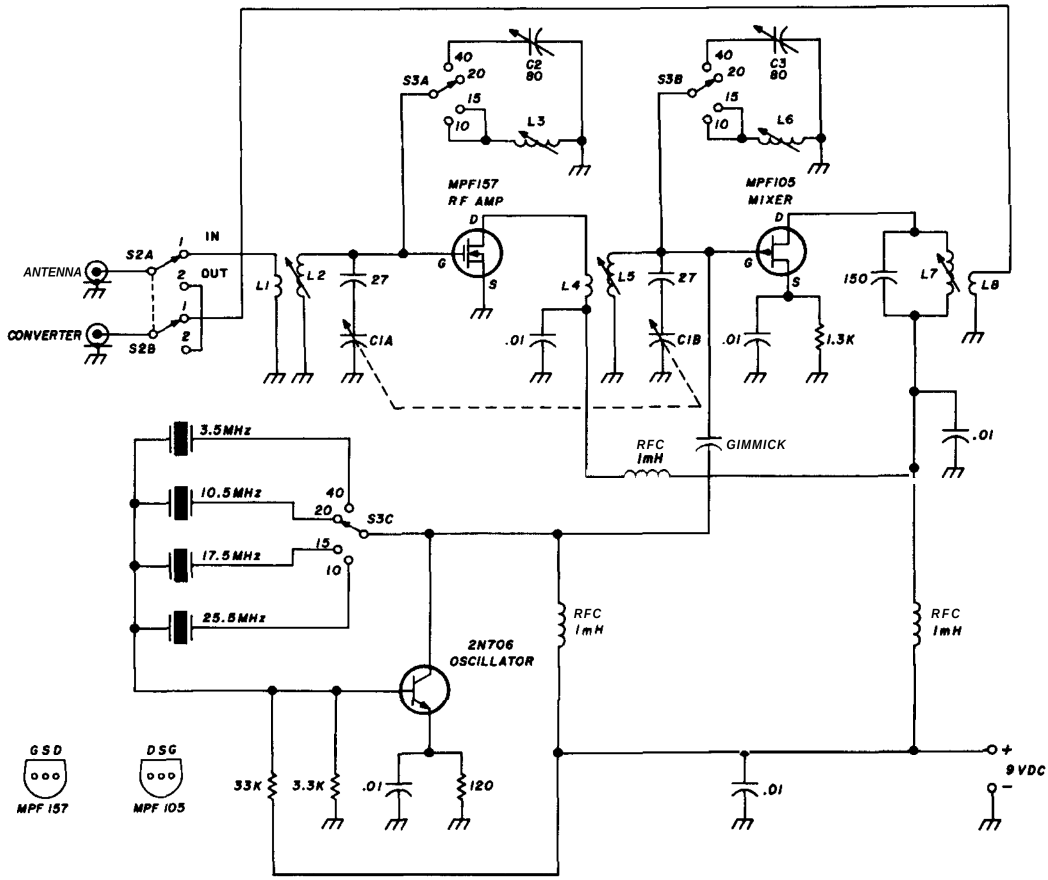
As the frequency of operation increases, the reactance of the rfc increases. This makes up for injection losses caused by decreasing oscillator efficiency. High-Q tank circuits in the rf amplifier and mixer minimize any problems caused by undesired harmonics from the oscillator. A simple "gimmick" capacitor (two paralleled wires twisted together for 112-inch) provides all the mixer injection necessary.

The second problem was not so much solved as ignored. I had hoped that with sufficient shielding and decoupling, it wouldn't be necessary to neutralize the rf amplifier to obtain the desired stability. This "head in the clouds" attitude was almost the undoing of the project, because the converter developed instability early in the game. Nothing (short of neutralizing) seemed to help until in desperation I took a look at the basic principles.

Experiments indicated that the feedback was caused by capacitance within the FET itself; I had defeated all coupling between circuits long-ago. How then, to eliminate this troublesome capacitance? A look at the specs on the Motorola MPF105 FET (my old faithful) indicated a reverse transfer capacitance of 3 pF. This seemed to be the dragon to slay. A quick call to the Motorola man yielded specs on a new MOSFET with all the desirable qualities of my MPF 105 JFET, plus a reverse transfer capacitance rating of less than 0.2 pF. A quick substitution, and the problem was solved.

construction

The converter was built on an aluminum chassis 5-314-inches wide, 7-inches deep and 3-inches high, the size dictated by the size of the bandswitch. The cabinet is a Ham-



C1	2-saction variable, 7-72 **pF (Millen** 21075-R)
Ll, L4	5 turns number 26 closewound on the cold ends of **L2** and **L5**
L2, L5	**6 μH. (CTC 2084-2) 25 turns number 26 enamel** clorewound on a **1/4"** slug-tuned form
L3, **L6**	**22** turns number 28 enamel closewound on a **1/4"** slug-tuned form
L7	11 pH. 40 turns number **36** closmound on a **1/4"** slug-tuned form
L8	12 turns number 28 enamel on cold end of L7
S1	**SPST toggle switch** (in power **supply)**
S2	DPST toggle switch
S3	**3P4T ceramic** rotary **switch (Centralab** 2008)

fig. 1. **Bandswitching** MOSFET converter for 40, 20, 15 and 10. A war-surplus variable war used for **C1,** but the **Millen 21075-R** is a close equivalent.

mond 1401-B*, which just accommodates my home-bent chassis nicely. The shields shown in the photograph were custom shaped with a nibbling tool after installing the band-switch.

The wiring is shown in the photographs; it is not critical, but it's a good idea to keep all the leads as short as possible. Solidly mounting all components will result in good mechanical stability. I didn't use any "special" components, so it should be reasonably inexpensive to duplicate.

*if you can't locate a Hammond distributor, the Bud WA-1540 Portacab is a reasonable facsimile. Available for $10.60 plus postage from Allied Radio, 100 North Western Avenue, Chicago, Illinois 60680. Weight, 2-1/2 pounds.

alignment

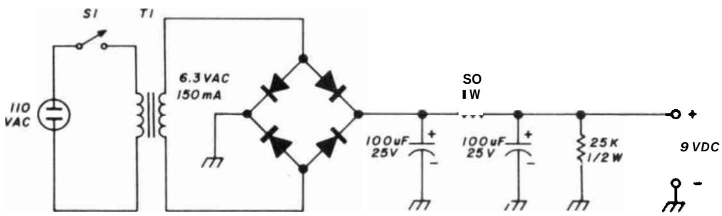
Alignment is simple. Switch the converter to the 20-meter range, and tune the i-f amplifier to 3650 kHz (14,150 kHz). Set the converter tuning capacitor to half-mesh. With a signal generator (or a received signal) at the input, adjust both rf amplifier tuned circuits for maximum signal output from the i-f amplifier. Adjust the mixer circuit first. Now peak the mixer output circuit for maximum signal output. This completes 20-meter alignment.

Set the converter to the 40-meter range. With a suitable input signal, adjust the shunt capacitor in the mixer circuit until the correct mixer beat is found, then adjust it for maximum signal output. Now, adjust the rf amplifier circuit for maximum signal output. This completes the 40-meter alignment.

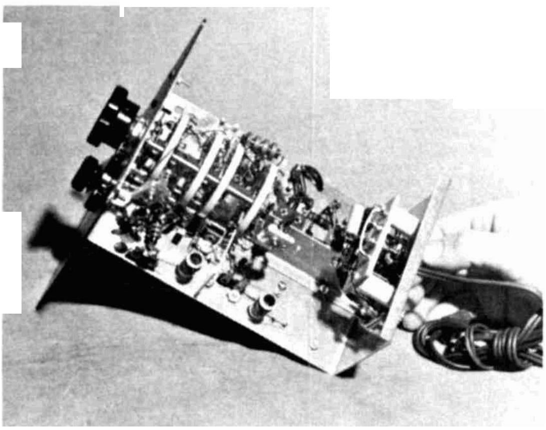
Alignment of the top two bands is a bit touchy because they are both covered by the same tuned circuits. However, with a little care, disaster can be avoided. Set the converter to 15 meters. Adjust the tuning capacitor for maximum mesh, and set the i-f amplifier for 3500 kHz (21,000 kHz). Tune the shunt inductance in the mixer circuit until the correct mixer beat is found, then adjust it for maximum received signal. Adjust the rf amplifier the same way. Now switch to the 10-meter position. If the tuned circuits are correctly aligned, it should be possible to peak up received signals on the bottom end of the 10-meter band with the tuning capacitor set close to minimum mesh. This completes converter alignment.

The photographs were taken before the shunt inductances were wired in; they are installed adjacent to the 20-meter coils. Room was provided for a total of five crystal sockets.

fig. 2. Powar supply for the bandswitching convertar. T1 is a Knight 54E1416 or similar; the diodes can be anything that will handle 150 mils.



Under the converter chassis. The mixer section is to the right near the front panel; the rf amplifier is toward the rear. The power supply components are mountad on the rear deck. Only three sections of the rotary switch am used.



performance

Performance of the converter has been quite satisfactory. It shows no tendency to overload whatsoever, is extremely quiet, and quite sensitive. Once the tuned circuits are set to the middle of a band, operation over the entire band is possible without repeaking them. However, peaking them on a weak one does improve matters. No spurious signals have been detected in the i-f amplifier when the tuned circuits were properly peaked. The tuning is very smooth and there is no evidence of instability.

ham radio

references

1. M. Goldstein, VE3GFN, G. Cousins, VE1TG, "High Quality Hybrid Receiver," 73, February, 1968, p. 42.

an ultrastable solid-state 10-watt 28-mhz transmitter

Robert M. Brown, K2ZSQ/W9HBF, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

Due to the perseverance of ITT's Gary Jordan — who should be credited for unearthing this unusual variable-frequency oscillator amid reams of British and Czechoslovakian technical literature — amateur designers now have at their disposal a VFO described by **The Electronic Engineer** as "the first oscillator showing promise of eclipsing performance of all other self-sustained oscillators."[1] The VFO actually uses a single transistor in a circuit with all appearances of a Clapp oscillator, except that the method of feedback is different.

exceptional stability

Three distinct performance characteristics combine to make the Vackar concept highly desirable. First, it has the greatest inherent stability of any known VFO design other than that with an independent external load feedback. Second, it can tune over a frequency range of nearly 3:1. Third — and especially significant for amateur purposes — the output over that frequency range is absolutely constant.

As with any unusual oscillator, however, problems have been encountered. The Vackar has a tendency to oscillate at audio frequencies simultaneously with rf generation. This problem can be overcome by decoupling the base-bias resistors from the collector load resistor — using an appropriate value resistor — and not an rf choke. A resistor will attenuate both audio and rf feedback simultaneously; an rf choke will not. As a precautionary move, it is also helpful to bypass the emitter both at audio frequencies and rf frequencies, although this is not absolutely necessary.

putting the Vackar on the air

Getting down to brass tacks, the extremely stable characteristics of the Vackar lends itself to hf applications where conventional designs fear to tread. The basic circuit is shown in fig. 1, adapted from Vackar's original concept (illustrated in **fig.** 2).

fig. 1. Want just the VFO? Here's a Vackar you can put to work immediately. Choose a 25-MHz coil/capacitor combination and you'll have stable drive available for your 6-meter transmitter.

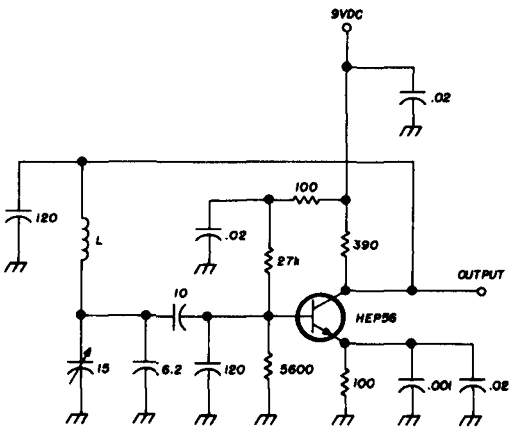
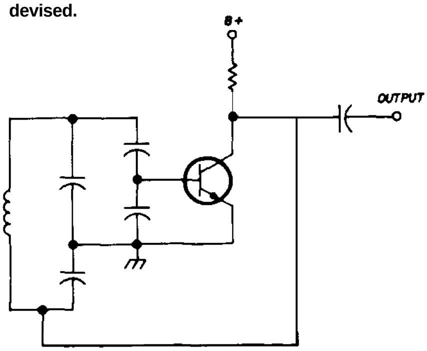


fig. 2. Jiri Vackar's original oscillator circuit, developed in Czechoslovakia in 1940. British and American engineers now tout it as one of the most brilliant VFO circuits ever devised.



A conventional small power amplifier chain has been built providing a power output of approximately 10 watts over the 28-MHz ham band. No creative claims are made for this portion, since it was constructed mainly of scrounged components found around the work-bench.

Standard VFO isolation techniques should be observed, of course. Decoupling has been included in the circuit to minimize audio oscillation and an emitter follower provides isolation. The final tank circuit uses a pi-L circuit with a capacitive divider to match the transistor output impedance. The transmitter can be built in a small mini-box equipped with wall-divider sections as indicated in the schematic.

performance

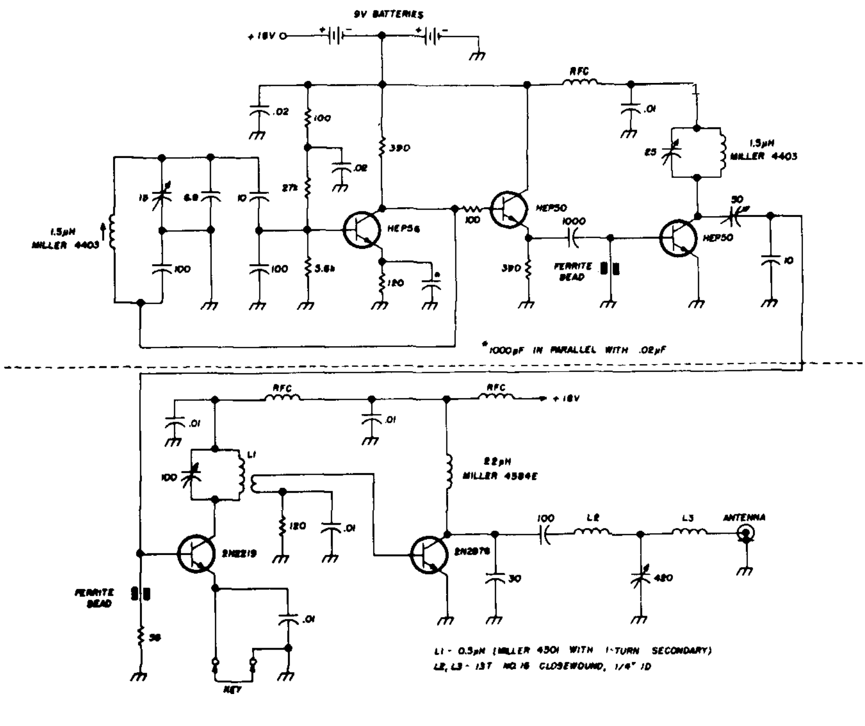
As a hf VFO, few designs can compare with the Vackar in stability and constancy of output over a wide frequency range. The ten-meter CW transmitter shown here can provide long hours of enjoyment to the ten-meter DX fraternity.

references

1. G. B. Jordan, "The Vackar VFO, a design to try," The Electronic Engineer, February, 1968, p. 56.
2. J. R. Fisk, W1DTY, "Stable Transistor VFO's," ham radio, June 1968, p. 14.
3. J. Vackar, "L-C Oscillators and their Frequency Stability," *Tesla* Tech Reports, (Czechoslovakia), December, 1949.

ham radio

fig. 3. Circuit of a 28-MHz transmitter using a Vackar VFO. Power output to the antenna is 10 watts.



transistorized 455-kHz bfo

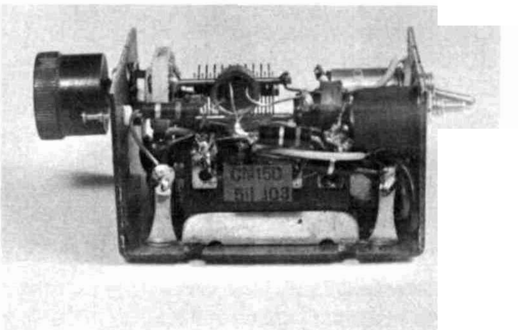
A stable outboard BFO for use with transistor radios

Ed Marriner, W6BLZ, and John Meredith, K5GXR/6, 528 Colima Street, La Jolla, California

Probably a lot of radio amateurs took advantage of the all-band transistor radio recently offered by a national oil company at a special low price of $39.00. The brochure they sent out showed a receiver with a vernier tuning rate and bandspread on 160, 80 and 40 meters of about one inch. Obviously, here was a portable radio that had all the possibilities of a picnic-table portable. All you needed to copy CW was a BFO. Needless to say, we bought the receiver.

The next step was to search through all of the old literature for a transistor BFO, and about a dozen circuits were found. However, none of them wanted to oscillate after they were built. K5GXR/6 took up the problem. With a Tektronix Oscilloscope, a frequency counter, a grid-dip oscillator, a variable-voltage power supply and some substitutions from the junk box, he did manage to make a BFO work. Kibitzing on the side by other amateurs who were digital-circuit engineers

Internal layout of the bfo.



also helped to make the project a success. By a process of elimination, the circuit described here seemed like the one most likely to succeed for the amateur with no test gear. A grid-dip oscillator or low-frequency receiver is a help if the tank circuit is very far off the design frequency of 455 kHz.

construction

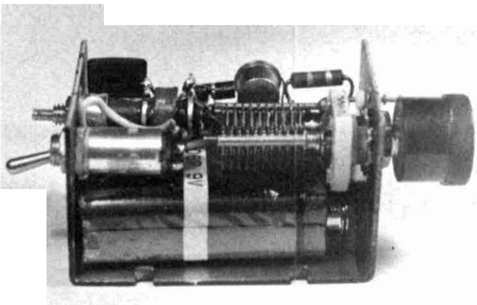
The BFO was built in an aluminum box 2-114 x 1-1/2 x 1-3/8 inches; LMB type MOO. At first we were going to build the unit in a plastic box, as in other articles, to take advantage of radiation coupling into the receiver. This idea did not work out because hand capacity made the signals wobble. In addition, the tuning capacitor had to be fastened down to the shield can for the same reason.

This unit was built quite small and you can see from the photographs that the parts are crammed into a tiny space. Some builders might want more room, hut nimble fingers should have no trouble soldering all the parts in place. All the holes were drilled by rule of thumb, a weather eye, and a little hand fitting. A four-terminal mounting strip was used to tie down the components.

It's a good idea to hook up the coil and capacitors and tune them to frequency before putting them in the box. Because of the low frequency, the coils are low *Q,* and some 1-mH slugged-tuned coils don't seem to tune. Once the circuit dips, put it in the box. After the oscillator is finished, set the variable capacitor mid-scale and adjust the coil slug to 455 kHz.

Use the same components that are called

The parts a n crowded, but they'll all fit.



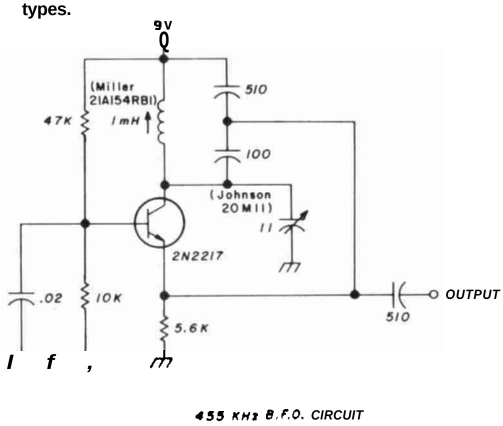
out on the drawing. Several of these units have been built and they all worked. It took a lot of fiddling to find the proper values, so don't duplicate our mistake by thinking something else will work—it might not!

The oscillator will run on any supply voltage between three and nine volts. The 9-volt battery was used because it fit into the bottom of the chassis box very nicely.

testing

About all that can be said about testing the unit is to set the capacitor half scale and turn the unit on. Hold it near the transistor radio or connect a wire from the feedthrough insulator to a piece of hookup wire wrapped around the transistor-radio antenna. This will couple in enough signal to beat with the i-f signal. Vary the coil slug until a zero beat is

fig. 1. Schematic diagram for the 455-kHz bfo. The two capacitors in the tank circuit should be silver-mica types.



received. Use the capacitor for fine adjustments. A wobbly signal is due to an unstable transistor radio; the BFO is quite stable. Tests over a three-hour period indicated a total drift of 186 Hz.

This circuit can be used for other oscillator frequencies by scaling down the LC ratio. Remember that the total tuning capacitance is in series across the coil and the total capacitance of the circuit is less than the smallest capacitor. While we didn't experiment with this circuit on higher frequencies, it seems like the easiest one that will oscillate.

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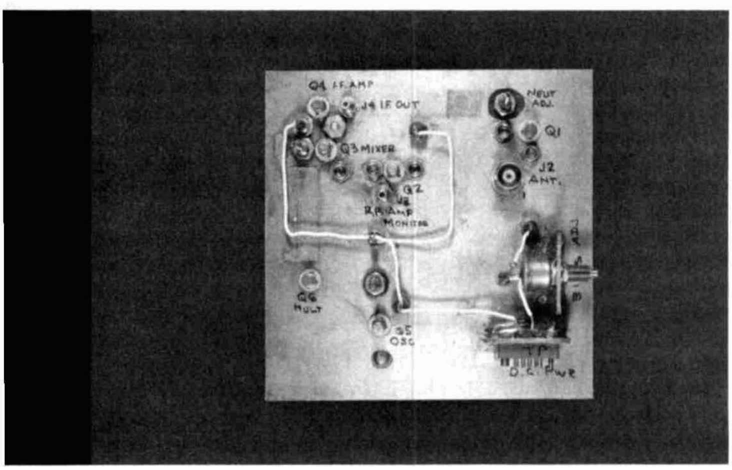


1.5-dB noise figure on two meters

Complete construction details for a high-performance FET converter for two meters

Bob Kolb, WA6SXC, 1388 West Oak Avenue, Fullerton, California 92633

I've built quite a few converters over the years, and I have to chuckle when I think of my first attempt. It was a single 12AT7 on six meters. I used 12-volts dc on the plates and filaments and down-converted to the broadcast receiver in my car. Other operators could hear my local oscillator radiating a mile away. My next endeavor used those new-fangled nuvistors on two meters. I really thought I had something! I built another unit for 432 and felt pretty smug.

Then noise figure became the rage. You just weren't in the "in crowd" unless your noise figure was less than 3 dB on two or 4.5 dB on 432. So, I got some 7077 ceramic triodes and built some preamps. I finally made the grade, but the fad switched to transistors; out came the soldering iron and the old 7077 preamps were replaced with 2N2857's, but there wasn't too much improvement in noise figure. I thought I was through for a while, but someone started making junction FET's for vhf. I tried a 2N3823 on two meters but was unimpressed; I wasn't pleased with TIS 34's in cascade either.

Then I built the 432-MHz receiver front end for the ill-fated ARIES satellite program and tried several varieties of exotic bipolar transistors. When the 2N4416 and 2N4417 junction FET's became available, I decided to try them. The results were astonishing. The 432-MHz converter design I came up with was published in the May, 1968 edition of **ham radio.**[1] Right afterwards, I began building the 2-meter converter described here.

the converter

This converter is an exceptional performer. It has a 5-MHz, 3-dB bandwidth and a calculated 1.52-dB noise figure. It will copy a carrier modulated 30% with 1 kHz at a level lower than –127 dBm (less than 0.1 μV). When 98- to 138- and 150- to 180-MHz signals at 0 dBm (1 mW) were inserted into the antenna connector, no output was detected when a field intensity meter (sensitivity less than –85 dBm) was tuned across the i-f band. Noise figure was determined by measuring an i-f signal 3 dB above the noise with a –121-dBm input modulated 30% with 1 kHz. A Stoddard NM-30A calibrated receiver with a 3-dB bandwidth of 140 kHz was used as the tunable i-f as shown in **fig. 3.**

circuit description

The rf amplifier section is a cascode arrangement using a neutralized common-source stage followed by a common-gate amplifier. The two stages are coupled inductively, but C1 was added to increase circuit bandwidth by overcoupling. FETs behave somewhat like tubes and have high input impedance. The loaded Q's are much higher than bipolar transistor circuits, and broad bandwidths are difficult to achieve without overcoupling. The increased coupling also increases the load on Q1 and makes the stage more stable and easier to neutralize.

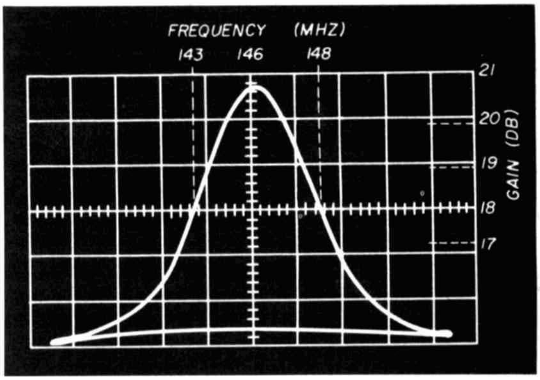
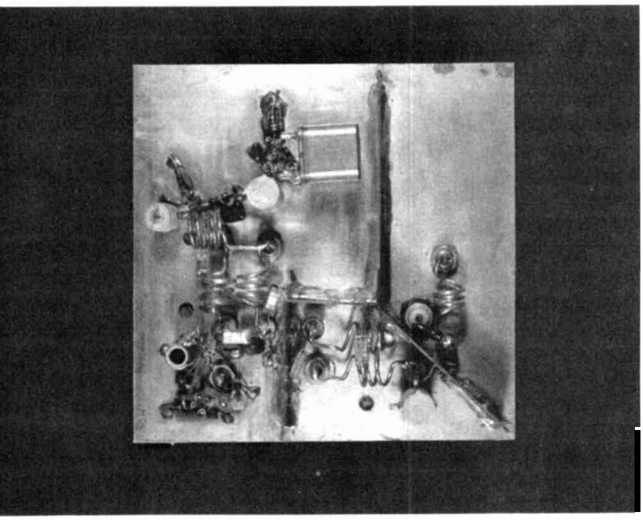


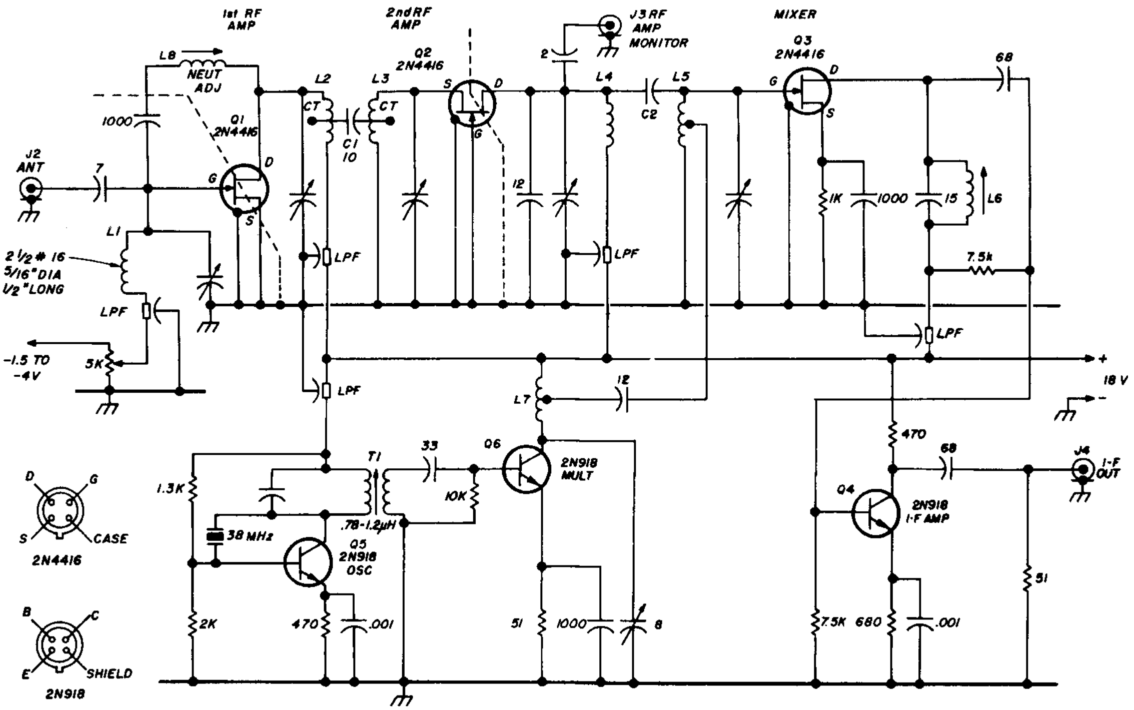
fig. 1. Bandpass characteristic of the converter. The 3-dB bandwidth is from 143 to 148 MHz; and 1-dB bandwidth is from 144.8 to 147.0 MHz. Gain at 146 MHz is 21 dB.



Parts layout of the 2-meter converter. The local-oscillator chain and mixer are to the left, the rf amplifier stages to the lower right. The shield partitions in the rf stages are constructed across tha canter of the transistor sockets.

The bandpass of the rf amplifiers as measured at the rf amplifier monitor jack (J3) is shown in **fig. 1.** If you're only interested in operating in the first one megahertz of the two-meter band, omit C1. The same coupling technique is used between the rf amplifiers and the mixer. The output of the mixer is applied to an untuned amplifier to drive the coax to the receiver and isolate it from the receiver input circuit. The mixer is operated slightly above pinchoff where it is fairly nonlinear.

The i-f for this converter was chosen for the 29.5- to 34.5-MHz converter band on the NC-300 receiver. However, the oscillator will tune to 38.6667 MHz, and C4 can be adjusted



L1	**2-1/2** turns number 16, space wound, 5/16" diameter, **1/2"** long	L6	20 turns number 24 on **1/4"** ceramic form (CTC PLS-6 with white core)
L2, L3, L4	3 turns number 16, space wound, 5/16" diameter. 1/2" long	**L7**	4-1/2 turns number 16, 5/16" diameter, **1/2"** long, tapped 2 turns from **B+** end
L5	4 turns number 16, space wound, 5/16" diameter, 1/2" long, tapped **1-1/2** turns from cold end	**L8**	1.7 to 2.7 μ**H** (J. W. Miller 4503)
		T1	0.78 to 1.2 *pH* (Vanguard **LT4J-7206)**

fig. 2. Schematic diagram of the low-noise converter for two meters. Except as noted **otherwise,** all variable capacitors are **10-pF** pistons (Johanson **JMC-2954); LPF's** are feedthrough filters (Allen Bradley SMFB-A2 or Erie 1201-050).

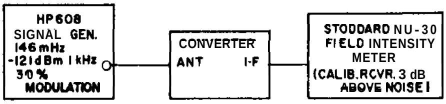
for 116-MHz output from the multiplier. This will produce a 28-MHz i-f at 144. Transistors such as the 2N918, 2N3564, 2N4324, and 2N2369 have been used with equal success in the oscillator, i-f, amplifier and multiplier circuits. The 2N3564 and 2N4324 are low-cost epoxy types.

The photos show a breadboard model of the converter. Each stage was evaluated independently and then the connectors were removed except for the rf amplifier-monitor jack (J3). You can see the extra holes in the breadboard. Optimum local-oscillator power is approximately +13 dBm or 20 mW. This may sound quite high, but the LO is very lightly coupled into the mixer to maximize LO isolation.

construction details

After considerable discussion with W6DQJ, I decided to use sockets for the transistors; this was in opposition to my personal preference. My attitude in this regard has only been strengthened by this experience. The use of a socket invites intermittents unless you solder the transistor to the socket. Another version under construction at this time will use no sockets or holders of any kind; only standoff terminals. The transistor will be soldered upside down as shown in **fig. 4** with its leads soldered to the terminals above the other components. With this arrangement,

fig. 3. Test setup used to measure the sensitivity of the two-meter converter. Noise figure of the unit was calculated to be 1.52 dB.



replacement or substitution of transistors is quite easy.

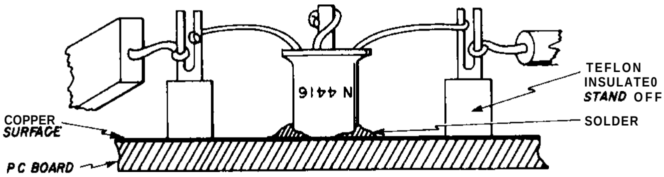
At the expense of being considered unsophisticated, I wound all the coils except L6 and L8 around the handle of a standard X-acto knife (5116-inch diameter). The turns were spaced approximately 118-inch apart with the blade of a pocket screw driver. The chassis is 1116-inch epoxy-glass printed-circuit board. All capacitors were glass (CY10 series) or 2% mylar Elmenco DM15's.

source and case connections to Q1 and the gate and case connections to Q2 are soldered directly to the shield. A small hole (about ,040-inch diameter) is drilled through the shield and the copper around it is removed in a 118-inch diameter circle to allow the lead from the 1000-pF capacitor in the neutralizing circuit to be connected to L8.

The drains of both Q1 and Q2 must be shielded from their gates and sources. All air-wound inductors are mounted with their center lines parallel with and 0.3 inches above the chassis as shown in **fig. 5.** All insulated standoff terminals are the press-in teflon type. All 1000-pF emitter and source bypass capacitors are 50-V disc-ceramic types with zero lead lengths. The crystal is connected into the circuit by tube-socket jacks removed from a phenolic 7-pin miniature tube socket. These work well with the HC-6/U crystal holders.

The copper can be removed quite easily from the printed-circuit board by cutting the outline of the area to be removed with an X-acto knife and lifting out the area with the tip of the knife as hot solder is flowed over the area. All ground connections were made

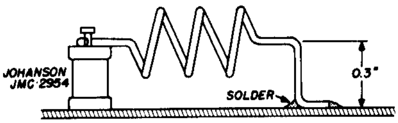
fig. 4. Method used for mounting the transistors when sockets aren's used.



It is important to strip the copper from around L8 so that the mounting nut does not contact ground because the tuning slug of L8 **must** be isloated from ground. Transformer T1 is a Vanguard LT64 series-tunable transformer (0.8-1.2 pH). All dc lines were bypassed with Allen Bradley FMB-A2 low-pass feedthrough filters, but I have found the Erie 1201-050 filters are also satisfactory. Standard teflon feedthrough terminals may be used if the dc points are interconnected with 10-pH inductors or Ohmite Z144 rf chokes. All feedthroughs should be bypassed with 1000-pF disc ceramic capacitors with zero lead lengths.

The chassis is 5-inches square with 1/32-inch thick x ,950-inch high partitions. The

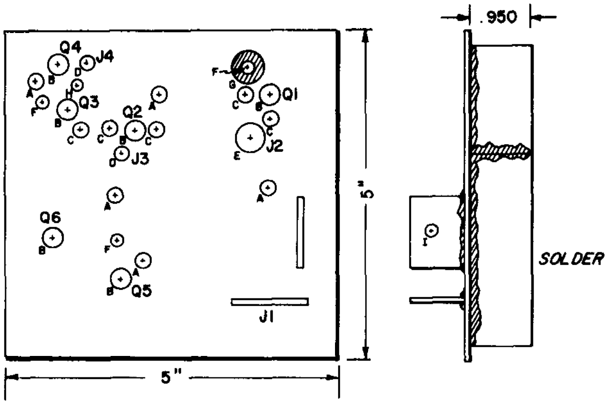
fig. 5. Mounting the air-wound inductors used in the converter.



directly to the nearest point on the board. The addition of the i-f amplifier was an afterthought; hence, the circuit is a little crowded.

tune up

Connect the B+ so that you can monitor the drain current ϕ of Q1 (10-mA meter in fig. 6. Layout of the copper-clad board used in the 144-MHz converter. The power connector (J1) is mounted on one vertical upright; the 5k bias-adjust pot on the other.



A .213" FOR FMB-A2 LOW-PASS FILTER
B 5/16" FOR TRANSISTOR SOCKETS
C .228" *I FOR JMC-2954 CAPACITOR
D .189" FOR SELECTRO #3102 CONNECTOR
E 3/8" FOR BNC CONNECTOR (J2)
F 5/32" FOR L6, L8, & T1
G 1/2" DIA. COWER STRIPPED FROM BOTH SIDES OF BOARD
H #28 FOR 1000-pF STANDOFF CAPACITOR
I 1/4" FOR BIAS ADJUST

series with +18V and LPF2). Connect the negative supply voltage and the antenna. Set the bias-adjust pot for 5 mA of drain current for Q1. Set the input trimmer capacitor at about midrange and the drain tank capacitor about 75% in with L8 nearly out. Rock L8 back and forth and make certain the drain current stays steady. Disconnect the antenna, and note any change in the drain current of the first stage (I_D). If I_D changes, Q1 is not neutralized. Repeat the adjustment of the two tank circuits and L8 with the antenna alternately connected and open circuited until I_D is stable at 5 mA. Attempts to neutralize Q1 by minimizing the signal through it without B+ applied will not work because junction capacitance changes significantly when B+ is applied.

The oscillator and multiplier are tuned by monitoring the dc voltage on the multiplier emitter resistor. Next, apply a signal to J2, and successively tune each circuit for optimum. Finally, tune in a weak station, and tune for maximum; then sit back and enjoy that well-earned QSO.

reference

1. R. Kolb, WA6SXC, "A Low-Noise 432-MHz FET Converter," ham *radio,* May, **1968, p. 18.**

ham radio

optimizing vhf converter performance

At this time of year a lot of vhf enthusiasts are getting ready for the DX season by putting up new antennas, replacing old gear, etc. During the big revamp, why not check the B+ feeding your converter? Few amateurs realize that most commercial converters (and many homebrew jobs) are overdriven to the point of decreased performance.

The B+ feeding the converter can be adjusted for optimum S/N ratio by inserting a pot temporarily in the line. With careful adjustment, you can often get 1 dB or more on weak signals and up to 7 dB on the stronger ones. After you've optimized the size of the resistance, you can replace it with a fixed value.

This is also a good time to get your signal generator out and tweak everything for maximum. Don't forget noise figure; it's important to keep it as low as possible. If the noise figure is reduced from 15 dB—a not uncommon figure—to 5 dB, it is the same as increasing the power of the signal you're listening to by 10 times.

ham radio

phase-shift ssb generators

There's more than one way to generate a ssb signal– the filter method was covered earlier: here's how the phase-shift method works

Forest H. Belt, 119 Baker Avenue, South Plainfield, New Jersey 07080

There are two ways to generate a single-sideband signal. In the one most popularly used, a double-sideband suppressed-carrier (dsbsc) signal is first developed by feeding an rf signal and an af signal into a balanced modulator. The dsbsc signal is then passed through a filter that removes one sideband. This is known as the **filter method** of obtaining an ssb signal.

The other way, seldom found in commercial ham equipment today, is called the **phase shift method.** The rf and the af signals are fed to a pair of balanced modulators, introducing enough phase shift in each so that, when the resultant is finally mixed, one sideband and the carrier are suppressed. Phase-shift ssb generation has been a mystery to many readers. Therefore, no series of articles on single-sideband would be really complete without a detailed explanation of how the phase-shift method works.

advantages and disadvantages

One of the major advantages of the phase-shift method of generating a single-sideband signal is that the method can be used at any frequency. In the filter method, an rf signal is used initially. After one sideband is removed, several stages of frequency translation are usually necessary to bring the ssb transmitter signal to a high or very-high frequency. Additional filters may also be needed, because every time the signal goes through a stage of translation, another double-sideband signal is generated (unless a balanced mixer is used). With the phase-shift system, the single-sideband signal can be generated right at the transmitter output frequency, even if it is all the way up in the vhf range.

The system isn't used much in commercial ham radio gear for two major reasons. In the first place, sideband and carrier suppression are not as thorough as with a well designed and properly adjusted filter system. A good filter-type single-sideband exciter can achieve carrier suppression from 40 to 60 dB, depending on the type of balanced modulator used. In the phase-shift generator, only the best design can achieve more than 30 or 35 dB of suppression.

This is adequate for practical use, and yet it can be a little annoying in crowded ham bands. A bit better suppression can be achieved with special tubes such as the RCA 7360, but even that depends on critical adjustment of the circuit.

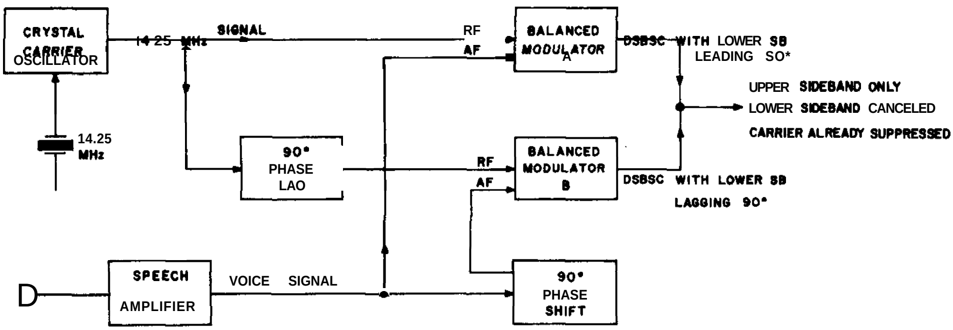
Which brings us to the second major reason why phase-shift ssb generation isn't very common in ham equipment. It is difficult to adjust properly. Later in this article, we'll discuss some principles of adjusting this type of single-sideband exciter. You'll see that correct adjustment does require a sound technical knowledge of how the circuit and the system works, and that it takes more time and test equipment than the less critical filter method.

There are other reasons for the lack of interest in phase-shift ssb in commercial ham equipment. In hf gear, cost is usually a little higher, because at least one extra circuit is required. However, a vhf transmitter would require extra stages and extra crystals anyway, for frequency translation; in the long run, the phase-shift method may actually be cheaper in vhf and uhf equipment.

For another thing, careful attention must be paid to the design of rf and af circuits, because the quality of suppression depends on precise phase-shifting. This is particularly true in the voice-signal bands where a range of frequencies from 100 to 3000 Hz must be handled exactly the same—phase-shifted the same amount. Since phase-shift networks are inherently frequency-sensitive, only careful design can keep them flat over so broad a ratio of low-to-high audio frequencies.

Furthermore, speech amplifier circuits must be very flat in response, and can introduce no phase shift of their own. If higher frequencies are phase-shifted more or less than those at the low end of the range, poor sideband suppression results. Even amplitude attenuation at certain frequencies may create balance problems that could upset the complete phase cancellation upon which sideband suppression depends.

fig. 1. Block diagram of a phase-shift exciter for upper-sideband operation on 14.25 MHz.



principles of phase-shift ssb

To clarify how a single-sideband signal is generated by phase shift, most explanations incorporate vectors. I propose to explain the operation without using mathematics, but to do so I'll have to break up the explanation into two parts. First, you'll get an overall view of the system, using the block diagram in **fig. 1**, and then a more detailed analysis of just how phase shift cancels one of the sidebands.

For the sake of this explanation, assume the ssb generator is to furnish a single-sideband signal at 14.25 MHz. A crystal oscillator initiates the 14.25-MHz carrier signal. In the simplest phase-shift system, the rf signal is sent in two directions. In one direction, the signal is applied directly to the rf input of a balanced modulator, which is labeled A. The other portion of the 14.25-MHz signal goes through a phase-shift network which retards its phase by exactly 90". This phase-shifted rf signal is then applied to the second balanced modulator, which is modulator B.

Meanwhile, the voice signal has been processed by the speech amplifier. It is also split, with one portion being fed directly to modulator A and the other being phase-shifted by 90° and fed to modulator B. In some transmitters, this 90° phase separation is accomplished by shifting one af signal forward by 45° and the other backward by 45". The important thing is that the two voice signals applied to the separate balanced modulators be exactly 90" apart.

The way balanced modulators work was explained in an earlier article.[1] Therefore, you should already understand how a double-sideband suppressed-carrier (dsbsc) signal develops in each balanced modulator. Actually, for our purpose, it is sufficient to know that the output from balanced modulator A consists of a double-sideband suppressed-carrier signal. The important difference between it and the output of an ordinary balanced modulator is that the lower sideband is effectively 90" out of phase with the upper sideband. An interesting point—and one to remember.

From modulator B, the out-of-phase sideband relationship also exists. The upper sideband from B is in the same phase as the upper sideband from modulator A. However, modulator B's lower sideband is effectively 90" out of phase with its upper sideband, but in the opposite direction from that of the lower sideband from modulator A.

With one lower sideband lagging by 90°, and the other leading by 90°, the two lower sidebands are obviously 180" out of phase with each other. Meanwhile, both upper-sideband signals are in phase with each other, and reinforce the upper sideband. When the outputs of the two balanced modulators are combined, only the upper sideband is produced; the two lower sidebands cancel each other. Of course, the 14.25-MHz carrier signals were canceled in each balanced modulator, by normal balanced-modulator action.

Now, summarizing this generalized description of the effective action of phase-shift single-sideband generation depicted in fig. 1: the rf signal is split and one part of it shifted in phase by 90"; the two signals are applied to separate balanced modulators. The af or voice signal is also split into two parts that are separated by 90' and fed to the two balanced modulators. The output of one modulator has one of its sidebands leading by 90"; the same sideband from the other modulator is lagging by 90". As a result, when the outputs are mixed, that sideband is canceled. The other remains, and is the single-sideband output of the generator. The carriers are eliminated by regular balanced-modulator operation.

The system shown in fig. 1 generates an upper sideband. All that is necessary to make it generate a lower sideband is to reverse the phase of the 14.25-MHz carrier fed to modulator B. That is, instead of it lagging the signal of modulator A by 90°, it is made to lead the signal by 90". When the signals come out of the two modulators and are mixed, the lower sideband is reinforced, and the two upper sidebands are 180° out of phase and therefore cancel.

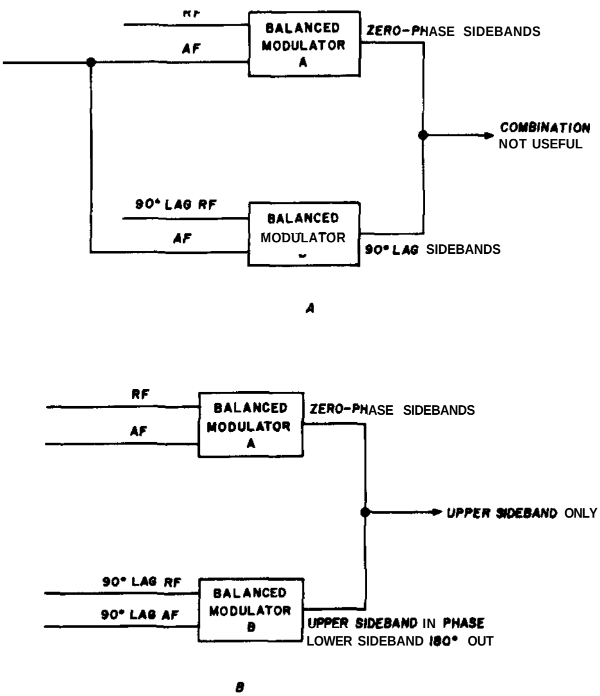
how the sidebands cancel

Now for the detailed analysis I promised, which can help explain further how just one sideband is developed in the system of fig. 1. The generalized version you've already read gives you some idea how, but the description isn't complete. Certain points were overlooked in the interest of simplification.

First of all, there is something to remember that will eliminate confusion about the phase relationships in these circuits: The phase relationship between the voice signal and the rf signal is irrelevant; it has no bearing whatever on what comes out of either modulator or out of the whole exciter system. What is important is the phase relationship between the two audio components when they are fed to the two balanced modulators, and the phase relationship between the two rf signals when they are likewise fed to the modulators. If you make the mistake of trying to visualize any phase relationship between the voice signal and the rf, the whole concept becomes very confusing. Even when the system is explained by vectors, a vector diagram can represent only one particular instant in time; at that instant, the audio signal could be at any particular point of any excursion and so could the rf signal. Consequently, I repeat: ignore any phase relationship between the rf signal and voice signal; there just isn't any that matters.

However, to understand the cancelation of sidebands, you do have to consider the phase relationship of the combined rf-af signal. In other words, even though we don't care about the exact phase of the af in relation to the rf, once they are mixed together, the phase relationships in the signals that result are the basis for the entire explanation. Once having entered the balanced modulators, the rf and af form double-sideband signals. From that point on, they must be considered as one; they are not rf and af any longer, but are double-sideband suppressed-carrier signals.

fig. 2. Principle of single-sideband generation using phase shifts. With a 90° phase shift to only the rf signal, the combinations from the balanced modulators does not produce a ssb signal. A 90° phase lag in both the rf and audio signals produces an upper-sideband, suppressed-carrier signal (B).



Bearing in mind these precautions, you can now consider first what happens in balanced modulator A (fig. 2A). An rf signal is fed in, and so is an af (voice) signal. What comes out of balanced modulator A is a double-sideband carrier-less signal. In all phase relationships there must be some reference point. You can consider the rf going into modulator A as being the zero-phase reference for the input rf, and the af going into modulator A as being the zero-phase for the input af. It follows, then, that you can also consider the sidebands that come out of balanced modulator A as being at zero phase. Remember that this last reference is necessary because you are going to consider this signal in relationship to the one coming from modulator B.

Next, consider what goes on inside balanced modulator B (still following the action in fig. 2B). Both signals going into modulator B lag those in modulator A by 90". Remember this 90° lag applies only as the rf and af signals go in. Because the rf and af signals fed into modulator B have this special phase relationship to the signals fed into modulator A, the sidebands that come out of B have a specific phase relationship to those coming out of A.

Without resorting to vector diagrams, what happens in the two balanced modulators can best be explained in terms of results. Considering the input rf as a temporary phase reference, the output of any balanced modulator is a double-sideband signal in which the phase of the sidebands depends on the audio signal. This may be a little hard to understand in view of the mention already made that the phase relationship between input rf and input af is irrelevant. However, keep in mind that you are dealing with two balanced modulators, and it is the phase relationships between signals. in the two that are important.

Start by analyzing operation under the signal conditions in fig. 2A. At any given instant, balanced modulator A produces two sidebands, an upper and a lower, that you can think of as bearing some arbitrary relationship to the input carrier signal. At the same instant, balanced modulator B produces two sidebands, an upper and a lower, that bear the same relationship to its input carrier signal. However, the input carrier signal at B lags that at A by 90°, and therefore the upper and lower sidebands from B also lag the upper and lower sidebands from A by 90".

That's simple enough, but it doesn't accomplish anything useful. The important action takes place when another change is made in the input signal conditions, and fig. 2B shows that change. A 90° phase lag is introduced into the af signal going to modulator B. The effect on the sideband signals from modulator B is to swing them even further out of phase with the sidebands from A.

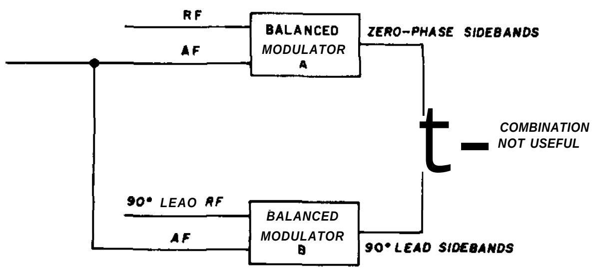
This new phase relationship between the output sidebands is one that is useful. The phase of the upper sideband from modulator B is shifted 90" backward from the position it held under fig. 2A conditions, and it is back in phase with the upper sideband from modulator A. The phase of the lower sideband at the same time is shifted 90° forward from its position described earlier, and it is now 180° out of phase with the sideband from modulator A. The result: a single-sideband output, which is what we want. The lower sideband is canceled, and only the upper appears in the combined output of the two balanced modulators.

The manner in which a single lower sideband is generated is diagramed in fig. 3. Starting with the signal conditions in fig. 3A, the action follows the lines already explained for producing an upper sideband. The chief difference is that the rf signal applied to modulator B leads that in modulator A by 90". The output, then, so long as the af signal applied to both modulators is the same, consists of two double-sideband signals in which the one from B leads that from A by 90". Again, this serves no useful purpose.

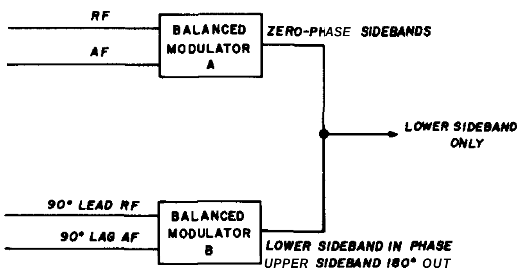
When the af signal fed to modulator B is shifted to lag that fed to A by 90°, a whole new relationship is set up in the sidebands coming out of modulator B. Fig. 3B depicts this set of signal conditions. The upper and lower sidebands from B are affected the same way they were in fig. 2B. The phase of the upper sideband is shifted backward 90" from the position it held under fig. 3A conditions, and the lower sideband is shifted forward 90°. Remember, however, that the rf signal reaching B now leads the rf signal in A instead of lagging. Shifting the modulator-B lower-sideband phase forward puts it in phase with the modulator-A lower sideband, and they reinforce each other. The result: only the lower sideband appears in the combined output of the two balanced modulators.

In summary of the overall effect, balanced modulator B shifts one sideband or the other 180" from the reference or zero-phase sidebands coming from balanced modulator A. Which sideband comes out 180° out of phase depends on whether modulator B's input rf signal leads or lags by 90".

fig. 3. Generating a lower-sideband ssb signal with the phase-shift system. The method shown in A does not provide a useful combination; the audio end rf signals must be phase shifted as shown in B to obtain a lower-sideband suppressed-carrier signal.



A



B

sideband switching

Examine that statement a little further. In the one case, the rf signal in modulator B lags by 90"; in the other, it leads by 90". The two conditions are 180" apart. Switching the rf signal in modulator B by 180°, which is the same as exactly reversing its phase, also exactly reverses which sideband is canceled in the combined output.

The two conditions suggest a means by which a phase-shift ssb transmitter can be switched from upper-sideband to lower-sideband operation. Rather sophisticated rf phase-shift networks are switched into and out of the circuit leading to modulator B, to institute whichever phase shift is needed for the mode desired.

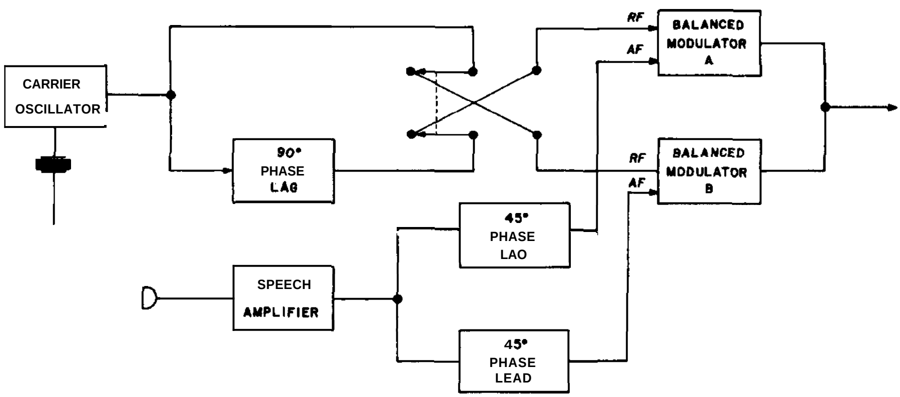
Another interesting method of switching sidebands is shown in fig. 4. Keep in mind that phase is always a matter of reference. If the rf signal at modulator B lags that at modulator A, it is just as true to say that the rf signal at modulator A leads that at modulator B. It's just another way of stating the same condition. In fig. 4, the 90" lag network can be switched from the input of modulator B to the input of modulator A. When the phase-shift circuit is between the rf source and modulator B, the system produces an upper sideband, as has already been described. Move the circuit between the rf source and modulator A, and conditions are right for producing only the lower sideband. That occurs when the rf signal in modulator B leads that in modulator A, and it doesn't matter if the relationship is caused by lagging rf signal to A.

Fig. 4 also shows the alternative way mentioned earlier of handling the audio phase shift. Instead of one network that must shift the broad range of voice frequencies the entire 90°, two are used—each shifting one portion of the signal 45". One causes the signal to lead 45°, the other makes it lag 45"; the resulting two outputs are 90° apart. A flatter response characteristic is possible in the networks when this method of phase-shifting is used.

adjusting phase-shift ssb

Proper adjustment of this kind of ssb generator is a matter of care and the right test equipment. There are several techniques, but the simplest—using an oscilloscope—will be covered here; it is adequate for all except the most exacting conditions.

fig. 4. Switching sidebands in a phase-shift ssb generator. With the switch in the position shown, the output is a lower-sideband, suppressed carrier signal. For upper sideband, the rf phase-shift network is switched from balanced modulator A to balanced modulator B. The 45° audio phase-shift networks are discussed in the text.



Two things are to be accomplished by alignment. First, the two balanced modulators must be adjusted individually for zero carrier output, with no voice signals applied. Second, the overall stage must be adjusted for maximum suppression of the unwanted sideband.

Obtaining the first requirement is the easiest. An rf indicator near the output of the exciter will let you know when each balanced modulator is adjusted for minimum carrier output. The controls to adjust are the "rf balance" or "carrier balance" adjustments; since there are two balanced modulators, there will probably be two such controls. Each balanced modulator may also have a "phase" adjustment—often a trimmer. Adjust each of them for minimum carrier output. You may have to juggle a little between the "balance" and "phase" adjustments of each modulator to find which position results in the least carrier output.

For the sideband-suppression adjustments, you must connect the scope to indicate the output signal from the ssb exciter. A pickup loop coupled loosely to the exciter and fed directly to the vertical deflection plates of the scope will usually work out fine. Tighten or loosen the coupling of the loop to get a usable display on the scope screen. The internal sweep of the scope should be set for any convenient submultiple of the test tone that is to be used to modulate the ssb exciter. If 1000 Hz is to be used, set the scope for 50 or 100 Hz, so you can see several ripples if any ripples exist.

The af test signal is from an audio generator, fed to the speech input connection of the exciter. Be sure to use as little signal as will modulate the exciter; too much will create distortion that can make adjustments misleading. The frequency to use depends on the phase-shift network used in the exciter, but 1000 or 1200 Hz will work out about right in any case.

What you are trying for is a pure sideband output, indicated on the scope by a smooth "bar" of rf, an inch or two high on the screen (its exact height is determined by how tightly the pickup loop is coupled to the exciter). The sign of incorrect sideband suppression is a ripple along the top and bottom of the rf envelope or "bar." The deeper the ripple, the poorer the sideband suppression. The ripple is caused by beat components generated when two sidebands mix with each other. Adjustments, therefore, are made to reduce any ripple that is visible.

There are certain requisites in the circuit for maximum suppression of the unwanted sideband. One is that the rf signals fed to the two balanced modulators be precisely 90° apart—no more, no less. The signals should be approximately equal in amplitude, but the 90° phase separation **must** be exact. Therefore, an "rf phase" control is provided, sometimes two. (The two doesn't mean one for each modulator, but represents a particular type of phase-shift network that may be used.)

A second important requisite for perfect sideband suppression is that the af signals fed to both balanced modulators be exactly the same amplitude. An "af balance" control takes care of that. In a few exciters, there is also an "af phase" control; but phase isn't as critical in the af signals as the amplitude match, so most transmitters keep af phase balanced by fixed resistors.

Setting these two (or three) controls is all there is to aligning the exciter for proper sideband suppression. You watch the pattern on the scope, and adjust first the rf phase and then the af balance until there is as little ripple as possible in the pattern. Once the adjustments have been made, it is well to go back and check the carrier-balance controls in each balanced modulator. Touch them up, then touch up the sideband-suppression adjustments.

Once ripple has been reduced to a minimum, switch to the other sideband. Best adjustment for one may not be best for the other. If there is any ripple on the other sideband, it probably can be reduced by readjusting the rf phase control. You may have to leave it set halfway between the first setting and the last, as a compromise to both sidebands.

With careful adjustment, using proper techniques and equipment, phase-shift single-side-band generation can be almost as effective as filter-type. The scope method of adjustment just described is good enough for a dependable 30 dB of sideband suppression. More than that is possible with more sophisticated adjustment procedures.

In the vhf and uhf bands, you may see phase-shift sideband generators used more often—since ssb has caught on. Better equipment and better techniques will probably make this equipment more popular. What you've learned here will help you make optimum use of this truly convenient principle.

reference

1. F. H. Belt; "Generating SSB Signals with Suppressed Carriers," ham *radio,* May, 1968, p. 24.

ham radio

GOING STRONG!

The latest evolutionary development of the Classic Model 350

The new SWAN 350-C
5 band, 520 Watt
SSB-AM-CW TRANSCEIVER

The introduction of the new model **350-C** at no increase in price over the popular *350* is a real triumph of value engineering.

Power rating of the **350-C** is the same as the Deluxe Model *500-C,* which is *520* watts P.E.P. input on single sideband, *360* watts *CW* input, *125* watts AM input. A pair of rugged, blast-rated **6LQ6** tubes handle this input with ease. Selectivity is provided by the same superb crystal lattice filter used in the **500-C,** with skirt selectivity and ultimate rejection superior to any other filter being used in amateur equipment today. Audio quality has the same degree of fidelity which Swan has stressed from the very first single band transceivers.

The *350-C* is designed to operate automatically on the normally used sideband with extended frequency coverage of all five bands, *10* through *80* meters. The high frequency crystal lattice filter is common to transmit and receive circuits. Bandwidth is 2.7 *KC.* Audio **bandbass** is essentially flat from *300* to *3000* cycles. Sideband suppression is greater than *50* db; carrier suppression is greater than *60* db. Grid block *CW* Keying is provided with offset frequency. The VFO is transistorized, temperature and voltage stabilized. Receiver sensitivity is better than .5 μv for *10* db signal-plus noise to noise ratio. Velvet smooth **dual**-ratio tuning is featured, as in all Swan Transceivers. Basically the difference between the *350-C* and the *500-C* is in the deletion of optional features which are not essential to communication. These include such things as crystal calibration, sideband selector, *CW* sidetone, automatic noise limiter, automatic level control, etc. For the operator who desires these features, we are proud to recommend the deluxe model *500-C.* However for powerful and reliable **communica**tions without all extras, we now offer the new *350-C,* and we are confident that you will rate it a truly exceptional value.

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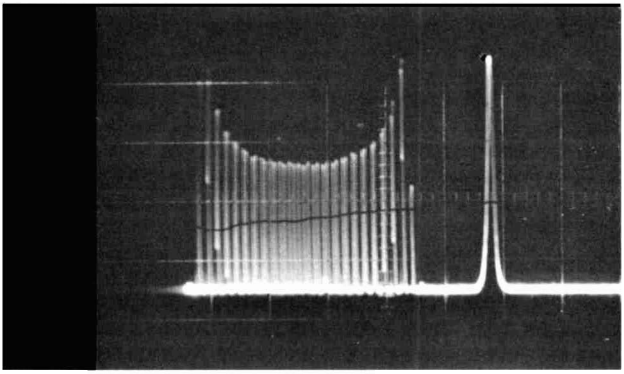


This is the way the output from a sweep generator looks on a spectrum analyzer. The "pip" on the right-hand side is a marker signal.

instrumentation
and the ham

You may not have access to some of the instruments described in this article, but it should give you an idea of some of the things that can be accomplished with modern equipment

M. J. Goldstein, VE3GFN, 22 Kingswood Road, Toronto, 18 Ontario, [redacted]a

It has been my lot in life, over the past eight years, to be closely associated with much of the better electronic measuring equipment — mostly with research and development labs and lately in the educational field. Needless to say, having access to such equipment has been a great benefit to my ham career since homebrewing is my greatest passion. It occurred to me, after several months of instructing engineering students in the use of instruments, that if graduating engineers find instrumentation a mystery, certainly Joe Ham could be in the same boat.

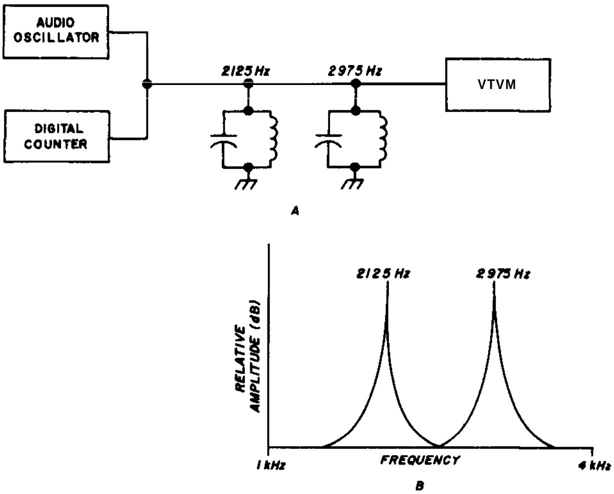
To the majority of hams who have no access to fine instruments, a discussion of this nature may provide an insight into a side of the hobby not previously encountered. Most dedicated homebrew artists seem to come up with any instruments they need, but you may find a new attack to some old problem in this article.

I would like to mention that the objective here is not to teach you how to use any instrument; only careful study of the instrument with the manual can accomplish this. Rather, it is to demonstrate what can be accomplished in a properly-equipped lab, and how measurements familiar to the ham world are obtained. While some of the equipment mentioned may appear to dwell in the science-fiction region, please be assured that all of it exists.

filters

A common problem in ham gear is obtaining the response characteristics of filters, whether they're used in exciters, receivers or what-have-you. Let's examine the several ways of measuring this response.

fig. 1. Test setup for measuring the response of a typical RTTY filter is shown in A; e plot of the measured characteristic is shown in B.



Suppose you have an RTTY converter which includes a filter that rejects all but 2125 and 2975 Hz. A popular system is to use two parallel hi-Q tuned circuits employing toroidal coils. An equipment set-up that can be used to measure the response of such a filter is shown in **fig. 1**.

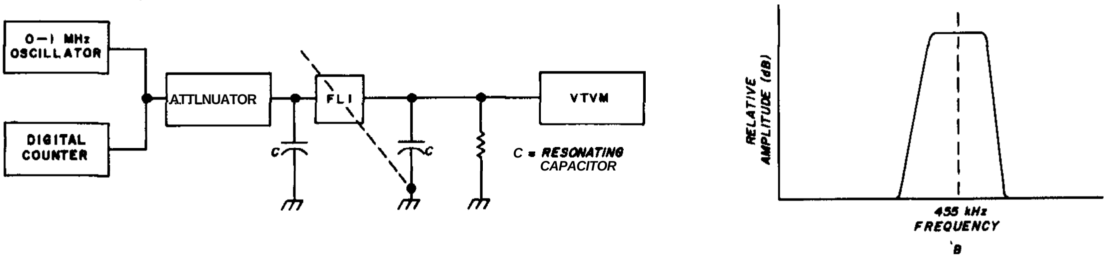
The digital counter indicates the oscillator frequency directly in frequency units with great accuracy, constantly sampling the frequency and indicating any change. The audio VTVM has a decibel scale and will indicate audio levels down to sixty decibels below one volt. The frequency limits of the VTVM depend on its quality, but most will respond accurately to the frequencies we are discussing here. Most response curves are plotted as amplitude levels in decibels, and the convenience of having an indicating device calibrated in these units is apparent. If such a VTVM is not available, a standard VTVM can be used; the level changes can be converted to decibel changes by using the nomograph in **fig.** 3. This chart is based on the formula:

$$dB = 20 \log_{10} \frac{\text{large voltage}}{\text{small voltage}}$$

The measurement procedure is as follows: set the oscillator frequency well below the lower filter frequency, and set the output level at one volt before the filter is connected in the circuit. Connect the filter. Move the frequency up the spectrum in 100-Hz steps, noting the relative level at the filter output at each frequency. Take readings at smaller frequency increments as the critical frequencies of the filter are approached to obtain an accurate response. Continue this procedure until you're well above the upper filter frequency. The correct form of the response characteristic, as it would appear when plotted on semi-log graph paper, is illustrated in **fig. 1**.

On the other hand, assume we have a mechanical filter with a center frequency of 455 kHz and bandwidth of 3 kHz. Another method of measuring a filter response is shown in **fig. 2**. The only precautions to be observed for this measurement are to shield the filter input from the output, and ensure

fig. 2. Another way of measuring the frequency response of a filter. The plotted output is shown in B.



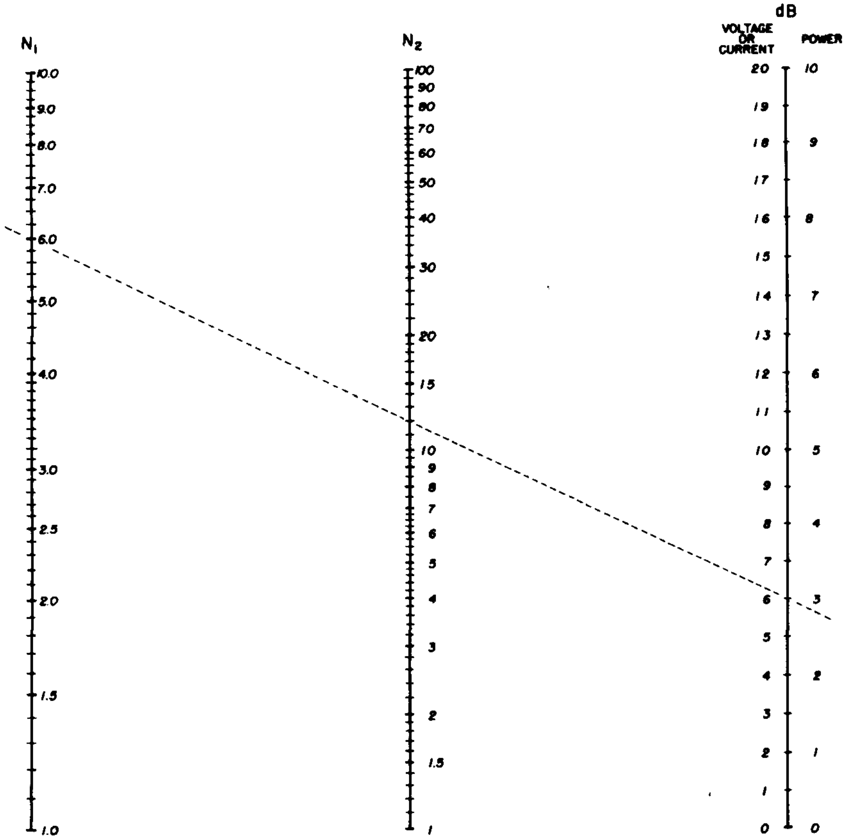


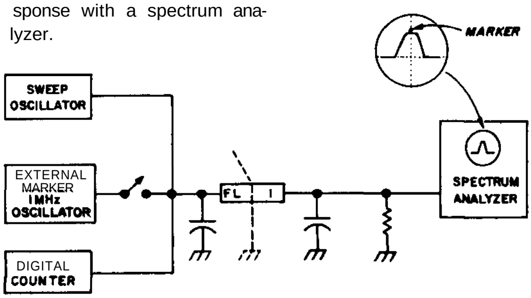
fig. 3. Nomograph for converting **voltage, current** and power ratios into dB, either gain or loss. The larger number is entered on the **N2** scale; the **smaller** number on **N1.** A straight line connecting these two points is extended through the dB scale—the point of intersection is the answer in dB. **If** for example, the voltage into an amplifier is 6 volts and the output is **12** volts, what is the gain of the stage? The answer is 6 dB as indicated by the plotted line. Or, if you put **12** watts into one end of a transmission line and measure 6 watts out, what is the loss of the line? The answer **is** 3 **dB as** indicated on the poww side of the dB scale.

the VTVM will respond accurately to 500 kHz.

The procedure is as follows: adjust the oscillator output to exactly five volts (before the attenuator). Set the attenuator to 0 dB. Sweep the oscillator frequency around 455 kHz until the VTVM indicates a peak. With this filter, the peak will be about 2-kHz wide, so adjust the frequency for the middle of the peak. The peak level should be some decibels below the five volt level, depending on the filter, Insertion loss is the decrease in gain that the filter will cause when put in a circuit. The oscillator output is set to a high level so you can measure very low attenuation levels at the filter output without running out of VTVM sensitivity.

Once the peak frequency has been adjusted, turn the oscillator output level to minimum. Adjust the attenuator for maximum attenuation (should be at least 60 dB). Set the VTVM to the most sensitive scale (–60 dB), and adjust the oscillator output level until

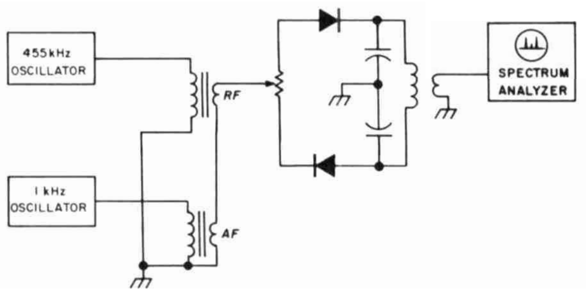
fig. 4. Measuring filter response with a spectrum analyzer.



the VTVM indicates "0 dB" on this scale. Remove one-half dB of attenuation, and adjust the oscillator frequency below the peak frequency until the VTVM again indicates 0 dB. Note the frequency. Continue in half-dB steps until 3-dB of attenuation have been removed. Proceed in 1-dB steps until 10 dB of attenuation have been removed. Continue in 5- or 10-dB steps until the remaining attenuation has been removed.

Now, set the oscillator to the peak frequency as before, and repeat the procedure on the high side of the peak frequency.

fig. 5. Checking the output of a sideband generator with a spectrum analyzer.



The same filter response can be examined, without plotting it by hand, by using some of the latest oscilloscope techniques. You can use a spectrum analyzer, which may take the form of a complete unit, or perhaps as a plug-in unit for a standard scope frame. The hookup is illustrated in fig. 4. The sweep oscillator is an oscillator which sweeps back and forth over a pre-adjusted band of frequencies. The horizontal axis of the spectrum analyzer can be adjusted to display the same band of frequencies; the vertical axis is calibrated in volts. Once the filter response is set up for display on the analyzer, a separate oscillator can be connected to the sweep oscillator as a "marker" oscillator. A pip will appear on the displayed response when the marker oscillator frequency is within the frequency band under display. The marker oscillator can be switched to drive the frequency counter so a method of accurately calibrating the horizontal axis of the analyzer is available.

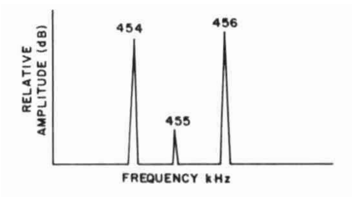
sideband generators

The major problems of designing a filter-type sideband generator are balancing the balanced modulator (a problem also encountered in phasing exciters) and setting the frequencies of the carrier crystals at the proper point on the response curve of the filter.

Fig. 5 shows a balanced modulator driving the spectrum analyzer. The 455-kHz oscillator simulates one of the two carrier crystals you will ultimately use. The 1-kHz oscillator simulates the microphone input signal.

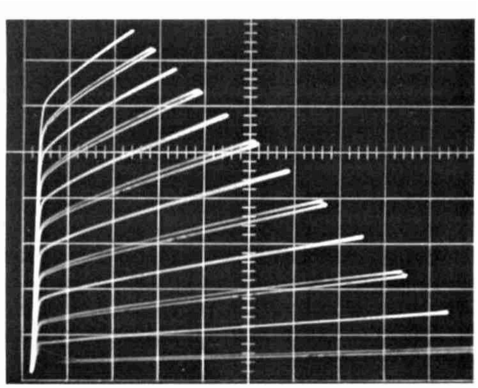
Theory states that the output of the balanced modulator should appear as in fig. 6.

fig. 6. Spectrum analyzer display of the output from the test setup shown in fig. 5.



This is the response which should appear on the face of the spectrum analyzer. The balanced modulator is properly adjusted when the carrier pip at 455 kHz is at minimum amplitude. You made the necessary measurements to set the frequency when you accurately measured filter response. The carrier crystals should be 1500 Hz above and below the "peak" frequency of the filter.

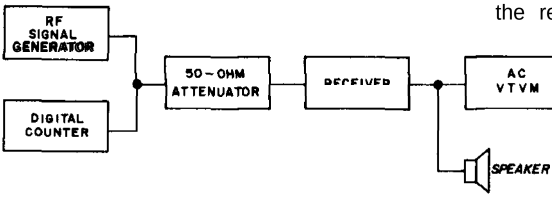
Characteristic curves of a 2N410 transistor as displayed on a curve tracer.



transmitters

All of the adjustments and measurements to be made on a transmitter can be made with a wide-band oscilloscope, several of which are now on the market. You can examine the sine-wave output of your two-meter transmitter, adjust your keying characteristic's rise and fall times within nanoseconds, and trace non-linearities in ssb rigs from the audio input to the 30-MHz output. You can measure the frequency of parasitics, and neutralize your final to within an inch of its life (does anyone neutralize anymore?). You can adjust each coupling circuit for absolute maximum coupling or optimum coupling, which is more realistic, and measure output power in PEP to the nearest milliwatt.

fig. 7. Measuring the frequency drift of a communications receiver.



receivers

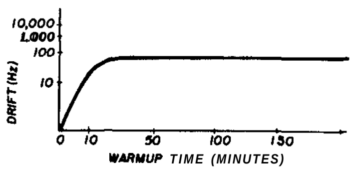
There are a multitude of measurements that can be made on a receiver to determine whether or not optimum performance is being obtained. Frequency stability is one. The simplest method of evaluating the stability of a receiver is to measure the stability of the variable oscillator over a period of time—preferably the warmup period of the receiver. This is done by connecting a digital frequency counter to the variable oscillator output and measuring the frequency difference in the time between turn-on and, say, two hours of warmup. The frequency difference can be read to the nearest cycle. A more realistic approach where the drift of the entire receiving system is involved is illustrated in **fig.** 7.

Since the ultimate purpose of a receiver is to make measurements by ear, it's quite "cricket" to use the speaker as an output device. The purist can use the VTVM to make absolute measurements of audio level. The attenuator is provided to decrease the relatively high output level of most signals to a level suitable for making measurements on a sensitive receiver.

Allow the instruments to warm up for at least several hours. The BFO on the receiver is turned on; the receiver is tuned to the frequency of the generator, then turned on. It is adjusted immediately for "zero-beat," as indicated by minimum audio level in speaker and VTVM. Note the indicated frequency of the signal generator. At time increments of ten minutes, re-adjust the generator for zero-beat, noting the new frequency each time, until all indications of drift cease. This drift characteristic can be plotted on semi-log graph paper, as shown in fig. *8.*

The mechanical stability of the receiver can be examined using the same setup. Adjust the system for zero-beat as before. Give the receiver a hearty thump, and measure any frequency change. Be sure that the frequency controls on the receiver are not moved while you're thumping it.

fig. 8. Typical frequency-drift characteristics of a communications receiver using the test setup shown in fig. 7.



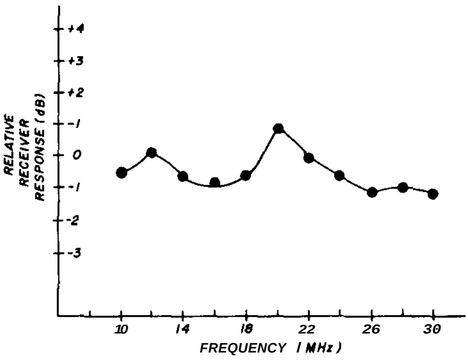
receiver tracking

The tracking of the receiver's tuned circuits can be easily checked over any desired frequency range by using the test setup of **fig.** 7. The VTVM is not used, and an audio power meter should be connected across the speaker terminals and set to the correct impedance. Adjust the signal generator to the lowest frequency in the desired range. Peak all tuned-circuit controls. Adjust the audio gain and BFO control for a convenient indication on the power meter.

From this point on, adjust only the rf tuned circuits. Adjust the frequency of the generator in convenient steps through the range of the receiver, peaking the tuned circuits each time for maximum indication on the power meter. Note the frequency and output power level each time the frequency is changed. The tracking characteristic can be plotted on linear graph paper as shown in **fig. 9.** Perfect tracking would yield a straight horizontal line on the plot.

If the tracking characteristic is undesirably shaped, it can be corrected by readjusting the receiver front-end while the receiver is connected as shown in fig. ***10.*** The time-mark generator is a device commonly used for calibrating oscilloscope time-bases. It provides marker pulses of short duration at selectable time intervals. With one micro-

fig. 9. Plot of raceiver tracking characteristics over the range from 10 to 30 MHz.



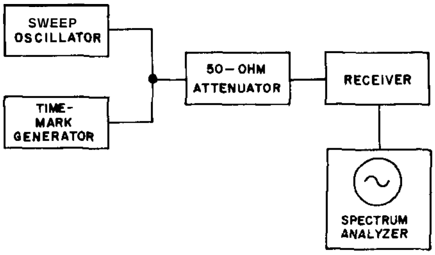
second markers, you have 1-MHz markers, since frequency = 1/time. Therefore, the time-mark generator may be used to provide frequency calibrations across the face of the spectrum analyzer. The spectrum analyzer is connected across the output of the last mixer, and provides a constant display of the tracking characteristic, making it very easy to observe the overall effect of any tuned-circuit adjustments.

adjusted balanced mixers

Many of the latest receivers use balanced mixers. These must be balanced for ac and dc to optimize their performance. This is most easily done with an oscilloscope having a differential input.

For a dc balance, the two scope inputs are first grounded, the inputs switched to "dc,"

fig. 10. Using a spectrum analyzer to adjust the tracking of a communications receiver.



and a zero reference obtained. The scope inputs are then connected to the two mixer inputs which require balancing. The oscilloscope will indicate any difference in dc potential between the two mixer inputs. The dc balance control on the mixer is adjusted for minimum dc potential difference. The ac balance is obtained by switching the scope inputs to ac, and repeating the procedure using the ac balance adjustments to minimize the amplitude of the waveform displayed on the oscilloscope. This waveform will most likely be a sine wave with amplitude proportional to the degree of ac unbalance of the mixer.

summary

There are a host of other measurements to be made on ham gear, but the ones mentioned here seem to be the most interesting. One of the nice ways to keep measurement records of oscilloscope traces is by means of Polaroid photographs, made with special cameras designed for that purpose. Some sample pictures have been included to show the versatility of this technique.

From the loop-and-bulb to the spectrum analyzer is quite a jump, but progress is what makes this old hobby of ours the fascination it is. Now where did I put that coffee-can grid-dipper . . .

ham radio

which way does current flow?

Positive to negative, negative to positive, or does it really make any difference?

Jim Ashe, XH, 03 Dryde oad, Ithaca, New York 14850

Strongly held expert opinion that current flows from plus to minus sharply opposes strongly held expert opinion that current flows from minus to plus. This situation has existed for many years. It complicates discussions between experienced workers in electronics, and has a specially confusing effect upon beginners and students.

Why have these apparently contradictory opinions persisted? The answer lies in the history of electronics and physics. As with many differences that are hard to settle, this is a matter of words. Both views work if applied **consistently** with slightly different meanings of key terms; and there is a good chance that both views are wrong!

The key terms are **current** and **charge.** Correct usage, having roots in history and physics, is that current is a rate of flow of charge. The latter usage, found in many electronics texts, is that current flows from plus to minus (or minus to plus) without any mention of charge. The plus-minusers point to history to support their view, and the minus-plussers point to electrons. If you can stand that word once more, both have missed the point.

the answer is in physics

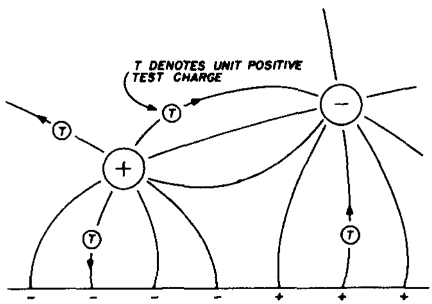
It is surprising to look back in history and discover a time, not so very long ago, when there was no field of electronics. We must look back much farther to find a time when there was no physics, but maybe the line can be drawn at Galileo. He introduced the idea, regarded as revolutionary in his time, of comparing the results of careful thinking with the results of careful experiments and changing the thinking if the experiments didn't agree. Previous techique was to ignore the experiment — if somebody happened to perform it.

As the science of physics developed, workers noticed and studied electrical effects, and came to understand something of what was actually happening. Ben Franklin introduced the terms positive and negative in 1747, and in the 1800's the first spin-offs from physics began to appear: telegraphy, power transmission and lighting. As work continued, new spin-offs appeared; some of them became the electronics we are familiar with today.

The close association between electronics and physics continues, and all of the electrical quantities in electronics are the same as or based upon the units of measurement used in physics. A physicist can provide the expert opinion required to understand the "current" problem.

When asked which way a current flows, he is likely to say, "Current is a scalar quantity. It is a magnitude, a meter reading. It has no direction. And since a current is a flow of charge, the common expression, 'flow of

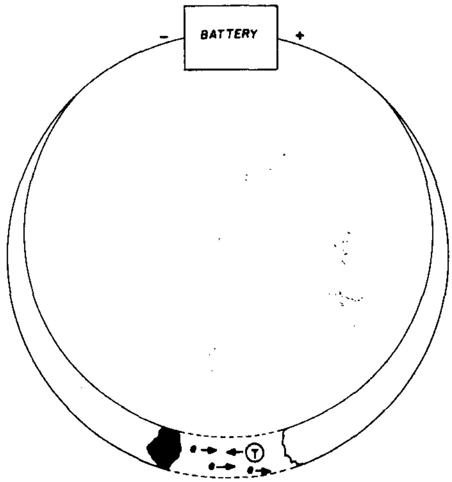
fig. 1. Two fixed charges, and two induced charges. Description of this arrangement is greatly simplified by the physics convention that the electric field st any point has the strength and direction experienced by a 1 coulomb positive test charge placed at that point in the field.



current,' should be avoided. That means, literally, 'flow of flow of charge,' not very good English. But 'flow of charge' is perfectly correct usage, and not at all ambiguous as to direction. Charge flows from plus to minus."

Well, that is a bundle of news and no mistake! Let's take this a bit at a time. Why does charge flow from plus to minus? Because everybody in physics agrees that it does: a convenience worked out before 1800 and used ever since. It is based upon a standard procedure for investigating and describing electrostatic fields. It is the same kind of notion as the modern equivalent circuit, Thevenin generator or Maxwell loop current:

fig. 2. Placing the positive test charge in a wire, we find it wants to move toward the negative battery terminal, against the flow of electrons.



it is a conceptual handle.

Careful tests by Coulomb in 1785 clarified the importance of having a basic quantity of charge, and once the notions of quantity and direction were worked out, people could start talking about electricity in a meaningful way. The terminology and concepts were well worked out when the electron was discovered in 1895 by Thompson, and named by the Irish physicist Johnstone Stoney. It turned out that electrons were negative, so, consequently, they passed from minus to plus; but people described the situation as transferring a charge from plus to minus. The great controversy was all set up — it continues to this day.

modern conventions

It's likely that the controversy will continue indefinitely. However, for situations where careful thinking and description are required, the conventions in physics are easy to use and surprisingly like some common experience with automobiles. The basic concepts are those of **field, charge** and **flow of charge,** or current.

If there is a fixed electrical field, there must be a charge or charges somewhere that produces it. The size and location of these fixed charges is often unimportant. Few electronics experimenters are interested, for example, in the exact electrical fields and electrodes inside a transistor.

Of more concern is the amount and motion of movable charges affected by the fields. Charges are commonly specified in terms of coulombs, a very definite amount of electricity. One coulomb will plate about 1.1 milligrams of silver from a standard bath, exert a 1-newton force on an equal charge 1 meter away, fill a l-farad capacitor to 1 volt, or meet some other standard definition.

When coulombs are under motion, a charge flows, and we say there is a current. The current in amperes is the rate of flow past a point or through a surface in coulombs per second. An ampere-second is 1 coulomb, but it's probably best to leave the details to the physicists and think of coulombs as ampere-seconds.

This entire picture has a very close analogy in cars and traffic. The mile, as a unit of distance, corresponds to the coulomb, the unit of charge. The speedometer, indicating miles per hour, is like the ammeter, indicating coulombs per second. The physicist knows that a charge can travel in any direction, but is most conveniently moved along wires; the driver knows he can drive in any direction, but most conveniently along roads. The ammeter doesn't know the difference, nor does the speedometer. The similarity would be even closer if cars didn't have engines and had to roll downhill. Then we would label the hilltops plus, and the valleys minus, and call the geographical terrain a **field.**

We see the 'flow of current' term used every day, and it works. You may as well continue to use whatever you're accustomed to. But if you find yourself in a deep discussion, or if you need to be specifically clear in your thinking, try using the conventions used by the physicists.

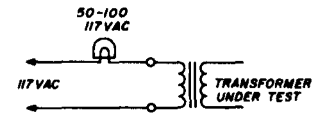
ham radio

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 phenomena 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 exper-

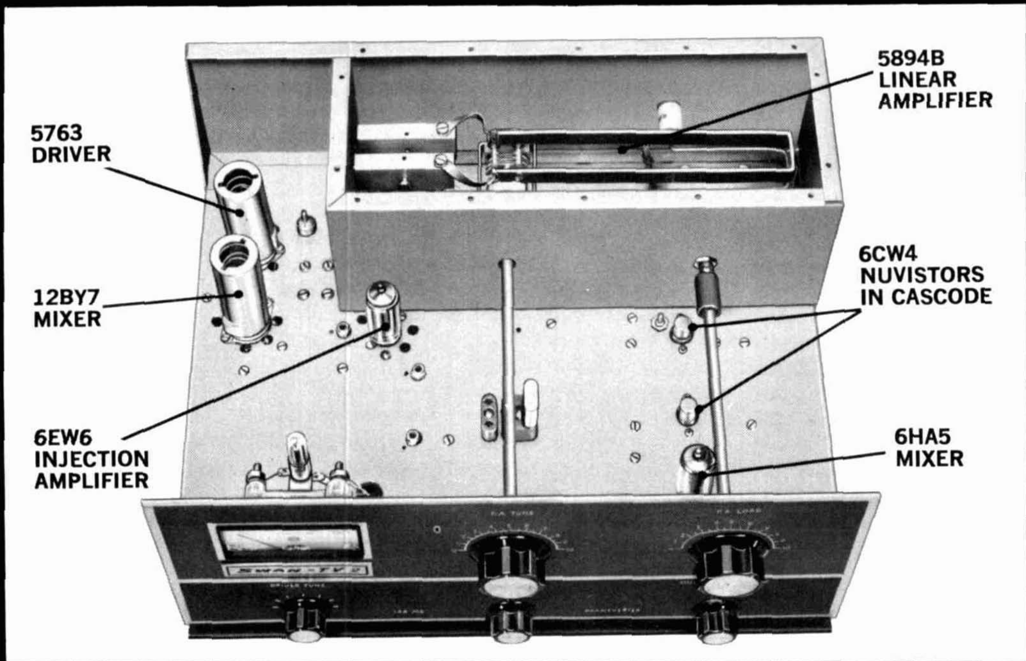


iment should only be used with low-voltage transformers. Pull the line cord while shorting, and keep your hands off while the power is on—we don't want you at the bottom of a hole. High-voltage ac has a propensity of going through insulation to your fingers, so please, put the clip leads on a long stick!

Ed Marriner, W6BLZ

2 METER SINGLE SIDEBAND

144-148 mc 240 WATTS P.E.P. INPUT



THE NEW SWAN TV-2 TRANSVERTER

A receiving and transmitting converter for the 2 meter band, designed to operate with Swan Transceivers, models 250, 350, 350-C, 400,500, and 500C.

SPECIFICATIONS:

14 mc intermediate frequency is standard. Thus, when operating the Transceiver from 14 to 14.5 mc, the Transverter functions from 144 to 144.5 mc. Additional crystals may be purchased and switched in for other portions of the 2 meter band, such as 144.5-145, and 145 to 145.5 mc. Three crystal positions are available.

Alternately, the TV-2 Transverter may be ordered for an I.F. in the 21, 28 or 50 mc bands, if desired. Of course, for use with a Swan 250 six meter transceiver, the Transverter must be ordered for 50 mc. Otherwise, the standard 14 mc I.F. is recommended since bandspread and frequency readout will then be optimum. The Transverter can easily be adjusted in the field for a different I.F. range, if required.

A 5894 B Power Amplifier provides a PEP input rating of 240 watts with voice modulation. CW input rating is 180 watts, and AM input is 75 watts.

Receiver noise figure is better than 3 db, provided by a pair of 6CW4 nuvistors in cascode.

Only a Swan Transceiver and Swan AC power supply, Model 117-XC, are required. The power supply plugs into the Transverter, and the Transverter in turn plugs into the Transceiver. Internal connections automatically reduce the power input to the Transceiver to the required level.

Tube complement: 58948 Pwr. Amp., 5763 Driver, 12BY7 Transmit Mixer, 2N706 crystal osc., 6EW6 Injection Amp., 6CW4 1st rec. amp., 6CW4 2nd rec. amp, in cascode, 6HA5 rec. mixer.

The Swan N-2 may also be operated with other transceivers when proper interconnections and voltages are provided. A separate Swan 117-XC power supply will most likely be required.

Dimensions: 13 in. wide, 5½ in. high, by 11 in. deep. Weight: 13 lbs.

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MODEL 500C 520

MODEL 117-XC
AC POWER SUPPLY . . $105

MODEL TV-2
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choosing diodes
for
power supplies

Ed Marriner, W6BLZ, 528 Colima Street, La Jolla, California 92037

Diode rectifiers have largely replaced tube rectifiers in amateur transmitters. This has reduced transformer size because the rectifier filament winding isn't needed any more. Heat inside the cabinet is also reduced. This is all very good, but you now have to decide what PIV diode you want before doing any construction.

In the old days, all you had to do was reach in the junk box for an 80, a 5R4GY or an 866. Since life is more complicated now, it's handy to keep a diode reference chart on the wall, or at least a diode source book handy.

The circuit shown in fig. 1 is a full-wave bridge. The maximum peak inverse voltage (PIV) is 1.14 x Erms (the transformer output voltage). Let's assume we have an 850-volt dc transformer (rms). How many diodes will be required in each leg of the bridge? According to the book, the inverse peak voltage will be 1.14 x 850. This gives us about 970 peak volts across each leg of the bridge. The diodes in each leg will have a peak-to-peak voltage of 970 volts impressed across them, so it is obvious that one 400-PIV type is not going to do the job. It will take several of them in series plus a safety factor. Actually, a single 1600-PIV or four 400-PIV diodes in series would be a good choice. Let's work this out a little further:

$$\text{Edc (output dc)} = \frac{\text{Erms} - 2\text{NV}}{1.11}$$

Where:

N is the number of rectifiers used in each leg of the bridge

V is the voltage drop per rectifier at 80% of the rated load current. (Use a value of 1.5 V)

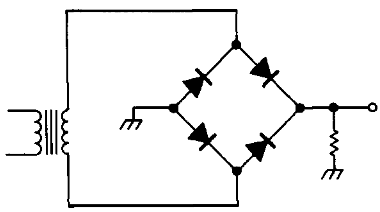
With these factors in mind, we find that the actual dc output is 755 volts.

The bridge circuit is handy because you can always use the center tap for a low-voltage supply, even though the copper losses go up. In ssb use, the transformer will handle the extra load because power is taken off in pulses.

The diodes should be mounted in an area away from any heat, and use a heatsink to keep the heat from damaging them when you're soldering them into a circuit. If nothing else, hold the leads with long-nose pliers.

Since all of the constants make such a small change, the PIV across each leg of the bridge can be used as a starter plus a safety factor for line transients of perhaps 100 volts. Diodes are quite inexpensive, and the higher ratings are not that much more expensive.

fig. 1. Full-wave bridge rectifier circuit.



The full-wave center-connected circuit in **fig. 2** is figured differently. Here, the peak voltage is 1.414 x 350 (example) or 494-V

table 1. PIV **values** for full-wave bridge and **full-wave,** center-tapped **rectifier circuits.**

transformer (Erms)	full-wave bridge (PIV per leg)	full-wave ct (PIV per leg)
200	283	566
300	424	848
400	565	1132
500	707	1414
600	848	1696
700	990	1980
800	1131	2262
900	1273	2546
1000	1410	2820

peak. If we call it *500* to round it out, then the peak inverse will be —500 plus 500 or 1000 volts stress across each leg. This tells us it would take three 400-PIV diodes plus a little safety factor for each leg—a total of six diodes.

In case you can't remember how to do it next time, **table 1** is a handy little chart which shows the peak inverse voltage across each leg of full-wave bridge and full-wave center-tapped power supply circuits.

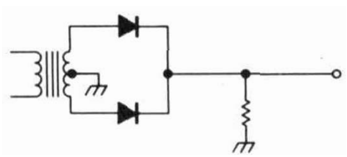


fig. **2.** Full-wave, center-tapped rectifier circuit.

Add a safety factor to the values in **table 1** when picking a diode. This could be anybody's guess since the amplitude of a transient depends on what part of the cycle it is passed through the transformer. Allow at least 100 volts on small supplies and more on high-voltage types. It's also a good idea to put a transient suppressor across the primary of the power transformer. You can use a commercial device such as a General Electric Thyrector or Sarkes Tarzian Klipvolt or a transient-suppression circuit.'

references

1. J. Fisk, W1DTY, "Ham Notebook," ham radio, July, 1968, **p. 75.**
2. "Silicon Rectifiers," Published by Sarkes Tarzian. Inc., Bloomington, Indiana, 1960.

ham radio

the dual-channel compressor

This approach to audio compression uses a stereo preamp; one channel for control, the other for compression

John J. Schultz, W2EEY, 40 Rossie Street, Mystic, Connecticut 0655

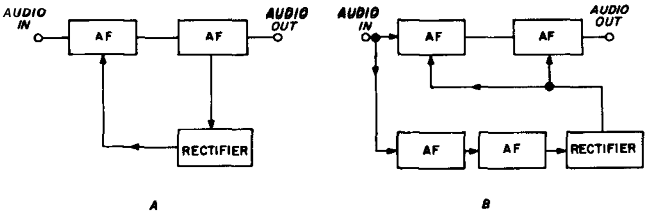
If you have any old audio equipment in a forgotten corner of the shack, you can probably turn it into a very useful piece of gear. In this article I'll show you how to turn a stereo phono preamplifier into a very effective audio compressor for use with a ssb transmitter. The preamplifier was set aside when I bought a new high-fidelity amplifier. I used it briefly as a microphone preamplifier for the transmitter and then it occurred to me that it could be made into a unique type of audio compressor. The same idea could probably be applied to any similar unit you might have. Two separate amplifiers, which by themselves might be considered of limited value, may also be used.

audio compressor circuits

The operation of a conventional audio compressor is shown in **fig.** IA. Here, the input audio signal passes through two or more stages of audio amplification; part of the audio output of the last stage is rectified to produce a dc control voltage. This control voltage is used to regulate the gain of the first audio stage. It may control a transistor switch to regulate the first-stage supply voltage, or, a diode attenuator. The methods of using the control voltage vary widely, but the principle remains the same.

The preamplifier I use contains two identical low-noise audio channels. I could have used the conventional scheme of audio compression in the second stage of either channel but, instead, I developed the method shown in **fig. 1B.** In this circuit, the inputs to the two audio channels are connected in parallel. One channel is used **only** to develop the control voltage; it regulates the gain of the other audio channel. This approach has

fig, 1. Audio compressor systems. The conventional compressor is shown in A; the dual audio amplifier system is illustrated in B.



the advantage that a large amount of control voltage can be developed. In **fig. 1A,** part of the audio output must be rectified. Also, since the gain of both stages in the audio channel are controlled, you obtain more effective compression.

The audio channels I used were matched since they were part of a stereo preamplifier. However, there doesn't appear to be any reason why two separate monaural amplifiers with different characteristics couldn't be used to build a similar unit.

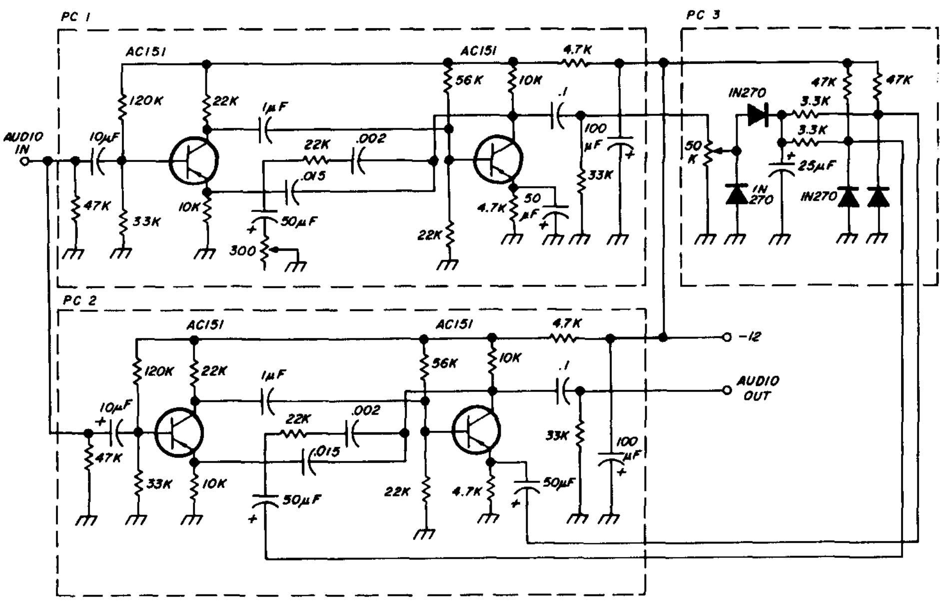
operation

The schematic of the dual-channel compressor is shown in **fig.** 2. If PC3 is forgotten for the moment, and the two leads between PC2 and PC3 are connected to ground, you can see that both PC1 and PC2 are conventional two-stage audio amplifiers. A feedback circuit from the collector of the second stage to the emitter of the first stage improves the frequency response and stability.

The only difference between amplifiers is the 300-ohm variable resistor in series with the 50-μF emitter bypass capacitor of the first stage of one amplifier for gain adjustments. The function of this potentiometer is worth noting since the same action, automatically controlled, is used with PC3 to provide compression.

The output from one channel is connected to a 50k-ohm potentiometer on PC3 which serves as a **compression level** control. The voltage frorn the potentiometer is rectified and coupled through 3.3k-ohm isolating resistors to 1N270 diodes in series with the emitter bypass capacitors of the other chan-

fig. 2. Schematic of the dual-channel compressor. The blocks show the three printed-circuit boards. The control amplifier board is on top, the controlled amplifier below.



nel. The diodes are also connected to the 12-volt supply through 47k-ohm resistors.

When low-level audio signals pass through the amplifier channels, the negative voltage from the 12-volt source forward biases the diodes. The two 50-μF emitter bypass capacitors are shorted to ground, and amplifier gain is maximum. As the signal level increases, an increasingly positive voltage is developed across the 25-μF capacitor on PC3, and the diodes are reverse biased; the emitter bypass capacitors are no longer effective, the stages of the second amplifier become increasingly degenerative, and gain and output decrease.

adjustment

Adjustment of the compressor consists mainly of varying the 50k-ohm **compression level** control for the desired amount of compression. Although there is some noise buildup during speech pauses when using this compressor, it is less noticeable than with other circuits. This is probably because the amplifiers were designed for low-noise, broad-frequency response and include generous amounts of feedback for stabilization.

construction

You can duplicate the circuit shown here, but more likely, you'll want to use existing amplifier circuits. In that case, you only have to worry about the components shown on PC3. Some variation in values may be necessary to suit individual applications. The 25-μF filter capacitor and 3.3k-ohm isolating resistors may be varied to provide different time constants for compressor action. The 47k-ohm isolating resistors determine the point where compression becomes effective. If a positive supply voltage is used, all the diodes on PC3 should be reversed as well as the 25-μF filter capacitors.

The components that are a part of PC3 can be mounted in any convenient location. In my compressor, they were mounted on a piece of Vectorboard and placed in one corner of the preamplifier. Connections from PC3 to the amplifiers should be as short as possible; hum may be a problem if the leads are too long.

If you have an amplifier where it's inconvenient to unground one end of the emitter bypass capacitors to obtain gain control, **fig. 3** shows two alternate methods for the same result. In the circuit shown in **fig. 3A,** a large bypass capacitor is connected between one or more stages in the controlled amplifier and the control diodes on PC3. When the diodes are forward biased by large audio signals, increasing amounts of the signal are bypassed to ground.

The one disadvantage here is that the frequency response of the amplifier will change with compression since a reactive element is used to bypass the signal to ground. This effect can be eliminated by placing a resistive T attenuator between the controlled stages as shown in fig. **3B.** The control diodes on PC3 then act as a variable resistance leg in the T attenuator.

Since resistive elements are used, the frequency response of the controlled amplifier remains essentially unaffected. The shunt capacitor in **fig. 3A** and the resistors of **fig. 3B** will require some tailoring to your particular amplifier, but the values shown in the diagrams should work with most transistor audio amplifiers using conventional grounded-emitter circuitry.

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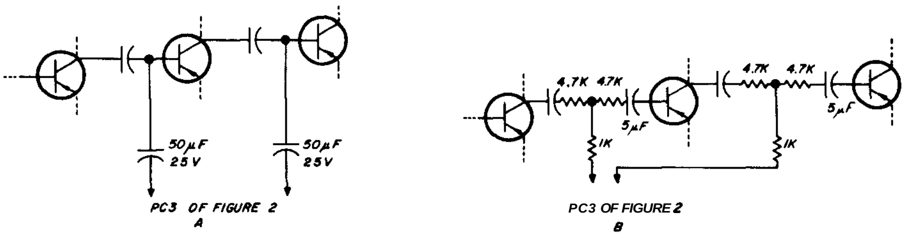
fig. 3. Alternate methods of putting a gain control between the stages of an audio amplifier.



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transistor rig for 40 meters

A crystal-controlled five-watt transmitter for 7 MHz using a 2N3553

Ed Marriner, W6BLZ and John Meredith, K5GXR, 58 Colima Street, La Jolla, California 92037

If you want to go solid-state, here is a crystal-controlled 40-meter transmitter that has an output of two-and-one-half watts. It was built more or less as a curiosity, but it worked far better than expected. It's simple to build, uses one transistor, and is powered by a standard 12-volt lantern battery. The battery should last for several months of normal CW operation.

We often forget that in the early days of wireless, a five-watt transmitter was the average power most amateurs had on the air. Although there weren't as many layers of QRM then as there are now, an awful lot of QSO's were made. Even today, a low-power transmitter can give some surprises; one evening I worked a station near Sacramento, 500 miles away, with a 599X report. It was impressive. A whole new world opened up.

The similarity between the "old" days and the "new" becomes apparent when you start to buy parts for a transistor transmitter. The transistor costs about the same as a "bootleg" 210 did in the early thirties. Let's face it, transistors and associated tiny parts are expensive, and small switches and components are not available yet on the surplus market.

This transmitter costs about $10.00 to build, but it depends somewhat on the parts you use. There are all grades, and prices vary considerably. Basically, the RCA 2N3553 power transistor costs **$4.75.** You need that! A 50-

microamp meter can run anywhere from $2.95 for an inexpensive imported tuning type without a scale up to $15 or so. The inexpensive one does just as good a job here because it only functions as an rf output indicator.

You also need a power source. A lantern battery is the most convenient, and costs about $3.00. An old 12-volt car battery and a trickle charger do a nice job; or you can build a power supply with a filament transformer—for more money.

circuit

This circuit is a Pierce-type crystal oscillator using a single 2N3553 transistor. The toroidal tank coil is used because the field is contained in a small space and miniature construction is possible. Capacitor C3 tunes the tank circuit to resonance on 40 meters, and C4 couples the load to the transmitter. Capacitor C1 is the feedback capacitor and is adjusted for proper keying and good circuit efficiency.

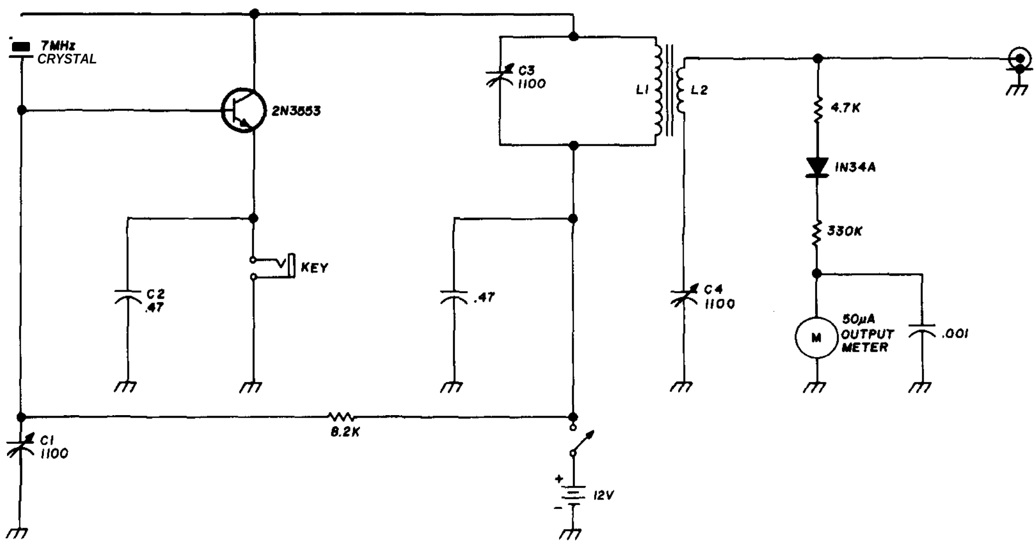
The antenna is inductively coupled to the collector of the 2N3553 through the final tank circuit and its link. Capacitor C4 controls antenna loading. The load impedance should be between 40 and 100 ohms resistive for proper operation. A relative power meter is used for rf output indication.

Keying is done in the emitter of the 2N3553. Capacitor C2 does a fine job as a key-click suppressor. The emitter current with a 12-volt collector supply is from 200 to 300 mA, depending on internal battery resistance as it increases with use. This represents 3 to 4 watts dc input. Collector-circuit efficiency runs about 70% at this level.

construction

The transmitter shown in the photographs is built in an LMB 138 chassis, 3 x 2-1/2 x 2 inches, although a larger chassis may be used for easy wiring. Placement of components and leads is not critical. The various holes in the chassis should be punched, drilled, and if possible, wired before starting assembly.

fig. 1. Schematic diagram of the one-transistor rig for 7 MHz. Power output of 2½ watts was measured with a Bird wattmeter. Construction of L1 and L2 is shown in fig. 2.



Construction of the toroidal tank coil, L1 and L2, is shown in **fig. 2.** L2 should be wound first using about ten inches of number-18 enamelled wire. Leave about one-and-one-half inches of lead, and count each turn as it goes through the center of the core. Space the turns so that the last turn is about one-eighth inch from the starting turn. Clip the excess wire so you have about one-and-a-half inches of lead left on L2. L1 is wound in a similar manner—starting 90 degrees clockwise from the starting point of L2 and winding counter-clockwise from that point. Leave one-and-one-half inches of lead on each end.

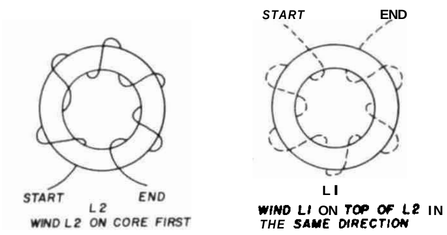
The trimmer capacitors, tie lugs, crystal socket, key jack, antenna jack, power lead grommet, switch and meter are mounted in that order. The transistor should be provided with a heat sink and soldered in the circuit last. Be careful not to heat the transistor leads; use a heat sink such as long-nose pliers between the solder joint and transistor.

If you can't find all the parts at your local radio store, they can be purchased from any of the larger mail order distributors such as Allied Radio or Lafayette Radio. It's also a good idea to send postcards to surplus dealers and get on their mailing list; you'll find inexpensive meters and other parts necessary to build modern equipment.

tuning

After checking the circuit over carefully for wiring errors, connect the power supply. Take extra precaution, and check the polarity of the battery leads—this may save you a burned out 2N3553. Next, plug in a 40-meter crystal and connect the transmitter to a dummy load. You can make one from three parallel-connected, 150-ohm, I-watt resistors.

fig. 2. Construction of the final tank coil. Each winding consists of 11 turns number 18 enamelled on an Amidon Associates' T-94-2 ferrite core.



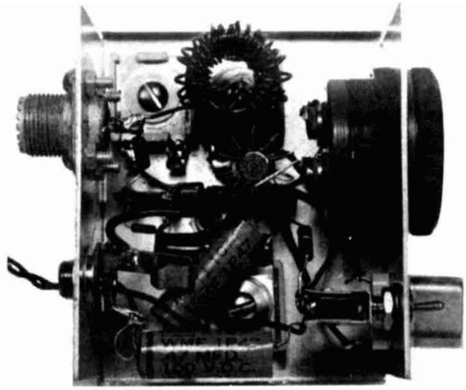
Tighten C1 to full compression, and then back it off two turns. Key the transmitter, and tune C3 until the output meter kicks. Tune C4 and C3 for maximum indication on the meter. If you don't get any output, try changing C1 while simultaneously tuning C3.

*Amidon Associates, 12033 Otsego Street. North Hollywood, California 91607.

Once you get output, tune C4 and C3 for maximum indication on the meter. Peak the output with C1 while listening to the keyed note on a receiver; set C1 for maximum power output consistent with good keying characteristics.

An antenna can now be substituted for the dummy load, and C3 and C4 tuned for maximum output indication on the meter. If the antenna does not provide a resistive load between 40 and 100 ohms, use an antenna tuner.

Internal parts layout of the 40-meter transistor transmitter.



To check power output, you can use an rf probe and a vtvm. You should measure 3 volts on the hot side of the 50-ohm dummy load.

results

This transmitter has been tested by several amateurs. WB6OGA had excellent results throughout Southern California. K6WC made a critical check and found the keying clean with no trace of chirps or clicks at a distance of 13 miles. The ground signal was a constant S7 at this distance, indicating good possibilities of 40-meter skip for long-distance work. The overall dc to rf output conversion efficiency of 70% is ideal for portable battery operation or as a driver for a higher power amplifier; we'll build one as soon as we can save the money to huy a high power transistor!

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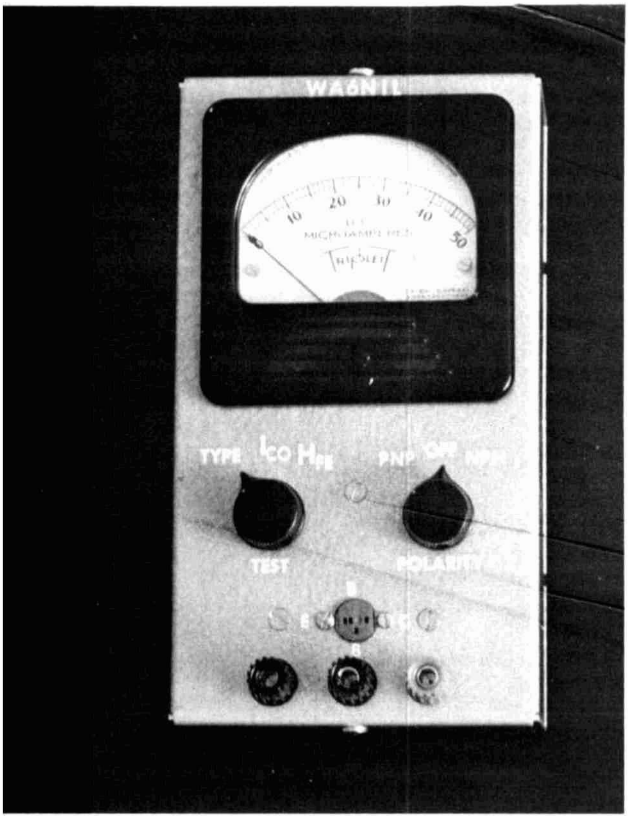


handy transistor tester

A transistor checker for measuring type, polarity, leakage and dc gain

Ja masney, WA6NIL, 25a1
Wave Ste wt, Palo Alt California 94301

A transistor tester is almost a necessity for anyone who experiments with the little three-legged fuses. This one provides direct indication of transistor type (germanium or silicon), polarity (NPN or PNP), as well as I_{CEO} (collector leakage current, and h_{FE} (dc beta or common-emitter current gain). It is fast and easy to use; large batches of transistors can be sorted into types, the open and shorted ones picked out, and the rest graded for gain. Several of these testers have been built by members of local radio clubs and they have all worked very well.

The photograph of the instrument shows the two rotary switches, test and polarity, the universal socket which accepts the standard TO-5 or older three-in-line base and the three binding posts connected in parallel with the socket. A worthwhile addition would be another socket to handle the small TO-18 transistors.

operation

Operation of the tester is straightforward. When starting with a transistor of unknown type, turn the **test** switch to **type** and the **polarity** switch **off.** Plug in the transistor and turn the **polarity** switch to PNP. A meter reading below 10 on the 0-50 scale indicates that you have a germanium PNP device, while a reading of 20 to 30 indicates a silicon PNP transistor.

A reading near full scale shows you have an NPN transistor; throw the **polarity** switch to NPN. Again, a reading below 10 shows a germanium NPN unit, and a reading of 20 to 30 shows a silicon NPN. If the reading is near full scale in NPN position too, the transistor is either open or not properly plugged in. If the test transistor is shorted, the meter will read near zero in both NPN and PNP positions.

Battery condition mav be checked on the meter: with no transistor plugged in, and the **test** switch in the **type** position, turn the switch to either NPN or PNP polarity and read the meter. When it drops below 45 on the 0-50 scale, the batteries should be replaced.

With the **polarity** switch in the correct position for the test transistor, turn the **test** switch to **I_{CEO}**. Read the leakage current in microamperes. A good germanium transistor should read less than ten, and a good silicon transistor less than one microampere. Now turn the **test** switch to h_{FE} and read beta on a 0-250 scale (multiply the 0-50 scale reading by five).

When sorting through a batch of the same type transistors to pick out high-gain devices, leave the **polarity** switch in the proper (NPN or PNP) position and the **test** switch at h_{FE}. Plug the transistors in, and note the beta reading of each.

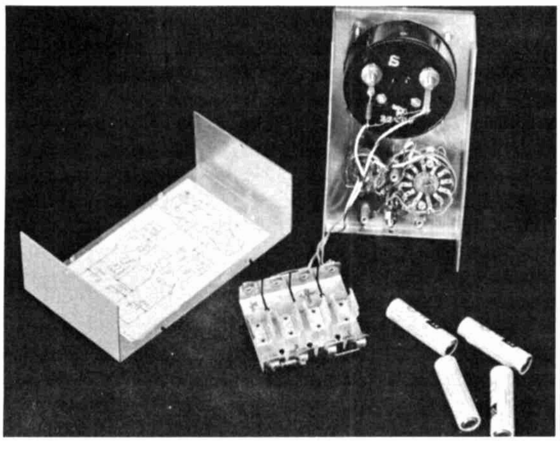
Diodes may also be tested on this instrument. Connect them to the **C** and B pins on the socket or binding posts, cathode to **C.** Turn the **polarity** switch to NPN and the **test** switch to **type.** A germanium diode will read below 10, and silicon, 20 to 30 on the 0-50 scale. If the meter reads full scale, you have the diode in backwards or an open device; reverse the leads or throw the switch to PNP. If the meter still reads full scale, the diode is open. With the proper polarity, turn the **test** switch to **I_{CEO}** and read leakage current. If the meter needle goes full scale, the diode is shorted.

the circuit

The circuit, fig. 1, appears more complicated than it actually is. To understand the operation of the various test circuits, refer to the simplified diagrams in **fig. 2.** For the **type** test, **fig. 2A,** the meter is connected to the battery through a string of resistors just large enough to make the meter read full scale at full battery voltage. A tap is made on the resistor divider at the point where the voltage drop across that part of the string including the meter is one volt.,

The transistor is connected across this one-volt drop so that its collector-base and emitter-base diodes are forward biased. With a germanium transistor with a forward drop of about 0.15 volt, the meter will read below 10 on its 0-50 scale; with a silicon transistor with about 0.5 volt drop, the meter will read between 20 and 30. An NPN transistor connected as shown will not conduct at all, and the meter will read full scale, indicating that the other position of the **polarity** switch

Layout of the handy transistor tester is not critical. A copy of the schematic is pasted on the back cover for ready reference.



should be tried.

The collector-cutoff leakage current circuit is quite usual **(fig. 2B)**. The battery is connected through the microammeter and a protective resistor to the collector-base diode of the transistor, reverse-biasing it. A good silicon transistor should show practically no leakage current, less than 1 microampere, and a small germanium type, less than 10 microamperes. Large germanium power types will generally show too much leakage for this tester.

The dc beta (h_{FE}) test circuit is shown in **fig. 2C.** The battery voltage is applied through R1 to the base-emitter junction of the transistor. The value of R1 is chosen so that the base current is approximately 20 microamperes. The battery voltage is also applied to the collector through the meter and R3. The meter circuit is shunted by R4 to read 5 milliamperes full scale.

The collector current is essentially the base current multiplied by dc beta. The meter may be calibrated directly in h_{FE} (dc beta) if you want, with a full-scale reading of 250 (5 mA ÷ 20μA = 250). I did not bother to do this because it's easy to mentally multiply the 0-50 scale reading by **5.**

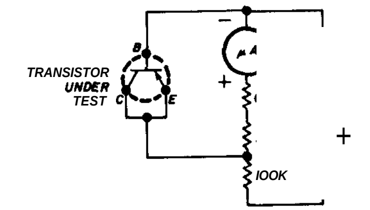
The value of R1 is a compromise. The voltage drop across it is the battery voltage minus the base-emitter diode drop, which is not the same for silicon and germanium transistors. Therefore, the battery voltage should be high in comparison to the change in diode drop, so that the current through R1 is nearly independent of the transistor type. On the other hand, the battery voltage should be low so that the collector-breakdown rating is not exceeded. A compromise must be made, and *6* volts is a reasonable value.

The circuits for testing an NPN transistor are exactly the same as shown in **fig. 2,** except that the polarity of both the battery and meter are reversed by the **polarity** switch.

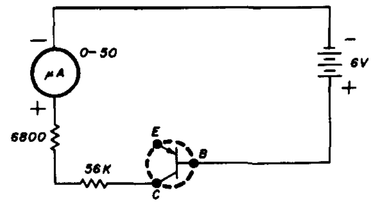
The dc beta reading is only moderately accurate and is affected by battery loading and aging, differences in emitter-base voltage drops between silicon and germanium transistors, and resistor tolerances. In addition, the collector-leakage current adds to the 20 microampere base current through R1 to make the beta read high, especially with germanium transistors. The readings are sufficiently accurate for most amateur purposes, however.

Returning to the overall circuit diagram of **fig. 1**, the **test** switch S1 selects any of the three circuits of **fig. 2.** The resistor R1 is put in series with the meter resistance (1100 ohms in my meter) to bring the total voltage drop

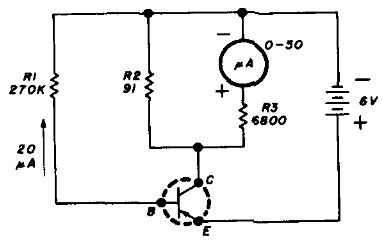
fig. 2. The three basic test circuits Incorporated in the handy transistor tester. PNP transistors ere shown here—for NPN devices, reverse battery and meter polarities.



A. Type test



B. Collector-cutoff leakage test (I_{CEO})



C. DC current gain test (h_{FE} or beta)

at full-scale meter current to about 0.4 volt. A protective diode D1 is connected across this part of the circuit; it draws practically no current when the meter is on scale, but will not allow more than about 50% overload current to flow. D1 may be almost any silicon diode **(not** germanium). It must be connected as shown—it will not protect the meter if it's put in backwards.

The three bypass capacitors at the socket are small ceramic discs with very short leads. These help to keep the very hot vhf transistors from self-oscillating under test.

construction

Layout and construction are not critical, but when you're wiring a circuit with a sensitive meter and a battery, mistakes can be expensive. For one thing, don't try to measure meter resistance with an ohmmeter! The unit shown in the pictures is built in an LMB 138 chassis-box, but anything big enough to hold the meter and other parts will do. The battery holder is mounted on spacers behind the rotary switches.

The most expensive part is the meter. A 50-microampere movement is desirable to measure the leakage current of silicon transistors, but less sensitive meters can be substituted as described later. Aside from this, the only unusual part is the **test** switch S1, a **non**-shorting, four-pole, three-position rotary switch. I used a Centralab type PA-2011, modified by using shorter spacers and sawing off the rear of the shaft to make room to mount the battery holder behind it. The **polarity** switch, also a four-pole, three-position unit, may be either shorting or non-shorting since the center position is **off.** I used an inexpensive imported switch (Lafayette #99-6156).

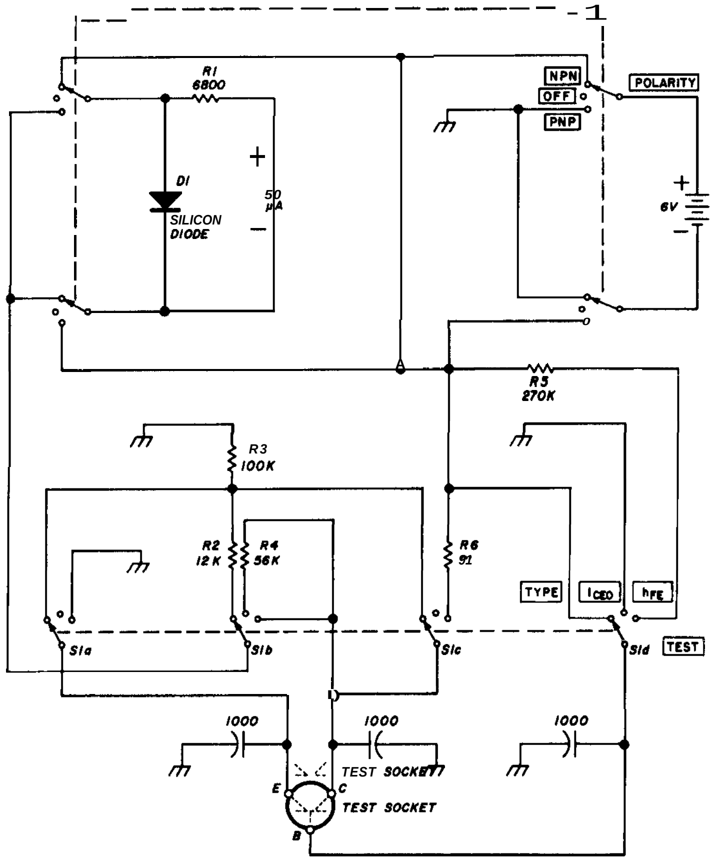


fig. 1. Schematic diagram of the handy transistor tester. Rotary switch S1 should be a non-shorting type.

Four 1.5-volt penlite cells are used for the power supply. Alkaline batteries are excellent, since their voltage holds up over long periods of time; mine are over a year old and still going strong. Even more constant voltage and longer life may be obtained with mercury or rechargeable nickle-cadmium cells, but their slightly different voltage (5.36 volts for 4 cells or 6.7 volts for 5 cells respectively) would require adjustment of resistor values.

Changing the circuit to accommodate a different meter, a different battery voltage, or a different dc beta range only requires a bit of Ohm's law. R1 should be chosen so that the total voltage drop across it and the meter is 0.4 volt at full-scale meter current. R2 is then picked so that the voltage drop across the meter, R1, and R2 is one volt at full-scale current. R3 is now calculated so that the total voltage drop across the meter, R1, R2 and R3 is equal to the voltage of new batteries at the same current. R4 is not critical—large enough to limit the meter current to about two or three times full scale if a shorted transistor is plugged in during the I_{CEO} test.

R5 is picked for an average 20 microamperes base current (slightly under 20 with a silicon transistor, slightly over with germanium). To compute R5, the voltage drop across it may be taken as the voltage of new batteries minus 0.35 volt.

R6 is selected to shunt the series combination of the meter plus R1 to read the desired full-scale current for the beta scale; that is, 20 microamperes times the full-scale beta reading. As an example, suppose you have a 0-150 microampere meter with 500-ohms resistance. The total resistance of the meter plus R1 is 0.4V ÷ 150 μA or 2667 ohms. R1 is 2667 ohms minus 500 ohms, or 2167 ohms; 2200 ohms is close enough.

Now, suppose you want a full-scale beta reading of 300; this will go nicely with a 0-150 meter scale. The total collector current will be 20 μA times 300, or 6 mA. Of this, 150 microamperes will flow through the meter, leaving 5.850 mA for the shunt, R6. The value of R6 is 0.4V ÷ 5.85 mA, or 68.4 ohms, and a 68-ohm resistor can be used. Judicious selection of resistors and checking them against an external meter will give better accuracy than just picking the resistors out of the junk box.

I assume that you know which leads are which on the transistor being tested. If you should run into an unknown oddball, a method of identifying the connections was described by W9QKC[1]. This is an excellent article and well worth rereading. Transistor base diagrams have also been published in various books[2,3].

I finished up my tester by gluing a copy of the circuit diagram inside the back cover, and short operating instructions on the outside. It's a great comfort to have this information available right on the instrument after a couple of years when you've forgotten what you put in it. For a really deluxe job, you could make up a special meter scale with "Germanium - Silicon - Reverse" sectors and a direct-reading dc beta scale.

The completed tester is fast and convenient enough to be kept right on the bench while you're building up experimental circuits. Whenever you suspect that the transistors you are using have been damaged, this checker saves a lot of time and exasperation. You'll soon wonder how you ever got along without it. Many thanks are due to Dave Annett, WB6DBE, for taking the photographs, and to Lloyd Provan for enlarging them.

references

1. Donald Grayson, W9QKC, "How to Check Transistors with an Ohmmeter," 73, March 1962, p. 14.
2. "General Electric Transistor Manual," Seventh Edition, 1964, pp. 575-589.
3. "RCA Transistor Manual," SC-13, 1967, pp. 449-456.

ham radio

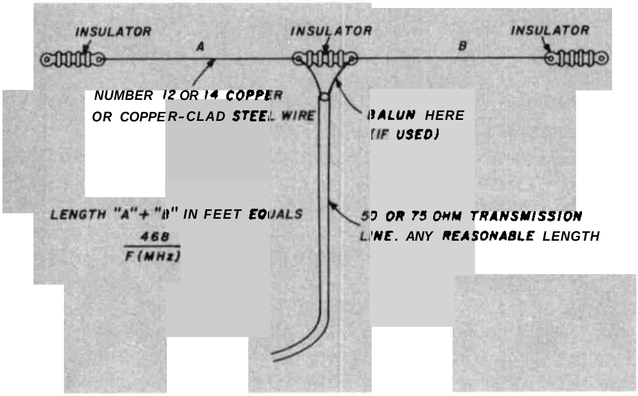


fig. 1. Basic half-wave dipole antenna.

simple 1-, 2- and 3- band antennas

A discussion of the merits and design of various multiband dipoles

Herbert S. Brier, W9EGQ, 385 Johnson Street Cary, Indiana 46402

The most popular single-band amateur antenna for the lower-frequency bands is the half-wave dipole fed in the center with flexible, low-impedance coaxial line. The dipole length is calculated from a formula that is found in most of the handbooks: length (feet) = 468/frequency (MHz). Calculated half-wavelengths for representative frequencies in the 3.5-, 7-, and 14-MHz amateur bands are shown in **table 1.**

Although the size of the wire, the method of fastening insulators, and proximity to other objects will have some effect on the antenna's resonant frequency, it's surprising how close a carefully measured and constructed antenna will resonate to the design frequency.

horizontal vs. inverted-v antennas

Many 3.5- and 7-MHz dipoles are installed as "inverted V's" (center high, ends low), because it is easier to put one short stick on the house to support the center of an inverted V than to install two high poles to support a horizontal antenna. However, for the same (center) height, the horizontal antenna usually gets out a little better because of its greater separation from power-absorbing objects.

antenna height

For daytime work on 80 and 40 meters, and for close-in work on the other bands, antennas 20- to 25-feet high perform about as well as higher antennas. But over longer distances, average results improve almost linearly with heights up to 50 feet, and more slowly for greater heights. Nevertheless, a low antenna outperforms a high one often enough to make it interesting—especially over medium distances.

length vs. frequency

When you want to resonate an antenna on a precise frequency, cut it slightly longer than the calculated length and put it up. Then insert an **accurate** SWR bridge in the transmission line and measure the SWR at different frequencies near the design frequency. When the frequency of minimum SWR is found, shorten the antenna to obtain minimum SWR at the desired frequency.

On the 3.5-MHz band, the resonant frequency of a half-wave horizontal dipole changes approximately 2.5 kHz per inch—slightly less near 3.5 MHz and slightly more near 4.0 MHz. On 40 meters, the change is about 9 kHz per inch, and on 14 MHz, about 36 kHz per inch.

antenna impedance

In free space, the theoretical impedance of the center of a lossless half-wave antenna of zero diameter is close to 73 ohms. Practical antennas installed at heights that are integral multiplies of one-quarter wavelength (70 feet at 3.5 MHz, 35 feet at 7 MHz and 17.5 feet at 14 MHz) have center impedances which are quite close to the theoretical value. The impedance fluctuates above and below this nominal value at heights that are odd integrals of 1/8-wavelength. For example, at heights of 3/8 wavelength (105 feet at 3.5 MHz, 52 feet at 7 MHz, and 26 feet at 14 MHz), the impedance goes up to 95 ohms. And at a height of 3/8 wave (170 feet at 3.5 MHz, 85 feet at 7 MHz, and 42 feet at 14 MHz), the center impedance drops to about 58 ohms. Similar, but gradually lessening, fluctuations occur at higher odd multiples of 1/8 wavelength.

Below 1/4 wave, the center impedance of a half-wave antenna goes down with decreasing heights, and over a perfect ground, the impedance will reach a very low value at heights of less than 3/16 wave. Over actual ground, losses increase rapidly as antenna height **de**creases, and the effective center impedance of a horizontal half-wave antenna stabilizes between 40 and 50 ohms for heights under 1/8 wave or so.

Translated into practical terms, 52-ohm coaxial cable matches the center impedance of horizontal 3.5-MHz dipoles at heights of up to 50 feet quite well. For greater heights, 75-ohm coax is a better choice. It is also recommended for dipoles on 7 MHz and higher for heights above 30 feet or so.

antenna baluns

Simple, center-fed dipoles first achieved popularity with the development of efficient, flexible, low-impedance, twin-lead transmission lines which matched their nominal 73-ohm center impedance. Even after transmitters with unbalanced pi-network output circuits became standard, many amateurs continued to use twin lead to feed their dipoles by grounding one conductor and connecting the other one to the hot output terminal of the transmitter.

Although this arrangement worked fine, the word gradually got around that you couldn't feed a balanced transmission line from an unbalanced pi-network. As a result, most amateurs dutifully switched from twin lead to coaxial cable. And our antennas worked fine —just as well as when fed with twin lead. Next, the theoreticians came up with the edict that you couldn't feed a balanced antenna with unbalanced coaxial cable unless you put a balancing device such as a balun between the line and the antenna.

This means that all the dipoles, Gamma-matched beams, split-dipole beams, etc., fed directly with coaxial cable really did not work; we just thought they did. Being great believers of theory, we quickly installed baluns between our coaxial transmission lines and balanced antennas—with just about the same results as before.

However, I do have to give the antenna

balun its due. When a balanced horizontal antenna is installed on a site free of trees, tall buildings, and utility wires that might distort its radiation pattern, and fed with coaxial cable and a balun, the local radiation pattern is symmetrical. When the balun is removed from the circuit, the radiation pattern skews slightly towards the side of the antenna connected to the center conductor of the coaxial transmission line.

Furthermore, with the balun installed, reception of local vertically-polarized signals is usually somewhat poorer than with the balun removed. This indicates that the balun prevents vertically-polarized signals and noise picked up on the outer conductor of the coaxial transmission line from getting into the receiver.

simple multi-band antennas

Undoubtedly, the most-efficient simple multi-band antenna is a half-wave dipole at the lowest operating frequency, center fed with open-wire transmission line coupled to the transmitter through an antenna coupler. The transmission line may consist of open-wire TV ladder line or a pair of number 14 or 12 conductors spaced two to four inches apart.

Although the SWR may be quite high with this arrangement, line losses are still lower than with a perfectly matched coaxial line of the same length on the same frequency.

table 1. Lengths of half-wave dipoles for 3.5. 7 and 14 MHz.

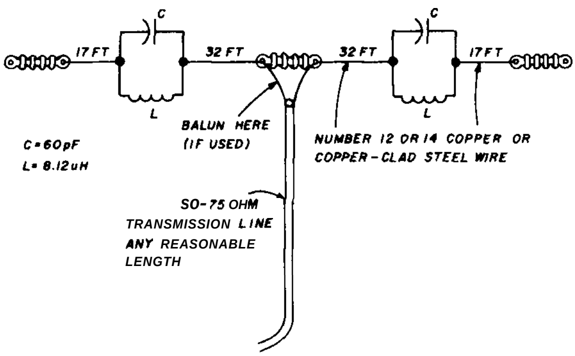
frequency (MHz)	length
3.5	133' 8".
3.7	126' 6"
3.8	123' 2"
3.9	120' 0"
4.0	117' 0"
7.0	66' 10"
7.1	65' 11"
7.2	65' 0"
7.3	64' 1"
14.0	33' 5"
14.1	33' 1"
14.2	32' 11"
14.3	32' 9"
14.35	32' 7"

Nevertheless, many amateurs shy away from open-wire feeders because they don't like antenna couplers. Also, open-wire line is somewhat more trouble to install than coaxial cable.

trap antennas

One popular multi-band antenna is the "trap" dipole. In a 3.5- and 7-MHz antenna for example, a pair of 7-MHz parallel-resonant "traps" are inserted in a 3.5-MHz dipole 114-wavelength (at 7 MHz) on each side of the center insulator as shown in **fig. 2**. At

fig. 2. An 80- and 40-meter trap dipole with typical values for operation on 7.2 and 3.9 MHz.



their resonant frequency the traps look like very high-resistances—like insulators, in fact —inserted in the antenna. On 7 MHz, therefore, the section of the antenna between the two traps acts like a conventional half-wave dipole.

At 3.5 MHz, the traps exhibit inductive reactance and act as loading coils to decrease the resonant frequency of the antenna. Resonance is restored by cutting off the end sections of the antenna. The overall length is usually reduced about 10 per cent.

Adjustment of the antenna consists of inserting an SWR meter in the transmission line and adjusting the lengths of the inner sections of the antenna for minimum SWR at the desired frequency in the 7-MHz band. Then, operations are transferred to 80 meters; the lengths of the end sections of the antenna are adjusted for minimum SWR at the desired frequency. If the lengths of the outer sections are changed only a reasonable amount, it will have very little effect on 7-MHz resonance.

Incidently, an almost endless combination of tuned-circuit values and conductor lengths may be used to obtain operation on two or more discrete frequencies, but adjustments can become quite tedious.

The efficiency of two-band trap antennas is almost the same as the efficiency of conventional dipoles on the lower frequency of operation. At the higher frequency, however, trap losses do reduce antenna efficiency. In one such 40- and 80-meter antenna using traps with a *Q* of 180, losses were just under 1 dB on 40 meters.

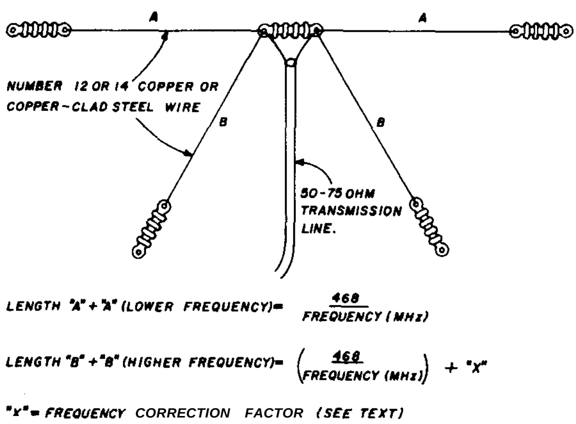


fig. 3. Two-bend multi-dipole antenna.

multiple-dipole antennas

Another simple multi-band antenna is the multiple-dipole shown in fig. 3. Referring to the figure, on the lower frequency band the longer pair of wires act as a conventional half-wave dipole, with the shorter wires having negligible effect on the operation. On the higher band, the short dipole radiates, and the long one goes for the ride.

As long as the two dipoles are resonant on harmonically related frequencies—say 3525 and 7050 kHz—operation is as described; otherwise, strange things occur. As an example, assume that the low-frequency dipole is resonant on 3900 kHz and the high-frequency one is tuned to 7250 kHz. At 3900 kHz, the antenna will perform as predicted; but at 7250 kHz, the SWR will be exceedingly high.

If the frequency is increased above 7250 kHz, the SWR will decrease to a minimum near 7800 kHz (the second harmonic of 3900 kHz). The explanation is this: at 7800 kHz, the 3900-kHz dipole appears as a simple high resistance across the transmission line. Below 7800 kHz, however, it looks like a capacitor in parallel with the resistance across the transmission line. The resulting capacitive reactance pushes the SWR up.

The solution to the problem is both simple and effective: increase the length of the higher-frequency dipole so it will present inductive reactance to the transmission line. This will cancel out the capacitive reactance presented to the transmission line by the lower-frequency antenna.

Assuming that the minimum SWR occurs at 7800 kHz and you want minimum SWR at 7250 kHz—a difference of 550 kHz—then the dipole should be lengthened approximately five feet for minimum SWR at 7250 kHz. This is based on the fact that the resonant frequency of a 40-meter dipole changes approximately 9 kHz per inch.

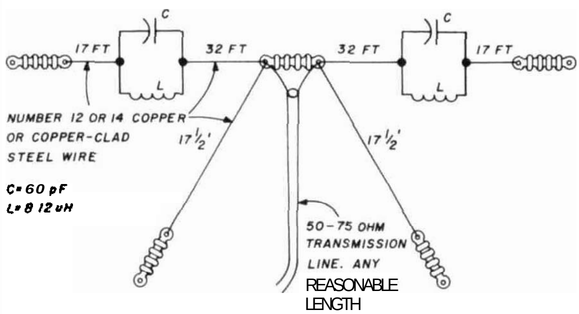
An important precaution in installing multiple-dipole antennas is to space the ends of the shorter dipole at least a foot (preferably more) from the longer one. Otherwise, slight variations in the spacing between the two antennas as the wind blows will cause a large variation in feedline SWR. One way to assure adequate spacing between the two dipoles is to install one of them horizontally and the other one as an inverted V.

More than two dipoles can be fed from the same transmission line, but the secret to success is to cut each one somewhat longer than the calculated length. Then, starting at the lowest frequency, trim each one for minimum transmission-line SWR at the desired frequency.

three-band antennas

A rather neat three-band antenna can be made by combining a 3.5- and 7-MHz trap

fig. 4. A three-band antenna **combining** the features of the trap and multi-dipole antenna.



dipole with an extra dipole for a third band, as shown in fig. 4. With the lengths shown in the drawing, the antenna is resonant on 3900, 7250, and 14200 kHz. While this antenna is no world-beater on 20 meters, it does all that can be expected of it on the three bands.

harmonic radiation

Some amateurs avoid multi-band antennas from fear of excessive harmonic radiation. Fortunately, if the transmitters or transceivers they are used with have adequate harmonic suppression built in and are properly adjusted, harmonic radiation should be no problem. Of course, it is always wise to check harmonic radiation with any new antenna, and if necessary, use an antenna coupler to prevent excessive harmonics from reaching the antenna.

ham radio

Why don't they.. .

1. make 10- and 15-meter transverters for the popular triband, single-band, and other transceivers?

Since there are thousands of limited-frequency range transceivers in use, why not? Could be of interest to a manufacturer or for home-brew articles.

2. make 432-, 1296-MHz and higher frequency elements for replacement of TV elements in uhf TV parabolic reflectors?

Souncls like a good idea. Perhaps some enterprising manufacturer who wants to sell more antennas will come up with it after reading this!

3. do away with credit slip refunds for small amounts by dealers?

Agreed! I have them in all different shapes, forms and amounts. It is annoying. I can see no reason why postage stamp refunds cannot he made. Hams can always use them.

4. make brass chassis for vhfers, or at least plates to fit standard chassis for home-brew converters, etc?

Here's an opportunity for someone in the business, or someone who wants to get into the business, to do so. Might as well make cavities, or boxes to use as cavities, for uhf, etc. Also brass tubing for baluns.

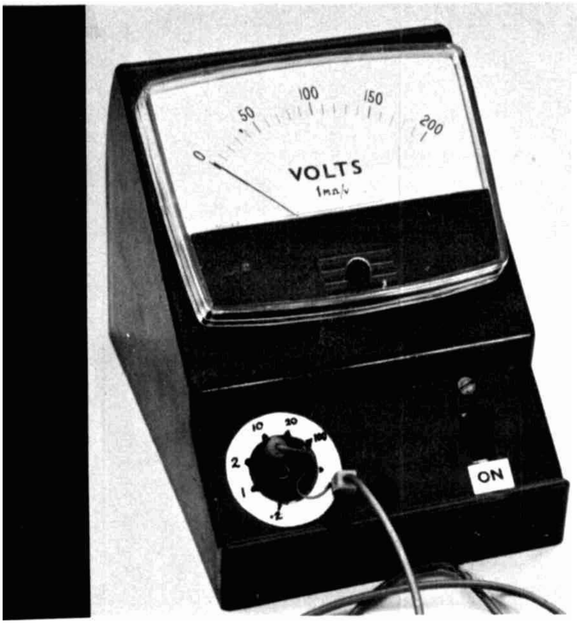
5. allow credit for commercial radiotelegraph license code tests for the Amateur Extra Class?

I don't know. Seems like it would accomplish the same purpose and save Uncle and the taxpayer money. Why not ask your ARRL Director?

6. include low-level output jacks on transceivers for low power so that the final can be turned off for local contacts, to drive transverters, etc?

Right. It shouldn't be necessary to "swamp" a 150 or 500 watter to transvert. I did this with a KWS-1 for 28 MHz without drilling or using a soldering iron. Will be glad to tell anyone how, if you send me a stamped, self-addressed envelope.

Chas. Spitz, W4API



transistor voltmeter

R. S. Maddever, Geelong Grammar School, Corio, Victoria

A transistor voltmeter built around the design featured in the April issue of ham radio is shown in the accompanying photographs. The amplifier components are mounted on a small painted-circuit board which is attached directly to the meter terminals. The multiplying resistors are grouped around a miniature 7-pin tube socket which is used with pin plugs instead of a multi-position switch. This conserves space in the meter box.

capacitance measurement

An interesting feature of the scale is the diamond-shaped mark at 37 divisions. This is **100/e,** where e is 2.7, the base of natural logarithms. With this mark, you can approximate the value of larger capacitors very simply with the aid of a battery. Its use is based on time constant; if a capacitor (C) in parallel with a resistor (R) is charged to a voltage (V) as shown in fig. 1, when the voltage source is removed, the voltage across the capacitor will decay to a value of **V/e** in RC seconds, where R is measured in ohms and C in farads.

In practice, R can be the input resistance of the transistor voltmeter, and the voltage source any suitable battery to provide meter deflection greater than 100. The transistor voltmeter is connected across the capacitor as

fig. 1.

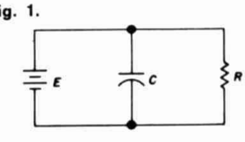
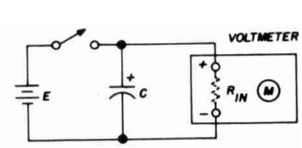
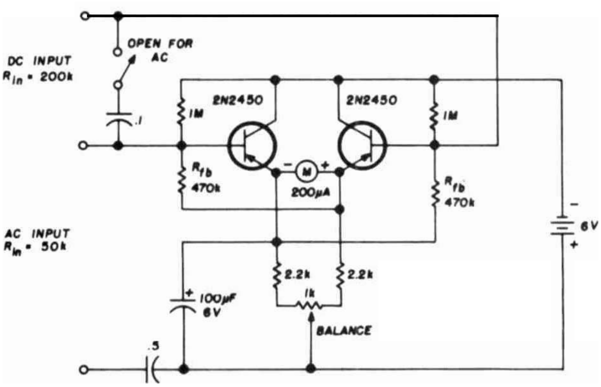


fig. 2.



shown in fig. 2, observing polarity if necessary. The battery is then connected to the circuit momentarily to charge C to anywhere above the 100-division mark. Then the length of time it takes the needle to fall from the 100

fig. 3. The transistor voltmeter. The 0.5 µF capacitor on the ac input should be a low-leakage, non-electrolytic type. The input terminal between the dc and ac inputs is common to both.



mark to the 37 mark is timed in seconds.

From the time-constant formula, the value of the capacitor can be calculated from $C = t/R$; where C is in µF, t in seconds and R in megohms. With the instrument pictured, the input resistance is 10 megohms on the 10-volt range. Therefore, a voltage decay timed at 25 seconds indicates capacitance of 25/10 or 2.5 µF.

This value is only approximate because we haven't considered leakage resistance in the capacitor itself. Strictly speaking, the value of R you use in the calculation should be the parallel resistance of meter input resistance and capacitor leakage resistance. If you want to see if the leakage resistance is appreciable, you can repeat the measurement on a different voltage range, and compare the two calculated values.

Alternatively, you can charge the capacitor up and measure the voltage across it on a range which does not discharge it very quickly; then disconnect it for a time and remeasure it. If the voltage has decreased appreciably, you can make an estimate of the internal leakage resistance.

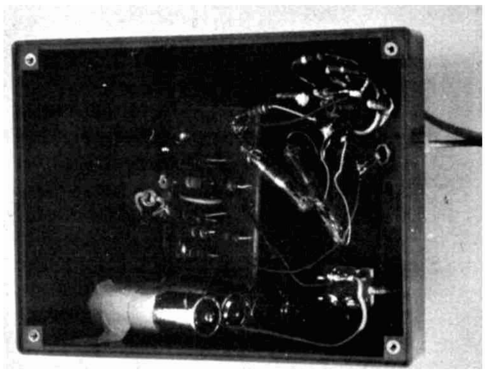
Don't use a battery voltage higher than the rating of the capacitor. For small capacitors, it is best to use the highest voltage range possible so the decay times are long and easily measured. The shortest useable times will vary with the inertia of the meter needle, but usually should be more than a second or two. On the 100-volt range (100 Megohm R_{in}) of my meter, the smallest capacitor that can be measured conveniently is about .01 µF. There is no upper limit capacitor size—if the decay time is too long you can always reduce the effective resistance with a parallel resistor.

ac measurements

I have found that by adding two capacitors to the meter circuit, it may also be used as an ac voltmeter of moderate linearity from at least 20 Hz to 200 kHz. The input impedance depends on the meter movement you use, but with a 200-µA meter, it will be on the order of 50k ohms per volt. The basic circuit is shown in fig. 3. The meter readings obtained with this circuit are listed in table 1.

The capacitor used to reduce hum and transient noise on dc settings must be switched out of the circuit for ac readings. If this isn't done, the same ac voltages are applied to both sides of the differential amplifier and there is no output. The large capacitor between the emitter and ground determines the low-frequency response of the instrument. With the value shown, the readings are within a few percent of the values shown in table 1 from below 20 Hz to well above 200 kHz. To obtain

Parts layout in the transistorized voltmeter is not critical.



meaningful results above 200 kHz, you have to match the generator and the volt-meter.

In this circuit, the emitter and base-bias resistors are larger than in the previous design. This was done to reduce standing current; the large quiescent current previously used was suitable for a 1-mA movement. For an even more sensitive movement (say 50 μA), the circuit resistances could be increased further.

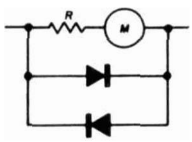
For applications requiring low input resistance or impedance such as bridge-null indications, the transistor voltmeter can be preceded by a balanced common-base preamplifier.

table 1. Meter deflection for the circuit shown in fig. 3.

input	scale deflections	
volts	dc	rms ac
0	0	0
0.1	25.2	11.0
0.2	50.3	23.6
0.3	75.0	36.0
0.4	99.0	48.6
0.5		61.0
0.6		73.4
0.7		86.2
0.8		98.8

D. K. Madden, of Hobart, Tasmania, has modified this design slightly for use as a null indicator. He used the meter protection system shown in fig. 4 for desensitization when the bridge was far from null. The resistor was

fig. 4. Meter protection circuit.

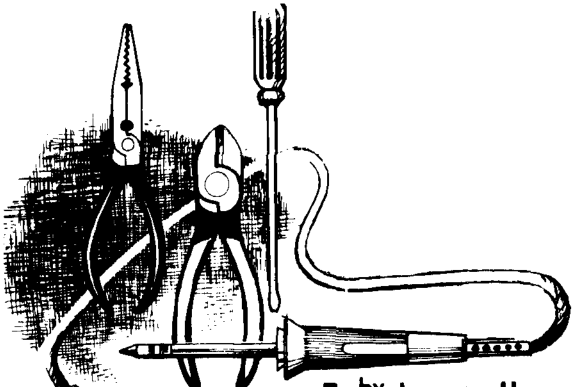


chosen to limit the meter reading to an on-scale value regardless of, the input overload. Because of the nonlinear characteristic of the diode, this maneuver didn't affect the sensitivity of the instrument near zero where it was needed.

reference

1. R. S. Maddever, "An Improved Transistor Voltmeter and its Applications," *ham radio*, April, 1968, p. 74.

ham radio



the repair bench

by larry allen

troubleshooting transistor ham gear

Hardly a month **goes** by that someone doesn't introduce new ham transistor equipment. It's about time we included some transistor troubleshooting information in this column. There are plenty of books on the subject, but it can't hurt to recap briefly some of the more fundamental transistor troubleshooting principles.

First of all, you should understand what a transistor does. In its most common use, a transistor is simply an amplifying device. For all practical purposes, you can see a transistor as a device through which controllable current flows. There are two kinds. In one, current flows from collector to emitter; in the other, from emitter to collector.

In either case, it is sufficient in this preliminary explanation for you to know that current in the transistor flows between these two elements. The phenomenon of amplification takes place because the transistor has a third element—called the base—that can control this flow of current.

Take a look at fig. **1A.** This will acquaint you with the schematic diagram of a PNP transistor, one of the two types (the other is called an NPN). A transistor needs dc operating voltages. For a PNP transistor, they are applied as shown in fig. **1A.** The emitter is usually grounded or connected to ground through a low-value resistor. Ground is therefore the common connection for all supply voltages.

The collector of a PNP transistor is connected to a strong negative voltage. This alone does not cause a flow of current in a normal transistor, but the possibility is there. It remains for forward bias to be applied to the base before collector current can flow. To cause normal current flow in this PNP transistor, a small negative voltage must be applied to the base. This makes the base more negative than the emitter (though still much less negative than the collector). With negative voltage applied to the base of a PNP transistor, the base-emitter junction is said to be forward biased, because the current flows easily across the junction from the N-material of the base to the P-material of the emitter. This base current is small, but it releases a large current flow between emitter and collector.

A small current in the base circuit controls a large current in the collector circuit; thus amplification is possible. Suppose a transistor, connected as in fig. **1A,** has a small audio voltage applied to the base along with the forward dc bias. What the audio voltage does is increase and decrease the bias, which in turn lets more and less collector current flow. Therefore, a tiny signal voltage or current in the base-emitter junction controls large amounts of signal current in the collector circuit. All you have to do is place a load resistor in the collector circuit and the current is converted to a strong voltage, and you have voltage amplification.

fig. 1. Power-supply connections for typical transistors: the PNP takes negative on collector and base (A), the NPN, positive on collector and base.

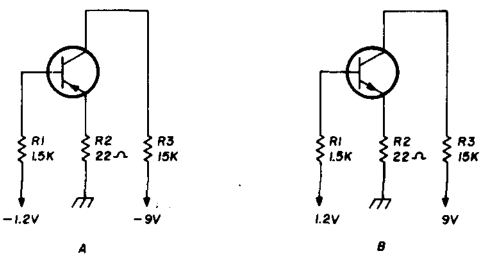


Fig. 1B shows the power supply connection for an NPN transistor. The collector of this transistor type is connected to a positive voltage. So is the base, but to a lower voltage. Again, any varying voltage, such as an af or rf signal, applied to the base will control the flow of current in the emitter-collector circuit.

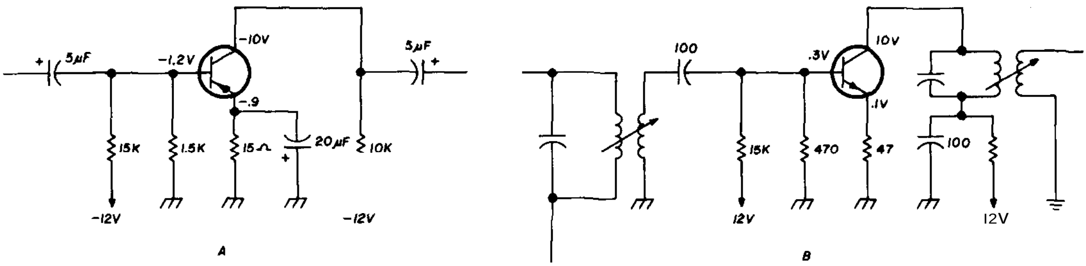
Amplification is thus accomplished by either type of transistor; the only difference is in the polarity of dc power-supply voltages applied to operate the transistor. **Figs. 2A** and **2B** show both types of transistor connected in normal amplifying circuits. **Fig. 2A,** using the PNP, is an audio amplifier, as you can see from the values of components. **Fig. 2B,** using an NPN, is an rf amplifier, which you can see from the tuned air-core transformers used to couple the signal in and out. Either polarity of transistor could be used in either circuit, merely by reversing polarity of the power-supply connections.

Normal operating voltages are shown in both schematics of **fig.** 2, as they would appear in a diagram of equipment you might want to troubleshoot. The "fun" of troubleshooting starts when the dc operating voltages on a transistor have changed from normal. One may be high and another low; one may be okay while another is way off; or they may all be wrong. Whatever discrepancy you find in measuring dc voltages at the elements of a transistor, your problem is to figure out what's causing it. The transistor itself could be at fault, or there could be a problem in one of the other parts.

One way to find out if the transistor is faulty would be to remove it from the circuit and check it either with a tester or with your ohmmeter. (I'll tell you how to use your ohmmeter for a quick test later.) The trouble is that most transistors are soldered in, and are not easy to get loose for out-of-circuit testing. As a result, it is better to be able to **interpret** incorrect voltages. It really isn't too difficult, if you stop to analyze the direction of current flow in the transistor, whatever is polarity.

Remember that current always flows from the negative power-supply terminal toward the positive one. Therefore, if the collector of a PNP transistor is connected to negative voltage, it can mean only one thing: current in that transistor flows from collector to emitter. Conversely, if you're dealing with an NPN transistor, the collector is positive, and the direction of current flow is from emitter toward collector.

fig. 2. Use of the transistor in typical circuits: PNP in an audio circuit (A), NPN in an rf circuit (B).



In most cases, however, knowing the direction of flow is not really as important as figuring out whether the voltage at the collector has increased or decreased. If it has increased—that is, if it's closer to the power supply voltage—you can reason quite easily that there must be less current flowing through the load resistor and consequently less voltage drop across it. Since the load resistor is in series with the collector circuit, it stands to reason that less current is flowing through the transistor, too.

If the power-supply voltage is negative, say –12 volts, the normal collector voltage may be about –10 volts. Suppose when you measure the voltage you find it to be –11.6 on the collector. This can mean only one thing: there is less collector current, signified by less drop across the load resistor. You can also be sure the cause is not the load resistor having increased in value, because that would cause a larger voltage drop and the voltage at the collector would be **less** than the normal –10 volts.

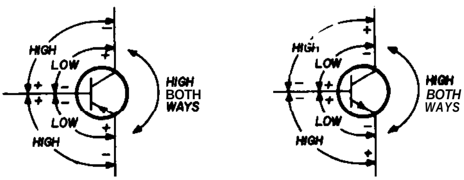
Exactly the same reasoning follows if you're dealing with an NPN transistor, where the collector voltage is positive. If the collector voltage shifts nearer to the power-supply voltage, whatever its polarity, it is a sign of reduced collector current. There is an outside possibility that the load resistor has lowered in value and therefore does not drop as much voltage. In that case, collector current would be unusually high, yet you'd still find a higher voltage at the collector. This seldom happens.

How many things can cause abnormal collector voltage? Or, phrased another way, depending on your conclusion from the first collector-voltage measurement, what makes a transistor draw less current than normal?

Two things: a faulty transistor, or reduced bias voltage (which causes reduced bias current). The way to find out which is to measure the bias voltage. If it is much lower than it should be, then chances are the bias voltage is the cause of reduced current; the transistor may be okay. If bias is normal, yet current is very low through the transistor, the transistor is probably defective.

Once you suspect by this method of reasoning that the transistor is okay, you might as well go ahead and check the other parts. Measure resistors to see if they've

fig. 3. Using ohmmeter readings to check the quality of a transistor. Both types should show high reedings in both directions between the emitter and collector terminals.



changed value, and check capacitors to see if they are leaky. Your ohmmeter is handy for both these tests.

If you decide to suspect the transistor, there are a couple more checks you can make before you disconnect the transistor for external testing. Clip your voltmeter—a vtvm is best—to measure collector voltage. If it reads at least slightly below the power-supply voltage it is connected to, proceed with this test. With a jumper lead, short the **base** of the transistor to the **emitter.** CAUTION: Be sure it is the emitter you short the base to; if you short it to the collector, you'll burn up the transistor. Watch the collector voltage as you make the jumper connection. If the transistor is operating normally, the collector voltage will jump upward, and read virtually the same as its power-supply source. If it doesn't, either the base element is open inside the transistor or the collector junction is leaky. Either condition signifies a defective transistor.

Now let's go back and see what we'd have done if the voltage on the collector were too low. Again keep in mind that this depends not at all on which polarity of transistor is involved; just consider "higher" voltage as being **closer to** the supply value and "lower" as being **further from** the supply value.

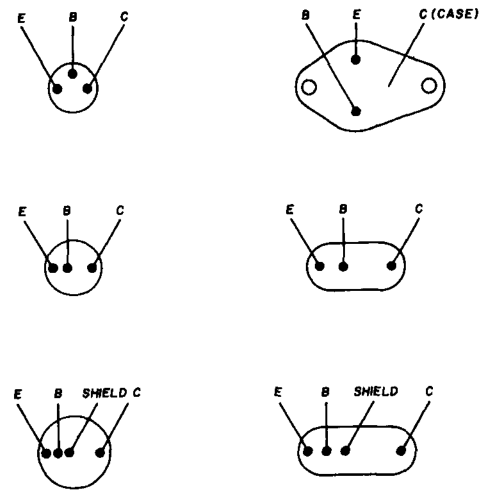
If the voltage at the collector is low, it means that either the load resistor has increased in value (unlikely) or more current is being drawn through it. If the latter, then more current than normal is flowing in the transistor collector. This in turn **may** be caused either by a faulty transistor that allows too much current to flow, or by too much bias current flowing between base and emitter.

Your voltmeter will tell you whether the bias voltage is too high. If it is, track down the trouble in the resistive network that develops the bias. You'll find any of several different resistor arrangements, but all of them are simple voltage dividers. You should have no trouble checking the resistors with your ohmmeter.

If bias is okay, the trouble is likely inside the transistor. Either it is leaky, allowing too much current to flow, or some trouble in the base-emitter junction is not letting the bias control the transistor as it should. Again, you can make the "jumper" test mentioned earlier: while measuring the collector voltage, connect the base terminal to the emitter terminal. The abnormally low voltage at the collector should suddenly jump up to the power-.

supply voltage. If it doesn't, the base circuit is not controlling collector current as it should. With zero bias, which is what you have when you short the base to the emitter, very little collector current should flow—only as much as is permitted by leakage in the transistor. With almost no collector current flowing, the collector voltage should be almost the same

fig. 4. Wide spacing of the collector lead is one clue to finding connections. The metal case of a flange-mounted power transistor is usually the collector connection. There are other configurations, but these are the most common.



as the voltage at the source end of the load resistor (no drop across the resistor).

Finally, suppose you've decided the transistor is faulty and you want a final double-check. You can get that with your ohmmeter. A transistor tester is handy, but your ohmmeter is adequate if you know how to use it. The secret of checking transistors with an ohmmeter lies in measuring the backward and forward resistances of the junctions in the transistor. What you should find is shown in figs. **3A and 3B. A** PNP transistor is shown at **A** and an NPN transistor at B. As indicated by the diagrams, the ratios of forward-to-backward resistances are more important than any specific values. The readings in one direction must be much higher than in the other. For small-signal transistors (they are physically small, too), the ratio should be 500:1 or better. In power transistors (the large ones with metal flanges), the ratio can be as low as 100:1, and occasionally even lower. When you're in doubt, check the readings of a suspected transistor against a new one of a similar type.

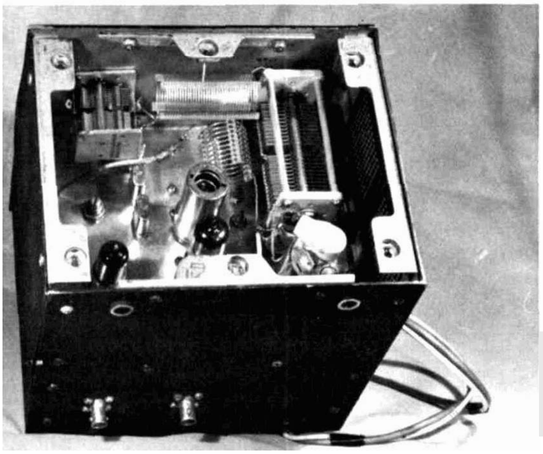
The mechanics of the test are simple. Connect the ground lead of your ohmmeter to the base; then with the other ohmmeter lead, check the reading from base to emitter and then from base to collector. Reverse the leads by touching the probe to the base and use the ground lead to check first the emitter and then the collector. If you use a small chart like those in **fig. 3,** which you can draw on any scrap of paper, you can jot down the readings and then compare them. This test method applies to either NPN or PNP, although the highs and lows are exactly opposite. Nevertheless, **ratio** is what is important; in either type of transistor, the ratios should be high.

To wind up this month's column on "quickie" transistor troubleshooting, I've included some base diagrams of several common transistor types in **fig. 4.** You'll find them easy to memorize, but you can also sketch them on a piece of cardboard and post them on the wall of your shack or on the back of your workbench.

Rarely will you find a transistor that varies from those shown in the diagrams. If you happen to have a piece of equipment with one that does vary, the manufacturer's service data that comes with the unit will show the proper lead configuration. Be very sure you are using the right configuration, particularly when you make the base-emitter "jumper" test. Remember, shorting the base to the collector, even accidentally, can ruin the transistor.

It is obviously impossible to cover all transistor troubleshooting possibilities in one short column. However, if you have specific questions about troubleshooting transistor equipment, drop me a line. Or, if you use some special technique in tracking down trouble in your own transistor gear, tell me about that. I'll use some of the more interesting and helpful ideas in future columns.

ham radio



high-level high-frequency transverter

The megahertz mover– a simple method for extending the frequency coverage of ssb transceivers

John Stanley, K4ERO, 1501 Robin Road, Maryville, Tennessee

Some amateurs are still using a-m on the MARS frequencies and ham bands because their gear only covers one band; many old sideband rigs get dusty on the shelf because they don't cover 15 or 10 meters. Recently, Heathkit recognized the demand for multi-band operation and added an all-band rig to the low-priced HW series of transceivers (the HW-100).

The **megahertz mover** will make the limited-coverage rig as versatile as the builder wants. I use the unit described here to put a HW-32 on 75, 40 and 15 meters, but you can use it on any frequency between 2 and 30 MHz with a simple crystal and coil change.

For those of you who tend to shy away from ssb homebrewing, let me point out that the **megahertz mover** is not any more exotic than a receiving-type mixer or single-tube linear amplifier. There are no balancccl modulators, sideband filters, or phase-shift networks; all of the complex circuits arc in the basic transceiver. In addition, tuneup is simple and requires no special test equipment.

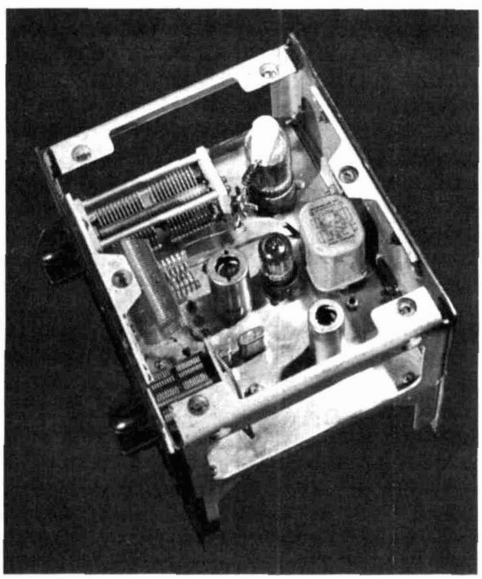
High-level transverting is not new. It's been used to put ssb on the vhf bands for years, and has even seen some use on the lower bands[1] However, the full potential of this approach has not been used Take a look at the schematic and see if this isn't the way for you to put a high-quality ssb signal on a new frequency with a minimum outlay of time and green stuff

the circuit

The transmitting mixer consists of a screen driven 5763, driven with reduced transceiver output. Output can be reduced several ways. I run my HW-32 with reduced plate voltage and with the audio gain turned down. If you are a purist for linearity and carrier suppression, you can use an attenuating network. However, scope and on-the-air testing indicate the method I use produces excellent results—with a saving in dc input power.

The output of the 6CL6 oscillator-multiplier is coupled into the 5763 grid. The exact frequencies involved depend on the transceiver you use and the frequency coverage you want. I used crystals on 10.350, 10.750 and 11.866 MHz to provide coverage of the top

Layout of the megahertz mover. The loading capacitor is in the lower left, the tuning capacitor at the upper left; two of the crystals can be seen in the foreground.

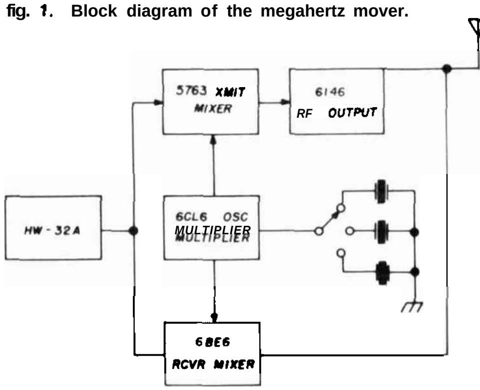


150 kHz on 75, 40 and 15 meters respectively To cover the MARS channels on the low end of 40 meters without retuning, plug in a 10 650-MHz crystal

The 6146 final amplifier is operated in class AB_1 and uses the output circuit popularized by the "Sideband Package"[2] This avoids any bandswitching in the final If you have enough contacts available on the bandswitch of course, you can use a conventional pi network

The receiver section of the transverter consists of a simple 6BE6 converter This has proven satisfactory up through 15 meters, al-

fig. 1. Block diagram of the megahertz mover.

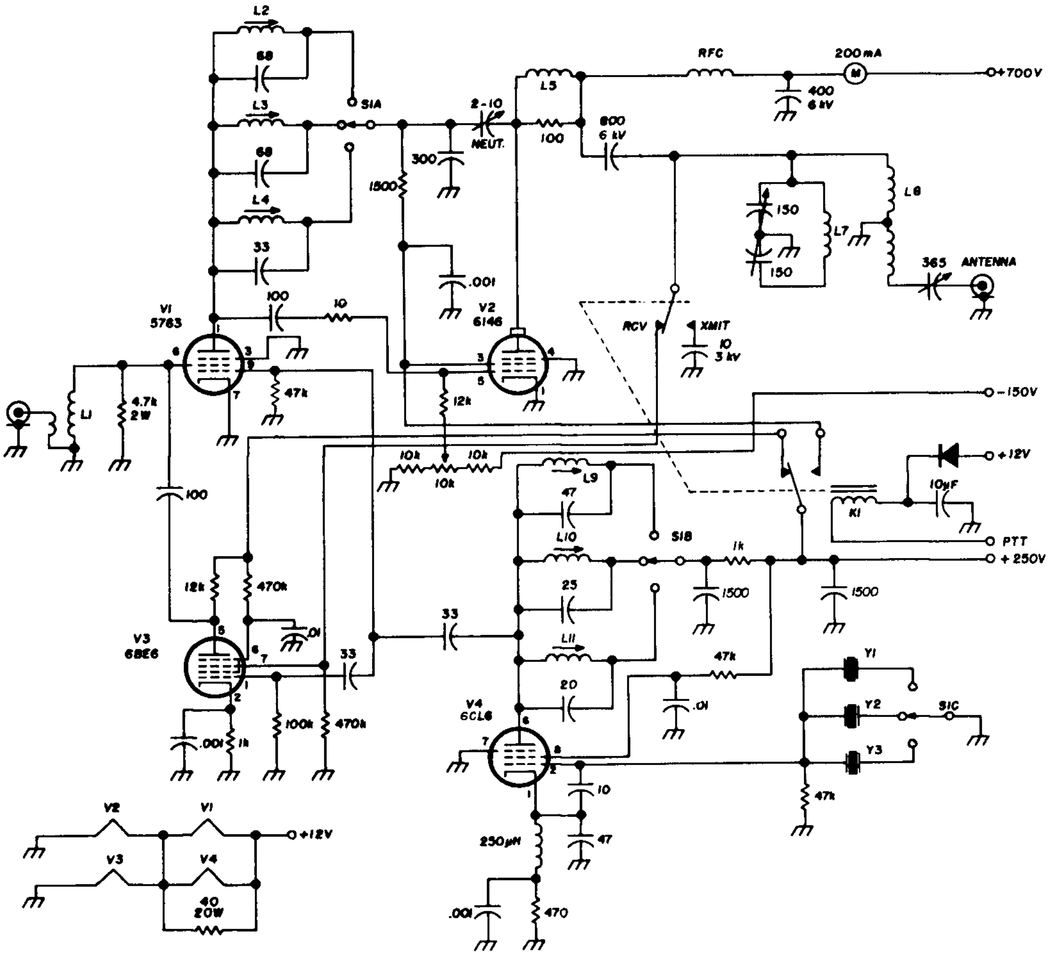


though a good rf amplifier might be useful if 10-meter coverage is desired. The rf stage would improve noise figure and sensitivity on the higher frequencies; on the lower bands, the extra gain would be detrimental to overall performance.

Since the input to the receiver grid is picked off the final tank circuit, peaking the receiver for a given frequency automatically tunes the final. Bandchanging entails throwing the bandswitch to the desired band and peaking the final and loading capacitors—instant QSY!

mechanical considerations

Parts layout is not critical. I used a surplus cabinet because I found one about the right size. An aluminum box about eight inches on each side provides adequate space. Some of you home-brewers who are gung-ho for miniaturization could get it in half the space. The only front panel controls are the bandswitch, plate tuning and antenna loading. I also used a surplus relay, but there are many



K1 12-volt DPDT relay

L1 Resonate to 14 MHz with tube in socket. Input link is 1/4 turn number 18.

L2 4 MHz. 90 turns number 34 on 1/4″ diameter slug-tuned coil form.

L3 7 MHz. 20 **turns** number 26 on 1/4″ diameter slug-tuned coil form.

L4 21 MHz. 9 turns number 22 on 1/4″ diameter slug-tuned coil form.

L5 4 turns number 18 on a 100-ohm, 2-watt resistor.

L7 9 turns number 18, 1″ diameter, 8 TPI **(B&W** 3014).

L8 21 turns number 20, 1″ diameter, 16 TPI **(B&W** 3015). Link is 14 turns of same on cold end.

L9 10 MHz. 20 turns number 24 on 1/4″ diameter slug-tuned coil form.

L10 20 MHz. 8 turns number 22 on 1/4″ **diameter** slug-tuned coil form.

L11 35.6 MHz. 6 turns number 22 on 1/4″ diameter slug-tuned coil form.

S1 **3** pole, 3 position rotary switch.

Y1 **10.35-MHz crystal**

Y2 **10.750-MHz crystal**

Y3 11.866-MHz crystal

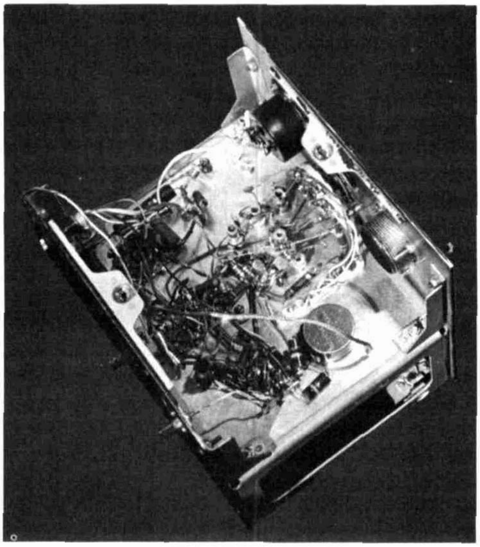
fig. 2. Schematic **diagram** of the megahertz mover—a high-level transverter that may be used on any frequency between 2 and 30 MHz with the proper crystal and coil changes. The unit is shown here for 3.5, 7 and 21 MHz output with 14-MHz drive.

commercial equivalents. On the rear of the cabinet are two BNC connectors (antenna and transceiver), the bias pot and the power cable.

power supply

In most cases, the voltages for operating the megahertz mover can he taken from the transceiver supply. If you have any doubt about the capacity of your supply, a separate supply providing 12 volts at 1.5 amps and 250 volts at 100 mA should be used. The bias voltage from the transceiver supply can be used in almost every case.

Below chassis construction of the megahertz mover.



If you lower the plate voltage or current in the transceiver when using the megahertz mover, the high-voltage supply can also be used for the 6146. If you want, you can add a power switch to the transceiver power supply so you can switch instantly from transvert to straight-through operation.

tune-up

To put the transverter into operation, all you have to do is peak the coils. First, adjust L1 for resonance with a grid-dip meter or by listening to signals coming through the unit. Tune the final to peak the receiver; then peak the oscillator-multiplier for the band being used.

After the receiver is tuned up, go to transmit and set the bias for about 15 mA of plate current. Then apply a test tone, preferably a two-tone, to the mike input and feed the output into a dummy load. Tune the driver coil for maximum output. The final should already be tuned from the receiver tune up. Neutralization follows standard procedure. Finally, peak up all the coils, especially the oscillator-multiplier coils, and check linearity with a scope.

general considerations

I used the 20-meter Heathkit HW-32 for coverage on 75, 40 and 15 meters. If you use another transceiver to cover some other frequency range, you'll have to work out the crystal frequencies. When doing this, there are several factors you should keep in mind:

1. In general, it's better to use a mixing frequency above the band you want to prevent birdies caused by crystal harmonics. You'll notice that I mix 10 MHz with 14 MHz to reach 4 MHz rather than using a 6-MHz crystal/tripler: 21 MHz injection is used for the 7 MHz range and a 35-MHz signal for the 21-MHz band. In the latter case it's quite obvious if a 7-MHz crystal were used with a 14-MHz transceiver to cover the 15-meter band, you'd have a lot of trouble with birdies.

2. Because of the birdie and doubling problem, you can't cover a band that is an exact multiple of the band you start with. I couldn't cover 28 MHz with the 14-MHz HW-32 because the second harmonic of 14 MHz would give me an extra signal on 28 MHz.
(Experimentally-minded amateurs might try it by using a carefully designed balanced or hybrid modulator.)

3. If your transceiver only has one sideband, use extra care when you mix frequencies, because some combinations switch the sideband. If the mix frequency is above the transceiver frequency **and** resultant frequency, the sideband will be reversed; other-

See Page 85

table 1. Crystal mixing frequencies.

transceiver freq	output-freq	mixer freq
14 MHz USB	4 MHz USB	10 MHz
14 MHz USB	4 MHz LSB	18 MHz
14 MHz USB	7 MHz LSB	21 MHz
14 MHz USB	21 MHz LSB	35 MHz
14 MHz	any MARS frequency	frequency ± 14 MHz
7 MHz LSB	4 MHz USB	13 MHz
7 MHz LSB	4 MHz LSB	3 MHz
7 MHz	any MARS frequency	frequency + 7 MHz
4 MHz LSB	7 MHz USB	13 MHz
4 MHz LSB	7 MHz LSB	3 MHz
4 MHz LSB	14 MHz USB	18 MHz
4 MHz LSB	21 MHz USB	25 MHz
4 MHz LSB	28 MHz USB	32 MHz
4 MHz LSB	50 MHz USB	54 MHz
4 MHz	any MARS frequency	frequency + 4 MHz
50 MHz USB	4 MHz LSB	54 MHz
50 MHz USB	7 MHz LSB	57 MHz
50 MHz USB	any frequency LSB	frequency + 50 MHz

Exact frequencies are not given since they must be calculated for each particular situation and depend upon the frequency range of the transceiver and the coverage desired. Note that on some bands the dial calibration as well as the sideband is reversed.

wise, the output from the transverter will be the same as the transceiver. This consideration may over-ride number 1 above. It's better to have a slight birdie in the receiver than to be on the wrong sideband. A table of suggested frequencies is worked out in table 1.

I have used the megahertz mover for nearly a year and have had excellent reports on all bands; the quality is essentially that of the transceiver. Power output is down slightly from the 200 watts PEP of the HW-32, but not enough to affect performance. If you are a power nut, put a pair of 6146's in the unit or drive a big tube.

References

1 "High-Level Converters," *SSB for the Radio Amateur*, ARRL, 3rd Edition, 1962, p 196
2 G K Bigler, W6TEU, "A Stdcband Parkagr," *QST*, June, 1958, p 24

ham radio

the ham notebook

six-meter tunnel-diode phone rig

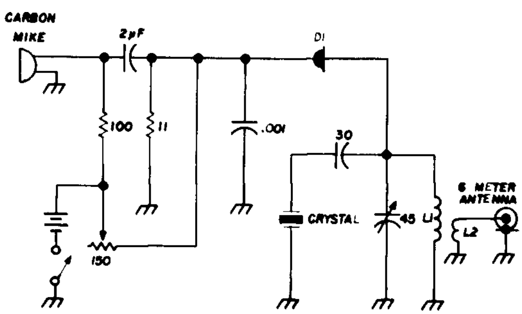
Those of you who would like to try your hand at flea-power tunnel-diode hamming should get a big kick out of the rig shown in fig. 1. With a tunnel diode, the little transmitter can be built into the tiniest housing, yet still provide adequate rf output to make local contacts.

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 appoximately 0.20 volts. At a bias voltage of 0.18, you'll notice a slight voltage upsurge which corresponds to an audible blast from a monitor receiver tuned to the transmit frequency. Adjust so that the TD 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/W9HBF**

fig. 1. Simple six-meter TD transmitter. Crystal is a 26-MHz overtone type; output can be coupled to any 50-ohm coaxial line. The battery is a 1.35 Vdc hearing-aid type. D1 is a tunnel diode such as the 1N3714.



10-minute timer from a Heathkit CA-1

A lot of amateurs have an old Heathkit CA-1 Conelrad alarm laying around that they don't need any more. Here's a way to convert it to a handy ten-minute timer with a diode, a capacitor, a resistor and a 6AK5. For the conversion, remove the chassis from the cabinet—you'll find a two-lug terminal strip near the ac line cord. Unsolder the resistor going to pin 1 of the tube socket and remove the terminal strip, resistor, rubber grommet and shielded cable.

Install a new terminal strip, connect the cathode end of a 500-mA, 400-PIV silicon diode to the anode of the old diode and the other end (anode) to the new terminal strip. Run a wire from the other lug on the terminal strip to pin 1 of the tube socket and connect a 20-μF capacitor between pin 1 and ground (positive to ground). Also connect a 4.2-niegohm resistor between pin 1 and ground.

Now, take both wires off the pushbutton switch and short them together. If you want, you can remove the short wire completely and connect the long wire directly to the relay—this accomplishes the same thing.

Connect two new wires to the normally-open contacts on the switch and run them through a grommet to the two lugs on the new terminal strip. Rewire the tube socket to accomodate the 6AK5; this is a simple exchange of the wires going to pins 5 and 6 on the socket; make sure that pin five is not shorted to pin seven. The diagram in *fig.* 2 shows all the details. This completes the conversion.

Turn the unit on. After it warms up a bit, the 6AK5 will turn on, pick up the relay and turn on the light. Push the switch. This applies –150 volts to the grid of the 6AK5 and turns it off—it also charges the 20-μF capacitor. When the capacitor is discharged by the 4.2-megohm resistor, the 6AK5 will turn on again and pick up the relay; this should take about ten minutes. You can make fine tuning adjustments by adjusting the tension on the armature spring on the relay. The ac socket on the front panel can be used as a power source to ring a bell or turn on another light; just replace the fuse with an appropriate value. My unit is timed at 9 minutes, 45 seconds; this gives me plenty of time to identify. I've used it for over a year and it hasn't changed a second.

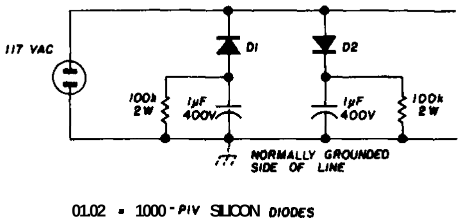
Harold Mohr, K8ZHZ

line transient protection

If you're troubled with line transients blowing out solid-state components, *fig.* 3 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. 3. Simple circuit for suppressing voltage transients on the ac line.

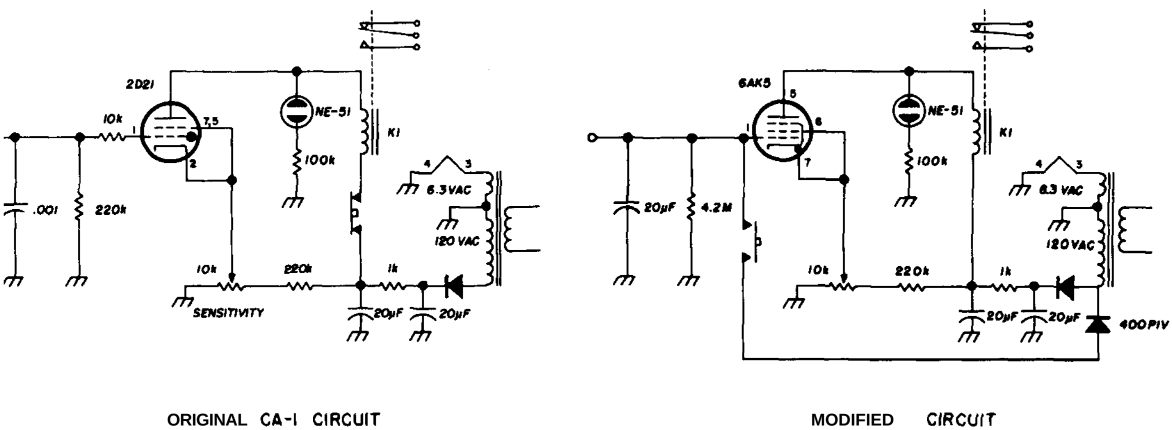


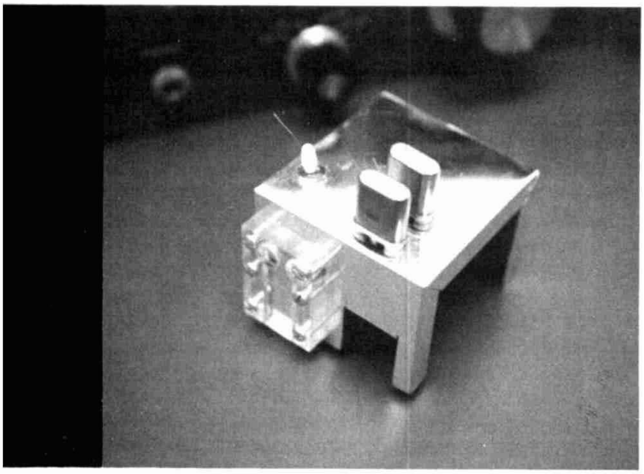
deburring holes

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.

W2DXH

fig. 2. Original Heathkit Conelrad Monitor circuit and its conversion to a 10-minute timer.





rtty reception

with the SB-300

Here's a way

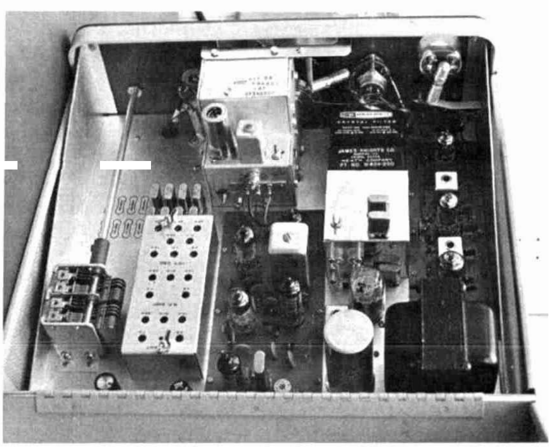
to receive RTTY

on the Heathkit SB-300

without any modifications

Maury Cox, W2ARZ, 3716 Frazier Road, Endwel New York 13760

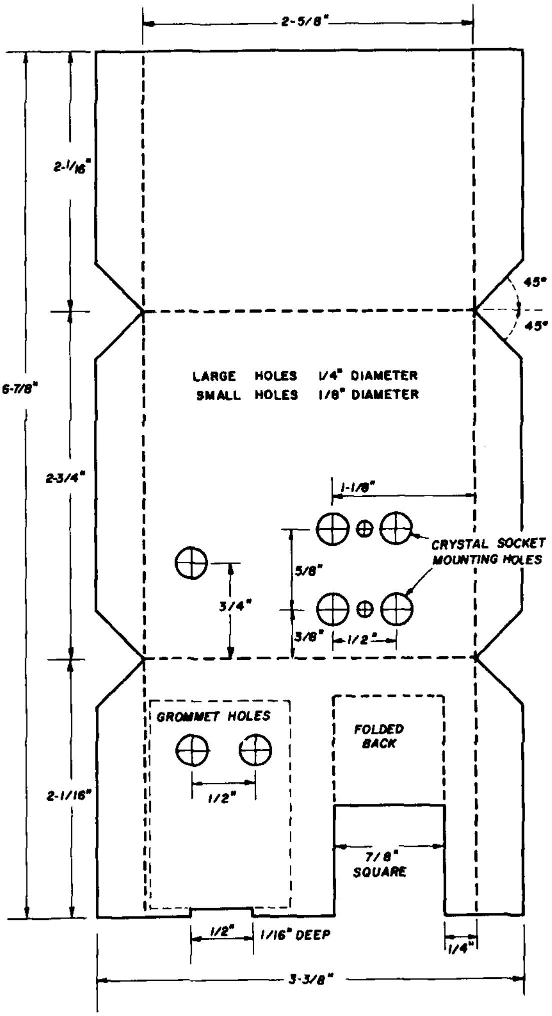
If you own a Heathkit SB-300 and are interested in RTTY reception, here is a handy-dandy little **gizmo** that effectively makes it into a model 301 as far as RTTY is concerned. What's more, no modification to the basic receiver is necessary.



Installation of the RTTY adapter in the SB-300 receiver.

The **gizmo** is an inexpensive adapter that is almost easier to build than to describe. It provides a means of switching (in USB mode) between the upper-sideband crystal and an added RTTY crystal in the bfo. The additional crystal puts the 850-Hz RTTY signal (differ-

fig. 1. Layout of aluminum sheet used to make up the RTTY adapter chassis.



ence between the mark, 2125, and space, 2975) within the bandpass of the ssb filter. You can buy the crystal you need from Heath*.

If you use the SB-300 with a transmitter for transceive operation, a word of caution: the bfo amplifier (V9C) should be disabled or disconnected during reception to prevent accidental transceive operation. In the SB-301, pins 11 and 12 of mode switch (MS3R) disable the amplifier in the RTTY mode by removing B-plus from the plate of V9C. If you use independent transmit-receive operation, there's no problem, and you can forget the bfo amplifier.

The RTTY adapter is simply an open aluminum box with two crystal sockets, the upper-sideband and RTTY crystals, and a miniature toggle switch. A block of clear plastic holds two pieces of number 16 bus wire that simulate the crystal contact pins. The adapter fits over the crystal filter nearest the bfo crystals and plugs into the upper-sideband crystal socket. The upper-sideband and new RTTY crystal are plugged into the two sockets on top of the adapter. Installation of the RTTY adapter is clearly shown in the photograph.

construction

The adapter is made in two pieces, and construction is quite simple. The two pieces—the aluminum box and the plastic block—are bolted together, the wires connected, and that's all there is to it! Except for the 112-inch spacing beween contact pins in the plastic block, there is nothing critical about any of the dimensions. Almost any available materials and hardware can be used.

To make the box, cut a rectangular piece of thin aluminum sheet 3-318-inches wide and 6-7/8 inches long as shown in **fig.1**. Cut the four 90" corner notches and the 1/2- by 1/16-inch clearance notch at bottom. You may prefer to drill all the holes for mounting the hardware now, or you can wait until the box is formed into its finished shape. Just follow the hole layout shown in **fig. 1**. Not shown on the drawing are the four holes used to support the plastic block.

Make the box by folding the aluminum 90" along the dashed lines in **fig. 1**. Cut out the 718-inch square piece shown in **fig. 1**, or fold it back inside the box so the box doesn't interfere with the lower-sideband crystal when it's plugged in. Mount the two crystal sockets,

* **Haath Company, Benton Harbor, Michigan 49022. 3392.110 kHz crystal, part number 404-280, $5.00**

toggle switch and two rubber grommets.

plastic block

Cut out a rectangular piece about 1-118-inches wide by I-518-inches high from a piece of plastic about 5/16 inches thick. The plastic doesn't have to be clear, but if it is, it will be a big help. Drill two parallel holes on 112-inch centers to a depth of 1-3116-inch as shown in

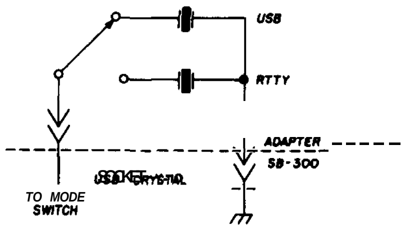


fig. 2. Wiring diagram of the RTTY adapter.

fig. 2. Locate these holes symmetrically with respect to the block centerlines. (Use a number-53 drill for number-16 tinned wire.)

Drill two 318-inch holes at right angles to the inner ends of these holes, so that the tip of the drill goes about 1116-inch beyond the wire holes. These 318-holes clear the grommets on the side of the box; the 1116-inch extra depth provides room for attaching the hookup wire to the wire "pins." Now drill four (or six, if desired) mounting holes. I used a number-33 drill for **4-40** screws.

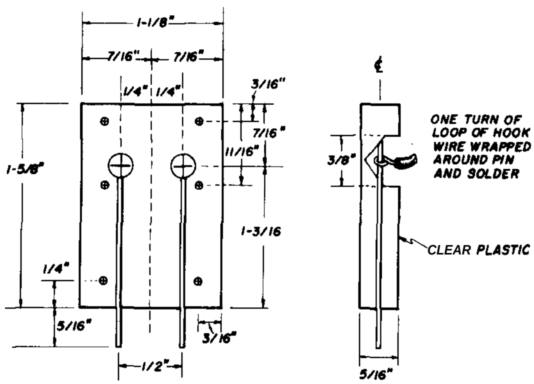
Cut a couple of ***straight*** pieces of number-16 tinned wire about I-314-inches long, and insert them **as *far as they*** will ***go*** into the wire holes. Remove about 318-inch of insulation from two 2-inch lengths of hook-up wire, loop one turn around the upper end of each of the tinned-wire pins, and solder. This completes the block.

Fig. 2 shows the details. This is the trickiest part of the whole operation because you're working in close quarters; a crochet needle or spring hook will prove invaluable at this point. Also, although there is no functional requirement for clear plastic, its transparency is a definite asset and helps construction.

assembly and wiring

Push the free ends of the two pieces of hook-up wire through the grommets, and bolt the plastic blocks to the face of the aluminum box so the bottom of the block is flush with the top of the 1/16-inch notch. Connect the two wires to the crystals and switch as shown in fig. 3. Be sure to connect the common (ground) pins of the two crystal sockets in the RTTY adapter so that they are connected to the common pin of the upper-sideband crystal

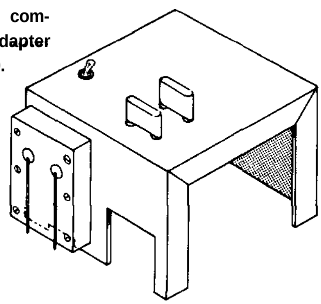
fig. 3. Plug assembly which goes into the upper-sideband crystal socket of the SB-300.



socket (Y-10) in the SB-300 when the adapter is in place.

You may also want to label the settings of the toggle switch on the adapter so you won't forget which is upper sideband and which is

fig. 4. The completed RTTY adapter for the SB-300.



RTTY. Plug in the unit, and you're ready to go. This adapter, of course, does not eliminate the need for an RTTY terminal unit. The TU may be connected to either the speaker or anti-vox output jacks, depending upon the impedance that is required.

ham radio

propagation predictions for july

High-frequency forecasts for the month of July plus some predictions for multi-hop sporadic-E propagation on 6 meters

R. Frank, 6KAP, 12450 Sk l . Boulevard, Woodside, California 940

The primary emphasis of this column will be on long-haul high-frequency communications over distances greater than 2500 miles (4000 kilometers). Over paths of this length, hf communications are dependent upon the F2 layer. The predictions will be as general as possible — applicable to almost any situation that the DX'er will run into.

Basically, these predictions will try to answer two questions: is the band open, and how far? At first, the format for the predictions may seem unusual. However, with the graphical presentations I have used here, they are much more versatile than a tabular form. Also, the graphs provide some insight to the underlying causes of propagation (or lack of it) over a given path on a certain band. It also permits mental "fudging" to allow for day-to-day variations in ionospheric conditions and shows the affect of geographic latitude on the maximum usable frequency (MUF).

maximum usable frequency

The propagation predictions are in two parts. The first part shows the maximum usable frequency for various latitudes in the northern hemisphere. The maximum usable frequency in this case is defined as the highest frequency that will be returned to the earth by the F2 layer in a given direction. The distance at which the signal will be returned by a single hop is about 2500 miles or 4000 kilometers. The reflection point is in the middle of the first hop — about 2000 kilometers from the end. This point is called the **control point** because its MUF determines the highest propagating frequency in that particular direction by the F2 layer. Basically, then, for long-haul communications, we're interested in the condition of the ionosphere about 2000 kilometers or 1200 miles away from the transmitter.

For communications over distances greater than 4000 kilometers by way of the F2 layer, a signal travels by successive hops between the earth and the ionosphere. If the MUF falls below the operating frequency, the signal may not be returned to earth, but will propagate by internal reflection in the ionosphere until it reaches a region where the MUF is high enough to return it to the earth. Therefore, there may be extended skip zones beyond the first hop—especially over the poles and temperate latitudes in advance of the dawn line.

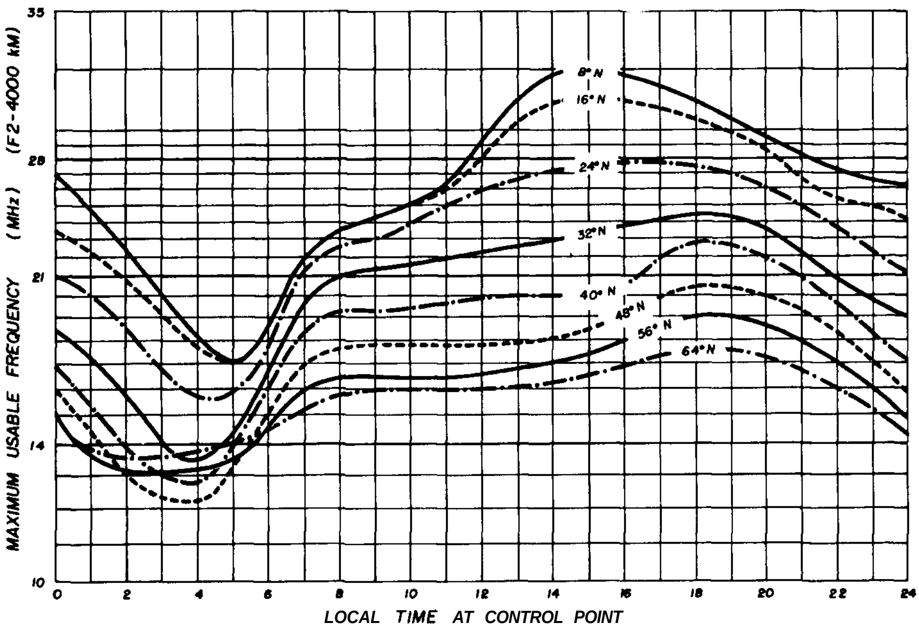
These skip zones are not forecasted on the basis of a single control point. For any long path, there are two control points, 2000 kilometers in from each end. Each will have a MUF that depends on latitude and local time, and to some extent, longitude. The two-control-point method forecasts that the overall path MUF will be the **lesser** of the two control-point MUF's. There is some evidence that during the summer months, sporadic-E propagation at the end of the path with the lowest MUF may raise the path MUF to the greater of the two control-point MUF's.

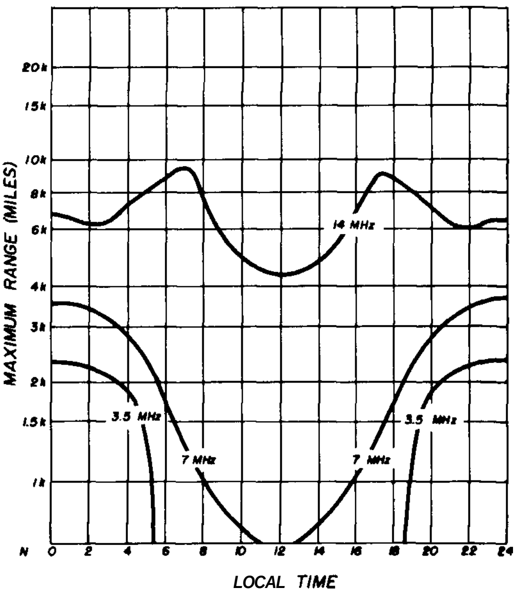
The MUF curves shown here are properly referred to as MUF **(4000) F2.** This simply means the maximum usable frequency at which the F2 layer will provide communications over a 4000-kilometer path. If you're confused by the use of kilometers, just remember that 4000 kilometers is approximately 2500 miles. Kilometers are used here because this is the standard nomenclature used in ionospheric propagation discussions and in propagation information available from the Institute of Telecommunications Sciences, Boulder, Colorado.

maximum possible range

The second part of the propagation predictions concerns the maximum possible range; this is determined by absorption. Absorption is a function of both the solar zenith angle and the angle of incidence at each penetration of the D layer, but will vary somewhat with atmospheric noise and the equipment

fig. 1. Maximum usable frequency curves for July 1968 baaed on 75° W longitude. For information on how to use these curves, refer to the box on page 84.





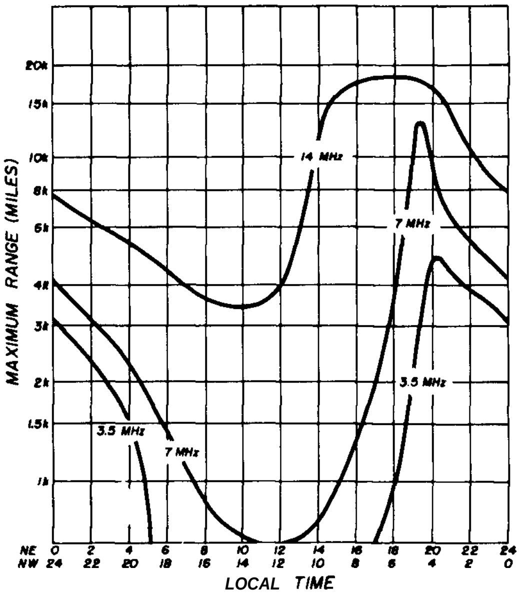
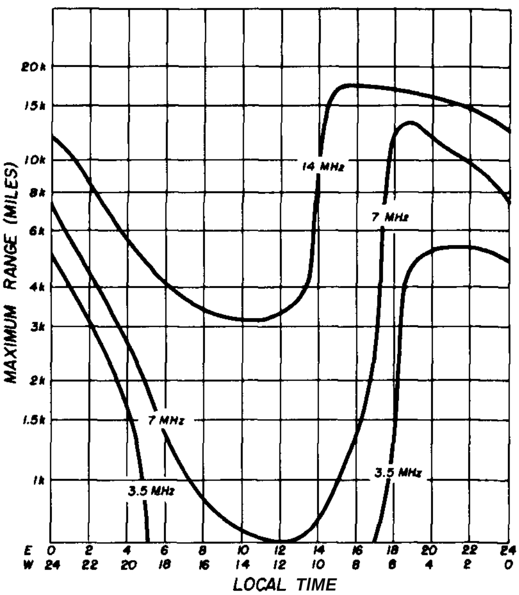


fig. 2. Maximum range to the north from **38°** N **latitude** due to absorption.

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



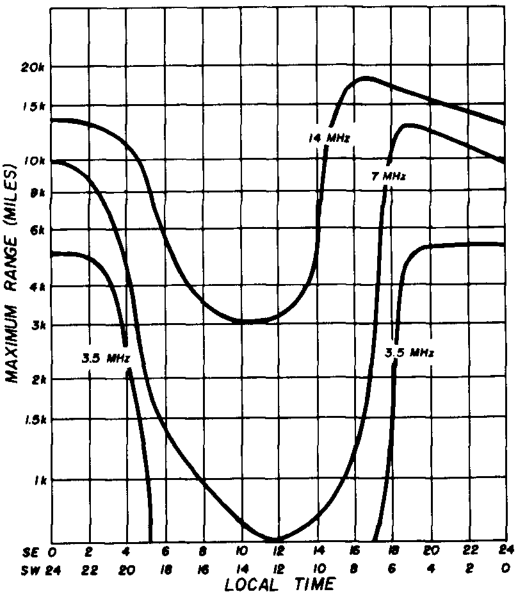


fig. 4. Maximum range to the east (top time scale) and to the west (bottom time scale) from **38°** N **latitude** due to absorption.

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

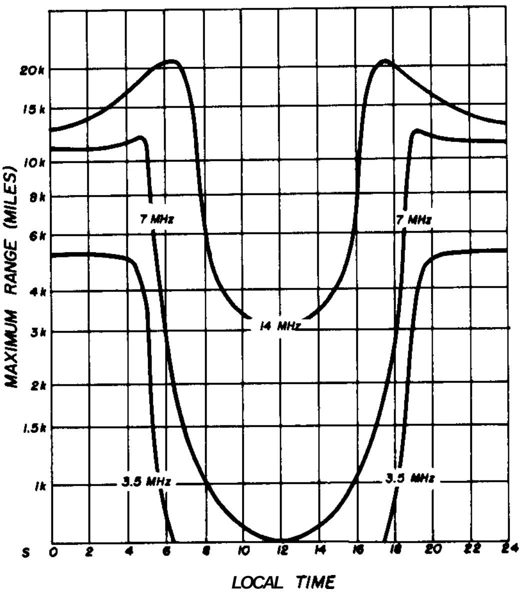


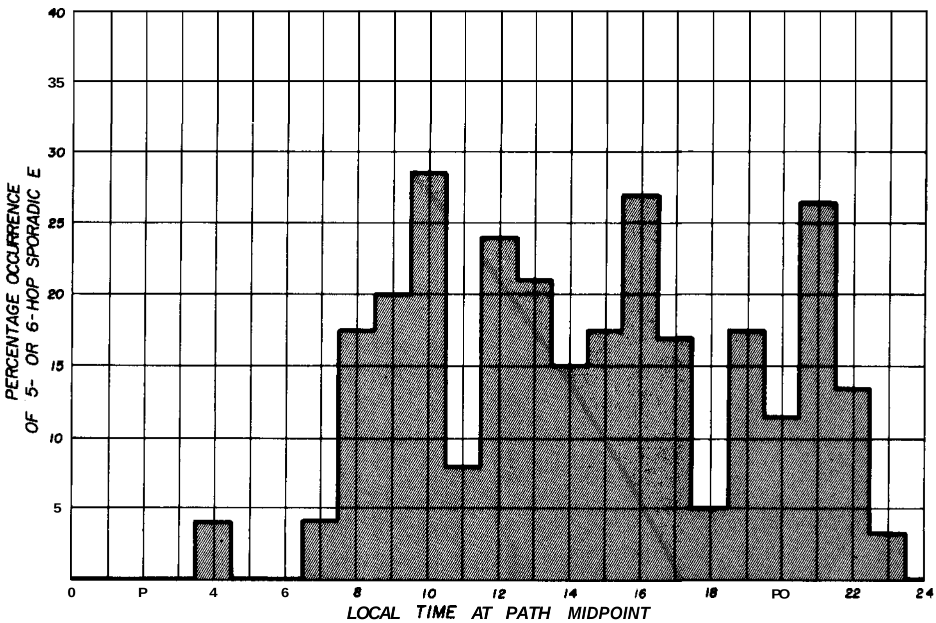
fig. 8. Maximum range to the south from 38° N latitude due to absorption.

you use. For example, a 10-dB increase in transmitter power or antenna gain above the values given in the box on page 84 will increase the opening time to any particular range by about one hour.

sporadic-E propagation in July

Past performance has shown that July is one of the best months for multiple-hop sporadic-E propagation. How far have you worked using this mode? Few 50-MHz stations have worked further than three hops or 3700 miles. However, this is probably due to the lack of activity in the proper places rather than a lack of propagation. When the Pacific Ionospheric Scatter network was in operation a few years ago, signals near 50 MHz from Wake Island were often heard on the West Coast, 4350 miles away. Perhaps more important, an ionospheric sounder operating over a 6200-mile (10,000-kM) east-west path lying below 50" N latitude observed 5- or 6-hop sporadic-C propagation at frequencies in excess of 20 MHz nearly 60% of the days of July 1966; on 28% of the days of July 1966, sporadic-E propagation was observed on frequencies greater than 30 MHz over the same path. This is plotted in fig. 7 with time of day at the path midpoint.

fig. 7. Percentage occurrence of multiple-hop sporadic-E propagation at frequencies greater than 30 MHz over a 6200-mile (10,000-kilometer) east-west path in the northern hemisphere during July 1966.



how to use these propagation charts

1. To find the maximum usable frequency **(F2-4000 kM)** in any particular direction from your **location, find the latitude of the control point from** table **1.** The control point will **be 1200** miles **(2000** kilometers) from your location.

The curves in fig. **1.** show the MUF for the latitude and local time of the control point. Since the control point **is 1200** miles away, **local time there is 45 to 90 minutes later than your local time if it is to the east, and 45 to 90 minutes earlier if it's to the west. Unless your station is located in the middle of a local time** zone, the standard time for your area is close enough for these calculations. Remember that **standard time is the time at 75° W (EST), 90° W (CST), 105° W (MST) and 120° W (PST). For** accurate time at your location, add four minutes per degree longitude west of the longitude **which determines the time zone for your area.**

Example: Your station is located at 34° N, you want to work east (90° beam heading) or west (270° beam heading). What would be the best operating times on **15** meters?

First, find the latitude of the control point from table **1—32°** N. From the MUF curve, you can see that **21** MHz will be open for distances **2500** miles **(4000** kilometers) and **beyond between 0800** and **2200** hours control-point time. The band **will be open to the east between 0630 and 0230 hours, and to the west between 0930 and 2330** hours local time.

2. To find the maximum propagation distance because of absorption, refer to fig. **2, 3, 4, 5** or **6,** depending on the direction you want to work. Note that the time scales are reversed for **west**ward propagation in fig. **3, 4** and **5.** These **curves are** based on unity signal-to-noise **ratio in a 6-kHz bandwidth with 100 watts output** power **(100** watts CW or **800** watts ssb), with combined receiver and transmitter antenna gains of **−12** dB on **3.5** MHz, zero dB on **7** MHz, and **+12 dB on 14 MHz. On ten and fifteen meters,** the communications range should not be limited by absorption to **less** than one transit around **the earth. However, anytime you expect minimum range on 14** MHz, round-the-world propagation **will be minimal on 21 MHz.**

3. To find the MUF for a particular path in the northern hemisphere, locate the other station's control point. Remember that it is **1200** miles **(2000** kilometers) toward you. The MUF curve **may then be used to make a crude approximation** of his control-point MUF. The path MUF is **the** lower of the two control-point **MUF's—yours** and his. These curves **are** not useful for the southern hemisphere.

The MUF curves should be accurate within a couple MHz between **45°** and **135°** west **longitude. They were prepared from basic propagation** predictions published monthly by the **Institute for Telecommunications** Sciences **(ITS), Boulder Colorado and available through the** U. S. Government Printing Office. The maximum distance curves were derived from standard formulas at 1000-mile intervals in each of **8** directions from **38°** N latitude.

One fact worth noting is that the amplitude of multi-hop sporadic-E modes was always lower than that of the multi-hop F2-layer modes, when present on the same frequency. In many cases, the difference was more than 20 dB.

table **1.** Control-point latitudes (degrees **N).**

your latitude (degrees **N)**	direction N	NE/NW	E/W	SE/SW	S
24	42	36	23	10	6
26	44	38	25	13	8
28	46	40	27	15	10
30	48	42	28	17	12
32	50	44	30	19	14
34	52	45	32	20	16
36	54	47	34	22	18
38	56	49	36	24	20
40	58	51	38	26	22
42	60	53	40	28	24
44	62	55	41	30	26
46	64	57	43	32	28
48	66	58	45	34	30

propagation summary for July

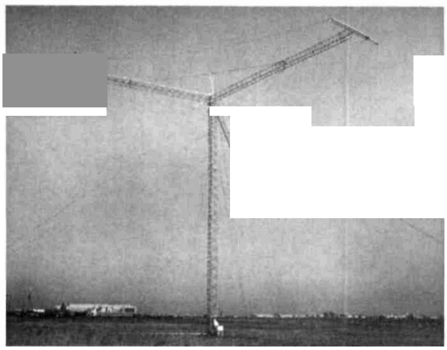
Below 30 **MHz.** High noise levels and absorption will continue to limit propagation on 40 and 80 to darkness paths. Don't expect any propagation over the North Pole on these bands because it is in continuous daylight. **Twenty** meters will continue to be a nighttime band for the DX'er. Look for openings on 10 and **15** during the day, particularly toward the south; short-skip sporadic-E propagation will be prevalent.

***6* meters.** Six should open up to distances between 1000 and 1500 miles more than one-third of the days of the month. Look for multiple-hop sporadic-E openings over paths greater than 3500 miles—if necessary, make schedules. Past experience has shown that the openings are there if you're around to take advantage of them.

ham radio

new products

rotatable log periodic array



Arc you interested in a really big antenna? If you are, try this new design from Granger Associates on for size. This antenna is 110 feet long and has a total width of 100 feet. It has sixteen active radiators and produces a gain of 12 dB and a front to back ratio of 15 dB. The frequency range is from 5.5 to 32 MHz with SWR of 2:1 maximum. Designed for the military market, this antenna has a power capacity of 20 kW average, 40 kW PEP, and will withstand up to 160 mile-per-hour winds with no ice.

The Granger Associates model 1730-3 Rotatable Log Periodic Antenna is complete with rotator, tower, balun, local-and remote-control units and complete assembly and erection instructions. Shipping weight is 9,000 pounds. For more information, write to Granger Associates, 1601 California Avenue, Palo Alto, California 94304.

Triplett solid-state vom

Triplett has just introduced a new battery-operated solid-state volt-ohm-milliammeter using FET's to provide 11 megohms input impedance on all ac and dc voltage ranges. In addition to the high input impedance, the FET circuitry provides improved stability over hattery life and temperature changes.

The Triplett model 601 features 52 range selections, accuracy of ±2% full scale on dc and ±3% on ac, and a low-power ohms circuit for IC measurements. With the low-power ohms scale, 75 millivolts is applied to the device under test, and the maximum power applied is 0.1 milliwatt. The ac frequency response of the model 601 is 50 Hz to 50 kHz.

A 100-dB scale range is incorporated into this new vom for high fidelity and stereo servicing. Using the single selector switch, the user can make measurements from –40 to +60 dB. The dc full-scale ranges go from 0.1 to 1000 volts; the ac scales, 0.01 to 1000 volts; ohms from 1 to 1 megohm, both conventional and low power; and ac/dc current ranges from 10 microamperes to 10 milliamperes. The range selector also includes a battery-condition check position.

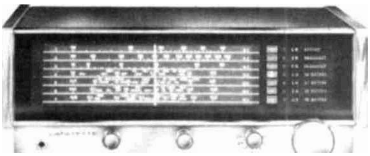
Battery life in normal usage is about one year. Although the model 601 is furnished with a complement of ten AA penlight cells, the instrument is designed so that it can use alkali, mercury, or nickel-cadmium batteries of the AA type.

The model 601 has a brushed aluminum

front panel and etched black range markings. The charcoal-grey case is constructed of high-impact material. A color-coded meter dial provides fast and easy readings. A slotted thumb screw on the back cover provides easy access to the batteries.

Because of the instrument's high sensitivity on ac measurements, the 601 uses a small, pencil-thin shielded probe. Price is $125 complete with hatteries, test probes and operating instructions. Check your local distributor, or write to the Triplett Electrical Instrument Company, Bluffton, Ohio.

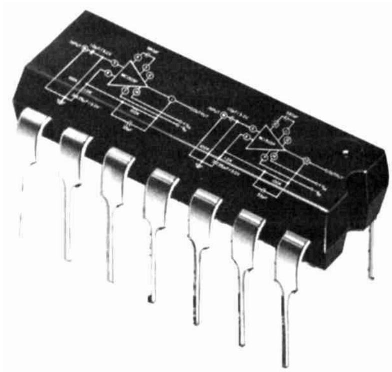
Lafayette 7-band receiver



Lafayette Electronics Corporation has just introduced a new seven-band fm-am, short-wave and weather/marine receiver. This new receiver features individually-tuned circuits for each band, avc, a mechanical filter, tuned rf stage and transformer-operated power supply. The seven bands cover 150-400 kHz, standard a-m and fm broadcast bands, plus the short-wave bands from 5.9-6.25, 9.45-9.85, 11.85-12.05, and 15.05-15.55 MHz.

The illuminated slide-rule dial on this receiver is labeled with the names of primary cities and countries. The rear panel has terminals for a-m and fm antennas, tape recorder output, extension speaker and 24-volt dc input. FM sensitivity is 2 microvolts or less; cross modulation, 70 dB; distortion, 1% or better at 20 microvolts; image ratio, 45 dB; and antenna impedance, 300 ohms. Sensitivity on the short-wave bands is as low as 2 microvolts. Priced at $99.95 from Lafayette Radio Electronics Corporation, 111 Jericho Turnpike, Syosset, Long Island, New York 11791. Order stock number 99-2601WX.

hi-fi stereo preamp ic



Motorola has just introduced a new integrated circuit which is specifically designed for high-fidelity amplification of low-level stereo signals. The MC1303P features a unique short-circuit-proof design which protects the device against accidental shorting in test or installation. The dual-channel preamp provides channel separation of **60** dB minimum at 10 kHz with less than 0.1% total harmonic distortion at the minimum rated output voltage swing of **4.5** volts rms.

The MC1303P data sheet presents application information including recommended equalizing networks that will provide flat frequency response in accordance with RIAA and NAB specifications. In addition, the data sheet has curves which show the noise and output loading characteristics of the device. The input bias current of the MC1303P is 1 μA; input offset current, 0.2 μA; input offset voltage, **1.5** mV; and dc power dissipation, 300 mW maximum. For more information, write to Motorola Semiconductor Products, Inc., Box 13408, Phoenix, Arizona 85002.

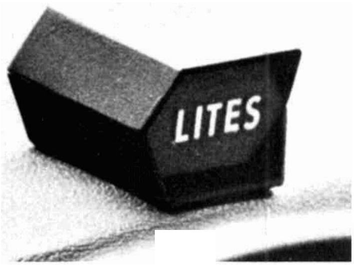
electronic circuit design handbook

How many times have you needed an electronic circuit to do a special job? A circuit you need right now and don't have time to design? If you have, this is a book you ought to have in your library—a compendium of more than 500 proven and tested circuits for all types of functions.

The circuits in this book were originally published in **EEE—the magazine of Circuit Design Engineering.** They were selected on the basis of their originality and practical application. This volume includes both simple and complex circuitry, as well as a lot of practical design data. If you have an application that isn't covered, usually you can find a circuit that can be adapted to suit your needs—they can also serve as stepping stones to almost any kind of circuit desired.

The circuit descriptions are supplemented by over 600 easy-to-follow schematics, diagrams, waveforms and illustrations. The tried and tested circuits constitute a vital source of ideas and techniques, and serve as "imagination triggers" for anyone who has an interest in circuit design and construction. $14.95 from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

headlight warning indicator



Although this may seem like a strange item to find in an amateur radio magazine, I wonder how many of you have gone away and left your car with your headlights or parking lights on. This little indicator flashes when your lights are on and your ignition is off—no more dead batteries! It features simple two-wire installation, solid-state circuitry, negative or positive ground and universal mounting. For all models and makes of cars, trucks and boats. Available for either 6 or 12-volt systems. $2.98 postpaid from Mike Gauthier, K6ICS, Gauthier Industries, P. O. Box 216, Lynwood, California 90262.

amphenol connectors

A new line of blister-packaged rf, audio and power connectors for amateurs and experimenters has been announced by Amphenol. The new packs, presently available at electronic parts distributors, include many hard-to-find connectors needed by project builders and electronics experimenters.

The new line of connectors is the result of two years of intensive in-the-field market research. Interviews with hundreds of buyers plus surveys of sales to experimenters resulted in a line that includes the broadest possible range of hobbyist needs. In addition to microphone and power connectors, the new line includes type-uhf coax connectors, straight, angle and tee adapters, push-on connectors and even in-line lightning arrestors.

Complete assembly information is printed on the reverse side of each blister pack. Detailed step-by-step instructions explain how the component should be assembled and used. Additionally, handy workbench reference booklets are available on each display. These booklets contain soldering information, schematic symbols, and resistor color codes as well as a complete catalog of Amphenol blister-pack components. Look for the new display at your distributor, or write to the Amphenol Distributor Division, 2875 South 25th Avenue, Broadview, Illinois 60153.

Motorola power amplifier/ switching circuits library

Motorola has just come out with a new technical reference tool aimed at the electronics designer who needs practical, comprehensive information about silicon power transistors in switching and amplifier designs. This book has 65 circuits that are devoted to saving the engineer time in design. It is highlighted by easy-to-understand, how-to-do-it topics like direct coupling to eliminate amplifier transformers, outguessing the effects of secondary breakdown transients, square wave forms, pulses, an exhaustive examination of ac-to-dc converters, FET's and high-voltage silicon power in entertainment and communications equipment and designing short-circuit-proof voltage regulators, just to name a few. This volume also includes a silicon power transistor selection guide which covers over 100 plastic and metal types of transistors spanning collector currents from 100 milliamps to 30 amps, collector-emitter voltages to 400 volts, power-dissipation ratings to 200 watts of f_T's up to 60 MHz. Copies are available by writing on your company letterhead to Motorola Semiconductor Inc., Box 955, Phoenix, Arizona 85001.

The Electronic Invasion

Robert M. Brown, K2ZSQ, has authored a new book on electronic bugging and debugging that should be of interest to every amateur. This noted electronics expert, amateur, author, vhf'er and licensed private investigator reveals the secret techniques, devices, circuits, users and suppliers of bugs and anti-bugs in this new book. Bob takes you behind the scenes in electronic snooping and tells you how the experts do it. He shows you the circuits they use, and some of the gambits they follow when duping the poor victim. He covers telephone bugs, eavesdropping microphones, miniature audio amplifiers, fm wireless microphones, bumper beepers, recording spies, bug detection, speech scramblers and lots of bugging and de-bugging circuits.

If you're interested in buying bugging equipment, this book provides a directory of manufacturers and dealers and tells you a little bit about the legal aspects of electronic bugging. Must reading if you're interested in finding out what is going on in electronic snooping today. $3.95 at your local electronic distributors or write directly to The Hayden Book Company, Inc., 116 West 14th Street, New York, New York 10011.

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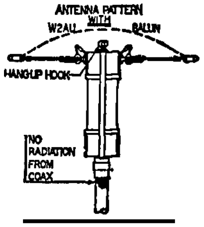

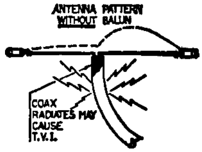

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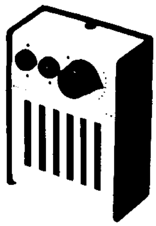

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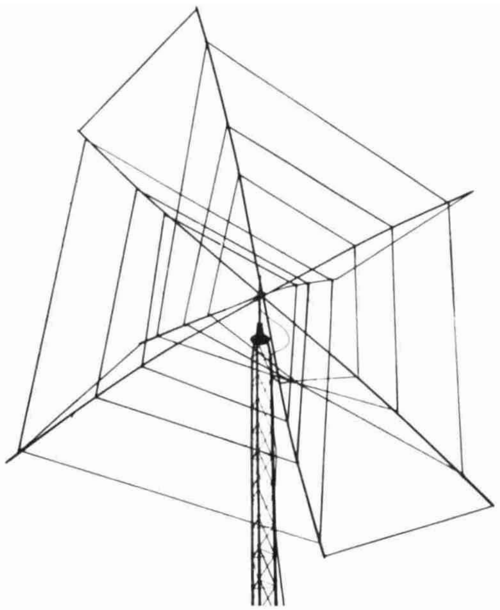

Present day transceiver and transmitter designs require an antenna match with as flat a resonant response as possible. In the olden days a typical 500 watt rig required a relay rack and 500 pounds of gear. The physical spacing and voltage parameter of the tank circuits were such that one never worried about VSWR; indeed, that term had not yet been coined! Flash over and mis-match were laughed at and tolerated.

What a difference today! In a table top box scarcely 1½ cubic feet and weighing less than 50 pounds we find a modern side-band transmitter and receiver thrown in as well. Take a good look at the components: fixed capacitors smaller than ½ a postage stamp; a loading capacitor with .005 inch spacing; and transmitting tubes that are in reality TV horizontal outputs. These, then, are what you find. Do you wonder then why the manufacturer tells us we must operate into loads of 2.5 to 1 or better? Do you wonder then why the factory says to tune up in less than 30 seconds? Is it a surprise to you to find that new finals are required so quickly?

Look at this problem more closely. The 6HF5's are rated 30 watts dissipation. Two of them in parallel equal 60 watts limit. Yet the rig has a PEP rating of 500 watts and can develop up to 265 RMS watts in the for-ward or incident position. Its power supply furnishes 800 volts. You adjust the bias for 50 mills average, or 40 watts. Here is two-thirds of the total dissipation rating in idling power. The danger comes into the picture because the average ham can't or won't con-fine his operation over that nsrrow spec-trum where his antenna is in resonance and where his VSWR is less than 2.5 to 1.

The pi - network output circuitry is only efficient with a SWR of 2 to 1 or less. Thus. when this SWR is exceeded, not only is your signal affected, but many components in your rig are endangered and reliab'lity suffers.

Sure — new SSB designs are in the works: transceivers whose receivers will hsve vari-able selectivity; whose transmitters will use better tubes; rigs that can take more guff. But can you wait that long or afford that much money?

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Wind resistance	4.5 sq ft
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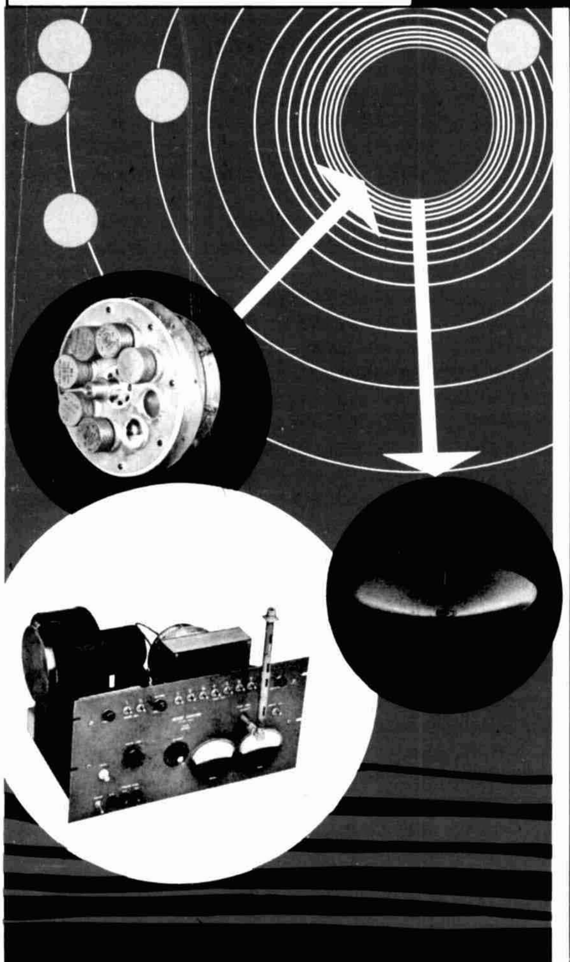

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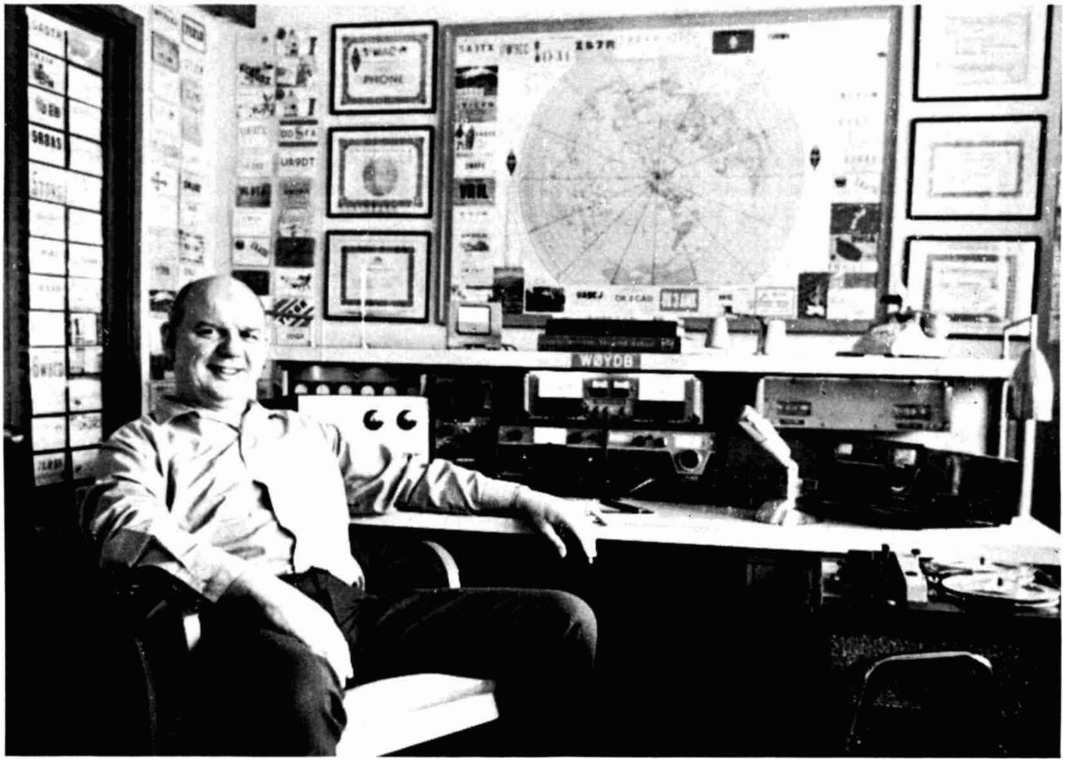

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says WØYDB, Minneapolis. . .

To quote in part from a letter received from W. C. Higgins, WØYDB, Minneapolis, Minn., dated May 10, 1968 . . .

"... Enclosed are several snapshots of my **hamshack** and equipment. Since the Drake 4-Line is so predominant, I thought that you might like to add to your photo collection of Drake-equipped stations. Granted, the gear is not the new B series but it is still the most satisfying and totally efficient that this old-timer has used in 32 years of amateur, military and commercial electronic experience. I earn my living as a Production Manager of (aerospace) electronic instrumentation production. . . and I think I can recognize excellence in electronic engineering design and performance when I see it.

"Again, congratulations for developing the **4-Line.** 73. . ."

(Signed] Bill, W. C. Higgins

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or write for detailed specifications:

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august 1968
volume 1, number 6

ham radio

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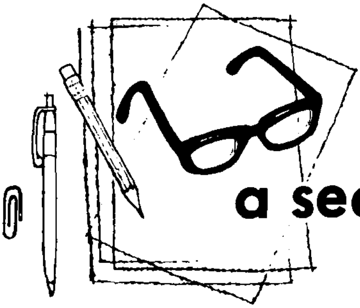
departments



a second look

by jim fisk

If you patronize the bands below 30 MHz, a QSO with a KL7, a KH6 or a PY is practically an everyday occurrence. But when you can chalk up contacts like this on six meters, the band is really popping. Add CE, LU, OA, VP7, KA, KP4, all states and provinces of Canada and it's extraordinary. But that has been the way of six meters over the past few months.

And six meters is not the only band that has been busy. Two meters has provided some unbelievable openings via sporadic-E. W1YQI in Massachusetts reports working stations in Florida, Georgia, Alabama and Louisiana with +9 signals both ways. One station in Florida was using 75 watts and a 75-meter vertical, so power is no problem when conditions are right.

The same evening all this activity was going on, television reception was lousy in New England. Here in New Hampshire television reception was unbelievably bad because of co-channel interference from distant TV transmitters—even on channel seven! It's going to be a long time before we see another evening like that.

Earlier in June, WB6CXF in Riverside, California, worked WØJYC in Colorado, 750 miles away. Six meters was open simultaneously, and according to our propagation expert, the likelihood that this was tropospheric propagation is extremely remote—due to horizon screening and unfavorable terrain. The distance covered in this case is much shorter than expected and indicates that the maximum useable frequency for a 1400-mile path with the same reflection point may have been greater than 180 MHz!

Reports are slow in coming in, but from all indications, the experiences of W1YQI, WB6CXF and WØJYC are not isolated cases. Apparently during the month of June there were more two-meter sporadic-E openings than ever before. Of course, there are far more openings than those recorded, but they occur at the wrong time of day and no one is on to take advantage of them. The opening on the East Coast was unusual because the band was still open at 9 o'clock in the evening. When that happens, there are a lot more stations around to take advantage.

Openings like this don't happen on 432 or 1296 of course, but on 1296, other things have been going on. WB6IOM has been hearing his own echoes with a kilowatt amplifier and a 10-foot dish. As far as I know, this is a record as far as antenna size goes. He is continuing tests with several well-equipped European stations, and it appears that a California-Europe moonbounce QSO on 1296 is just a matter of time.

Ray Naughton, VK3ATN, the famous moonbouncer from Australia, spent the early part of the summer in this country attending conventions and vhf get-togethers and planning his 1296-MHz moonbounce system. His home-made dish is coming along nicely, and as soon as he rounds up the necessary parts for his transmitter, I suspect we'll be hearing a big 1296-MHz signal out of Birchip.

If you're tired of working the world at the flick of a switch—if there's no challenge left on 10, 15, and 20—why not QSY up? The long, hot summer is only half gone. There's still plenty of time for putting up antennas and getting a vhf rig on the air. The sporadic-E season is just about shot, but the Perseids meteor shower in August and almost daily aurora displays offer some interesting opportunities for vhf propagation. So, if ham radio has lost some of its challenge, QSY up—the water's fine.

Jim Fisk, W1DTY
Editor

ham radio and you...

a word from the publisher

Over the past several months or so, we've had the privilege of meeting a large number of our subscribers during many of the larger spring ham shows, such as the Dayton Hamvention and the ARRL National Convention in San Antonio. This has been a very rewarding experience for all of us as it has given us a chance to learn first hand what the average reader thinks about **ham radio** magazine, and what he would like us to do to make it a more enjoyable and useful addition to his hobby.

It's been made very clear to us that **ham radio** is the kind of magazine that you've been looking for and that you don't want it changed. Don't worry — it won't be. It will only be improved.

Many good ideas were given to us, however. You'll notice several of them being used in months to come as we follow up on various suggestions for articles and possible improvements.

As we have met you, both by mail and in person, we have found a group totally interested in our new magazine and in its prospects for success. Many have asked what they might do to help with our growth. This month we are giving all of our subscribers an opportunity to do just that. Each envelope is being filled with two copies of this issue, and we are asking each of you to see that some ham friend of yours across town, across the country or around the world receives the extra copy along with your impression of **ham radio.**

The success of this idea depends on you. No one can sell **ham radio** as well as a satisfied reader. We hope you are that reader. Give us a hand, and we'll be able to give you an even bigger and better magazine each month.

If you can use more than one extra copy let us know, and we'll send as many as you would like to pass on to friends or to give out at your next club meeting. While on the subject of your club, remember that we have special bulk subscription rates which are well worth investigating. Drop us a line and we'll be glad to let you in on the details.

Many of you have been very interested in just how well we've been doing. The answer is a most pleasant one for me to digress on for a moment or so. Before undertaking this venture, extremely detailed operating forecasts were prepared to determine whether there really was a place for **ham radio** magazine. Both our anticipated growth, and the price that would have to be paid to achieve it were carefully estimated, not only to prove our feasibility, but also to be used as a guideline during our development. We can happily report that our rapid growth has come as a surprise, even to our staff, and that the original projections now appear to be quite conservative. In virtually every category we are now ahead of our estimates, yet this has been done within the original budget which we thought was necessary to put our magazine on its feet.

The speed and warmth of your reception has certainly helped us over many of the difficult periods common to any young enterprise. Your letters and comments have meant a lot to all of us. There is still much hard work ahead to complete all of our plans for **ham radio,** but with your support, the road will be a lot smoother and a good deal shorter.

Skip Tenney, W1NLB
publisher

EIMAC

3-500Z's used in Drake's linear amplifier for 2 kW PEP at 3.5-30 MHz

The R. L. Drake L-4B linear amplifier shown here uses two of EIMAC's new 3-500Z zero-bias triodes in grounded grid circuitry to achieve 2-kW PEP SSB input and 1-kW dc input on CW, AM, and RTTY. Drive power is 100 watts PEP and 75 watts CW, AM, and RTTY.

Drake chose EIMAC 3-500Z's because these rugged, compact, high-mu power triodes are ideal for grounded grid operation. They can provide up to 20 times power gain in a cathode driven circuit. And the two tubes have a total plate dissipation rating of 1000 watts.

For more information on EIMAC's line of power tubes for advanced transmitters, write Amateur Services Department. or contact your nearest EIMAC distributor.

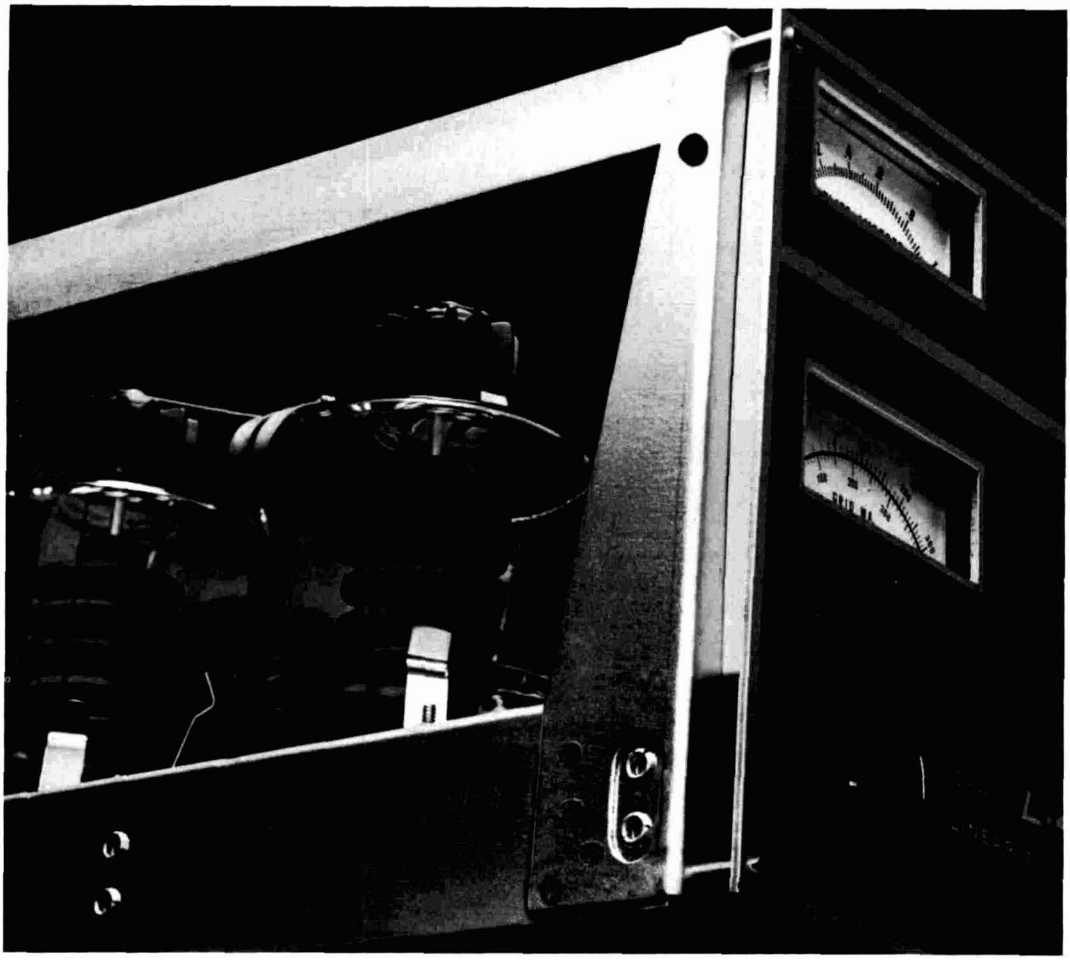
3-500Z TYPICAL OPERATION'

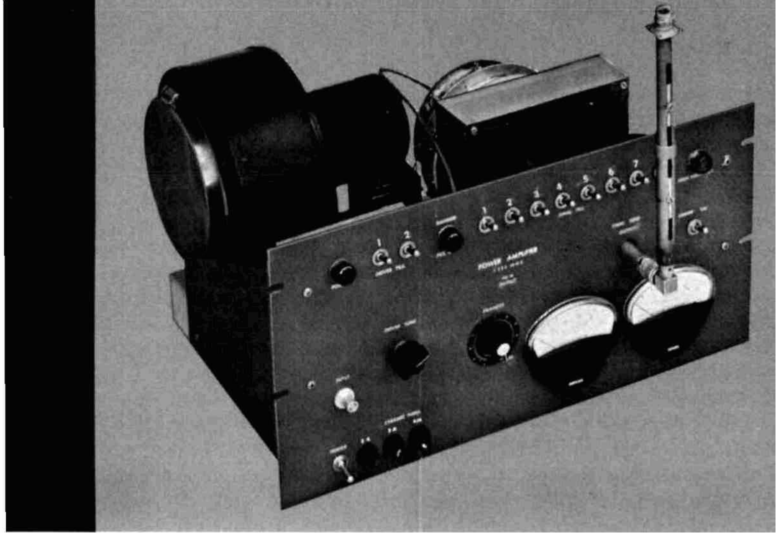
DC Plate Voltage	2500 V
Zero-Sig DC Plate Current**	130 mA
Single-Tone DC Plate Current	400 mA
Single-Tone DC Grid Current	120 mA
Two-Tone DC Plate Current	280 mA
Two-Tone DC Grid Current	70 mA
Peak Envelope Useful Output Power	500 W
Resonant Load Impedance	3450 ohms
Intermodulation Distortion Products	–33 dB

'Measured data from a single tube

**Approximate

EIMAC
Division of Varian
San Carlos, California 94070





high-power linear for 1296 mhz

500 watts out with an octet of 2C39 ceramic triodes

Peter Lakma m, 6IOM, 8001 Airlane Avenue, Los A geles, C3 l forn'a 9045

The difficulty of generating useful amounts of power on **1296 MHz** without using high-power klystrons has limited the interest of troposcatter and moonbounce enthusiasts. Commercially available high-power tetrodes such as the **7650** and **7213** are expensive and don't exhibit the power output and gain necessary for amateur use on **1296.** These tubes are rated for service as uhf TV drivers or **1215-MHz** radar use and are marginal at frequencies above **1215 MHz** because of sharply decreasing power output and gain at only slightly higher frequencies. I have used a **7650** for the past two years; the best output **I** could obtain was **350** watts with 120 watts drive and total plate power input of **7.5** kW.

For these reasons, I've spent considerable effort replacing the tetrode amplifier with a parallel arrangement of eight 2C39A's. The high-power 2C39A amplifier described here can be easily duplicated by an experienced uhf'er; it uses a single power supply and exhibits nearly as much power gain and output as the average lower frequency linear.

the 1296 linear

The complete linear amplifier consists of two cavities on a single large chassis. The first one contains a pair of 2C39's in a square cavity, while the final consists of eight tubes in a round "radial-vane" arrangement. The required drive for full power output is **30** watts. Total plate current for both the driver and final amplifier is about two amps.

The square driver cavity was originally developed to drive a 7650 amplifier and was described in **QST**[1] as well as the 1968 edition of "The Radio Amateur's Handbook." It can deliver as much as 120 watts output with 30 watts drive; with about 15 watts drive, it will put out about 75 watts. Make sure that the tuning-screw guiding nut (6-32) is soldered to part B in **fig. 4** of the **QST** article.

The driver and final are both operated with their grids at dc and rf ground. Since the filament of the 2C39 is in common with the cathode, floating filament transformers must be used. Grid bias is obtained from a Zener diode and shared by both amplifiers. Both amplifiers operate in class B for low excitation and move toward class C with medium and full drive. With peak signal excitation, the grid bias is about 40 volts; grid cut-off bias is only 9 volts.

If you're not interested in using this amplifier on ssb, you can leave ou the Zener diode and use two large cathode resistors instead. Peak output power is the same either way, but linearity is better with Zener bias. However, don't use fixed bias; it can't be used safely at the drive power levels used with these amplifiers.

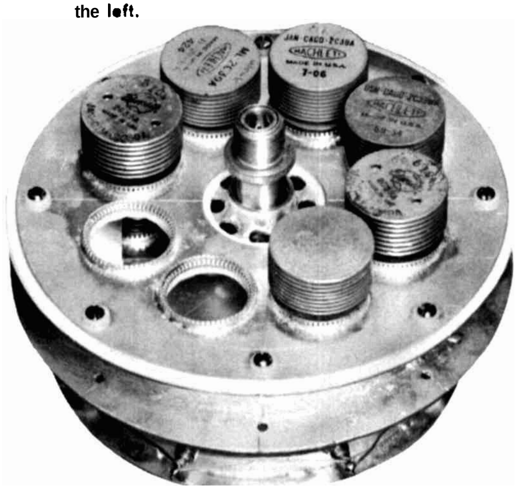
All filaments are wired to front panel switches so the condition of each 2C39 can be analyzed under operation. When a tube is turned off, a light comes on to warn against using the amplifier under full power.

A variable transformer is provided for adjusting filament voltage for the particular operating mode. For ssb and short bursts of CW or continuous operation as an a-m linear amplifier, full filament voltage is used. For long CW transmissions, the filament voltage should be reduced about 10% to prevent short tube life due to bombardment heating of the cathode by drive power. Operation of the amplifier at full output power without keying is not recommended for more than about 20 seconds, in which case the filament voltage should be dropped about 20%.

The 2C39 family is rated at 125 mA maximum cathode current. In ssb service the tubes will handle 300 mA peak current each. In CW operation, with normal keying speeds, they will provide excellent life when keyed up to 220 mA cathode current. When keyed at 300 mA, runaway effects can be noticed, and with higher currents the cathode is quickly destroyed.

Fortunately, cathode resistance keeps the current quite stable, so that amplifier operation is not critical. When operated at maximum power, the eight tubes in the final dissipate about 1000 watts peak and about half that when keyed. This is well within the ratings of the 2C39's, but the tubes must be well cooled to take the cathode-current overload.

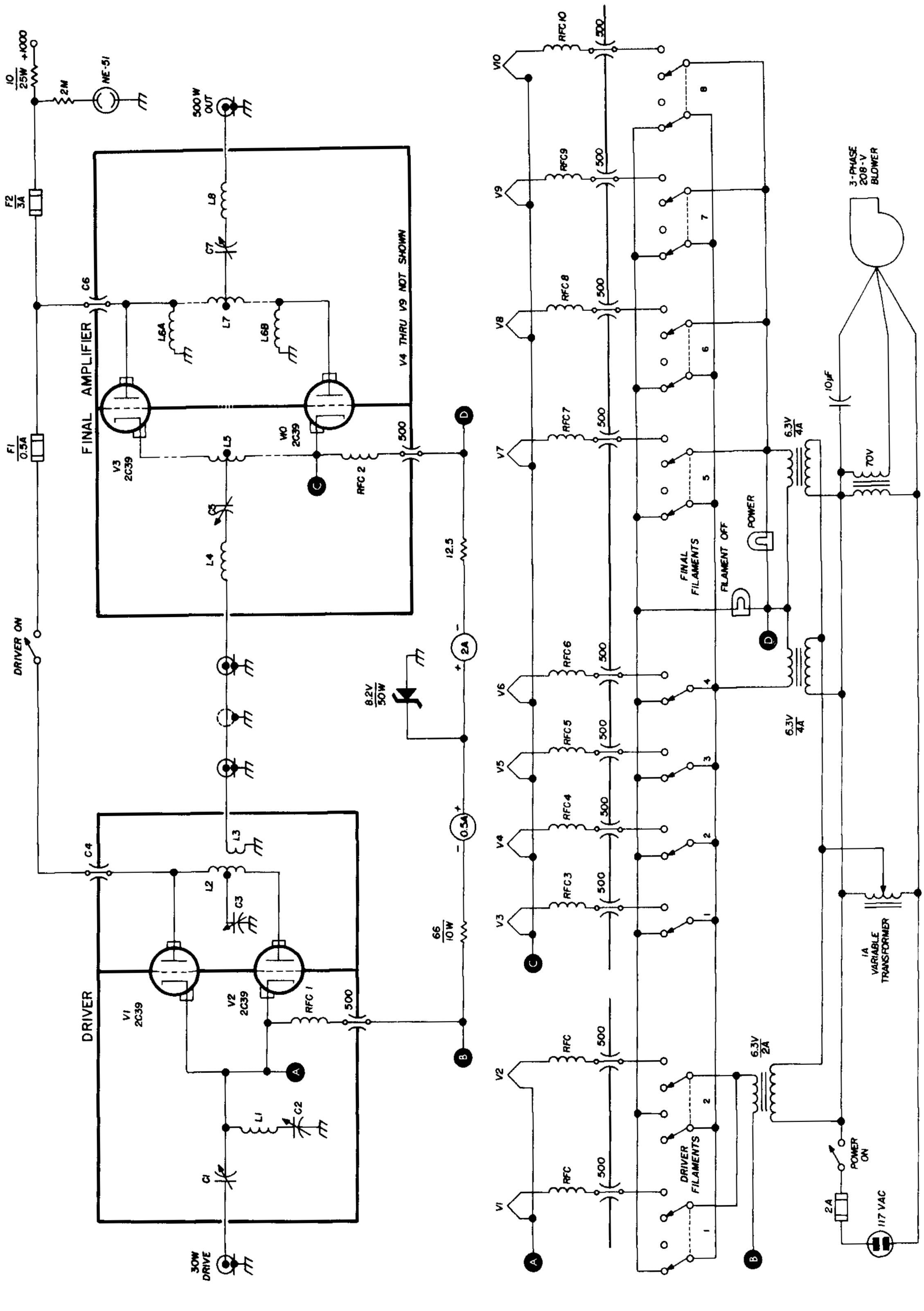
The high-power 1296 amplifier, plate ride up. You can see one of the tuning vanes through the open hole on the left.



cooling

The common blower for both cavities delivers about 300 cfm against a back pressure of two inches of water. The blower I use is a three-phase 220-volt unit operated with a capacitor and step-up transformer that adds 75 volts to the 110-V line. Two X-band waveguide sections are used as air outlets directly under each drive-tube radiator. The main air blast is directed into the center of the final tube ring (A) through a 3-112-inch hole in the square aluminum box as shown in the photograph.

The side of the box facing the tube assembly is lined with teflon and contacts the tubes; air escapes radially past the tubes, and holes are provided to pass some air through the cavity to cool the output coupler and tube



FINAL AMPLIFIER
DRIVER
V4 THRU V9 NOT SHOWN
V1
2C39
V2
2C39
V3
2C39
V10
2C39
30W DRIVE
500W OUT
DRIVER ON
F1
0.5A
F2
3A
10
25W
+1000
2M
NE-51
C1
C2
C3
C4
C5
C6
C7
L1
L2
L3
L4
L5
L6A
L6B
L7
L8
RFC 1
RFC 2
500
66
10W
12.5
0.5A
2A
8.2V
50W
RFC
RFC3
RFC4
RFC5
RFC6
RFC7
RFC8
RFC9
RFC10
DRIVER FILAMENTS
FINAL FILAMENTS
FILAMENT OFF
POWER
6.3V
4A
6.3V
2A
10μF
70V
3-PHASE
208-V
BLOWER
1A
VARIABLE
TRANSFORMER
POWER ON
117 VAC

fig. 1. (Left) Schematic diagram of the high-power linear amplifier for 1296 MHz. Construction details for the dual 2C39 driver stage are given in the January, 1968 issue of QST; part numbers are the same. Although a total of eight 2C39's are used in the actual final amplifier, for clarity, only two are shown here.

L4	Part of the input probe, fig. 4
L5	Cathode disc, part F, fig. 2
L6	Tuning vane, part J, fig. 2, 8 required
L7	Plate cavity. Consists of parts B, C, and D, fig. 2
L8	Part of output probe, fig. 3
RFC 3-10	10 turns number 22 enamel, 1/8" diameter, 1 inch long
RFC 2	4½ turns 1/1/16" wide brass strip airwound on 1/8" diameter
C5	Part of input probe, fig. 4
C6	500-pF feedthrough bypass
C7	Part of output probe, fig. 3

cathodes. The 2C39's can be easily removed by unscrewing the output coupler, removing the box, and removing the filament clips.

the final cavity

The final cavity is based on a different concept than the driver and is applicable to any number of tubes. However, since the power gain of the 2C39 is 6 dB at 1296 MHz under maximum-efficiency drive conditions, the number of amplifier tubes should progress by a factor of four from amplifier to amplifier. The following progressions are therefore naturals: 1, 4, 16 . . . or 2, 8, 32 . . . The first chain would deliver about 1 kW out while the second could deliver 2 kW. Eight tubes in the final is a good choice since it's compatible with FCC regulations regarding maximum input power for amateur ssb transmitters.

The eight 2C39's are placed around a circle 3-1/2 inches in diameter. At this diameter, adjacent radiators clear by about 1/16 inch. The minimum size cavity that will provide room for eight tubes has an inside diameter of 5 inches. However, the resonant frequency of this assembly is a good deal below 1296 MHz. To raise the frequency of the cavity, vanes are put into the space between tubes. These act inductively to raise the frequency of the cavity to any desired value. For 1296 MHz, the vanes protrude 3/4 inch into the cavity.

The vanes are fixed in place, and the cavity is tuned with a capacitive probe in the center. The capacitive probe also doubles as the output coupling circuit. It is split into two concentric rings with the outer one grounded; the inner ring is connected to the center conductor of a 50-ohm type-N coaxial fitting.

The probe is constructed from a dual female type-N chassis feedthrough connector (UG-30D/U). The threaded portion of a type-N male cable plug is soldered to the center hole of the plate cavity to serve as a guide for the dual female feedthrough. The longer end of the UG-30D/U is extended with a brass sleeve 1/2 inch in diameter and 318-inch long as shown in **fig.** 3. A 2-56 screw about 5/8 inch long is fastened to a 6-32 nut by screwing a 2-56 nut against the 6-32 nut as shown in fig. 3. The 2-56 screw just fits into the center pin of the UG-30D/U. The 6-32 nut should stick out 1/16 inch beyond the end of the brass sleeve. Wrap thin teflon sheet around the center probe until it is held rigidly by the sleeve. Solder can't be used to hold the 6-32 nut on the 2-56 screw because the high rf currents at this point will melt the solder.

The output is brought through the cooling air box and the front panel with a 50-ohm air-dielectric line to a type-C connector. The type-C connector acts as a tuning knob and rotary joint.

The output cavity does not have any provisions for adjustable impedance matching; its output coupler is designed to deliver maximum power into approximately 50 ohms. Exact impedance matching is accomplished with the slide tuner mounted on the front panel. This tuner was developed by W6DXJ and can match a 2:1 VSWR of any phase. It consists of a 9-inch air-dielectric 50-ohm transmission line with two moveable 1-1/2-inch long teflon slugs.

An exact impedance match between the driver and final can be established by alter-

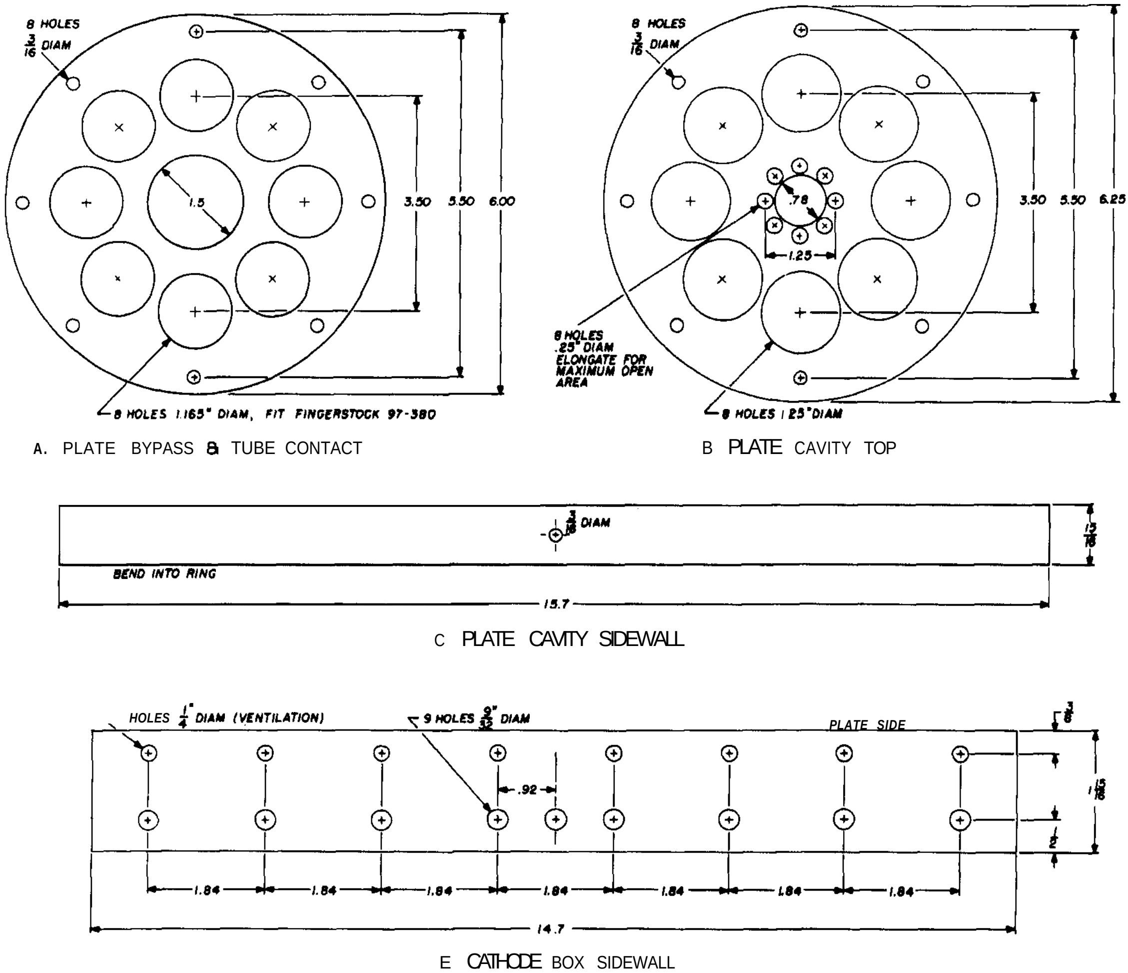


fig. 2. Mechanical parts of the high-power 1296-MHz linear amplifier. Material for all parts is .050" brass. Assembly of the parts is illustrated in fig. 5.

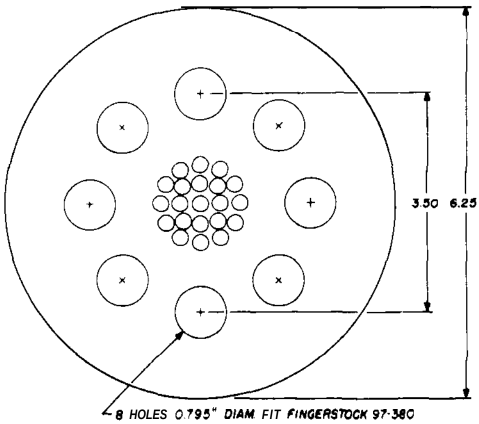
nately tuning the driver cavity and final amplifier input probe.

the cathode cavity

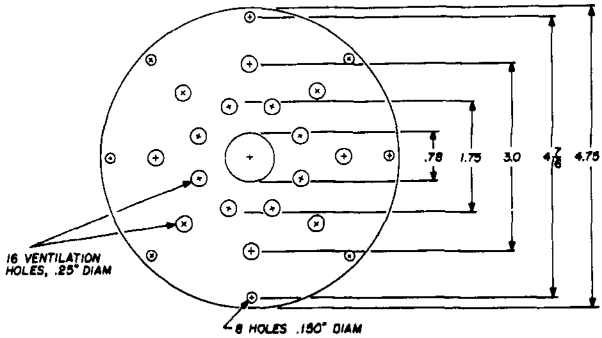
The cathodes of the eight tubes are connected to a 4.2-inch diameter brass plate with finger stock. This plate (F) is spaced 3/8 inch from the cavity bottom and is not resonant in itself. However, drive power is applied to the center of the disc with a resonant probe. Because of physical symmetry, drive power splits equally between the tubes.

Input coupling is provided by a UG-30D/U connector modified as shown in **fig. 4**. The threaded portion of a type-N male connector is soldered to the bottom of the cathode cavity as a guide. A 1/2-inch diameter brass or copper disc is soldered to a 1/8 inch wide, 1-1/4 inch long brass strip that is pushed into the center pin of the UG-30D/U female adaptor and soldered. The disc is spaced 3/4 inch from the end of the UG-30D/U. A hairpin inductor 1/4 inch long is soldered between the strip and the face of the UG-30D/U. This inductor is also made from 1/8 inch wide brass sheet.

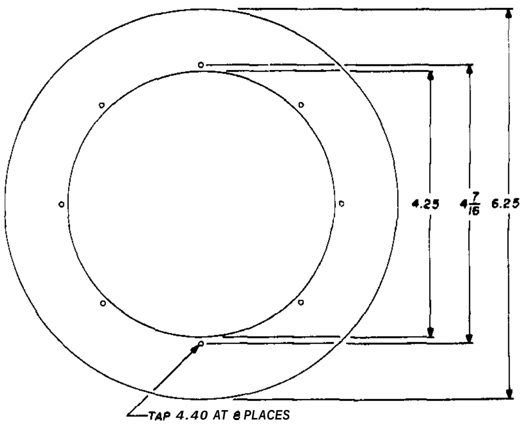
This arrangement provides an input impedance of approximately 50 ohms. It is not



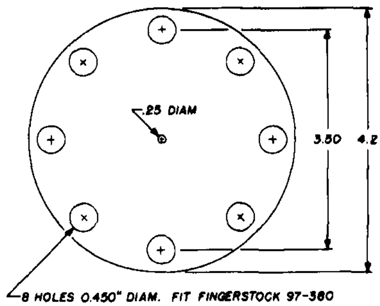
D. PLATE CAVITY BOTTOM



H. BOTTOM COVER

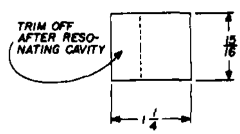


G. CATHODE BOX BOTTOM



F. CATHODE DISC

J. TUNING VANE (8 REQUIRED)



too critical because an exact impedance match is established by tuning both the driver and input probe for maximum power transfer. The coaxial cable from the driver to the final must be a low-loss, high-temperature teflon cable because the VSWR on this line may be very high. Don't make it any longer than is absolutely necessary.

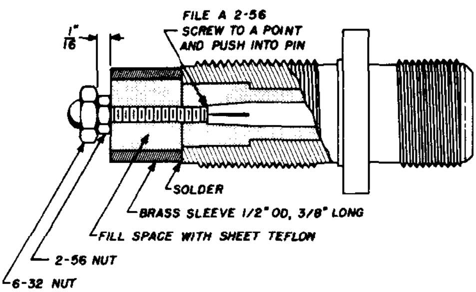
Filament connections to the 2C39's are made with U-shaped spring brass clips which make a snug fit in the filament cup at the end of the tube. The common cathode choke (RFC2) should be made of 1/16 inch wide sheet brass to handle the current. To minimize the filament current flowing over the cathode choke, the filament supply should be derived from two separate transformers or a center-tapped 12-V unit. If you use two transformers, they must be phased to operate like a center-tapped transformer so filament current through the cathode choke is zero when all tubes are turned on.

cavity construction

The cavity parts are cut out as shown in **fig. 2** with an electric or manual coping saw and filed to the proper dimensions. The tube holes may be punched with the nearest smaller punch and then reamed to the exact dimensions shown in the drawings. Size is critical for proper tube fit.

The first part in the assembly operation is the plate cavity. The grid finger stock* is preformed by hand until it fits the grid holes without a gap between the hole and the fingerstock. The two discs (B and D) are then aligned with each other and the cavity wall (C), is formed into as round a shape as possible.

fig. 3. Output probe for the 1296-MHz amplifier.



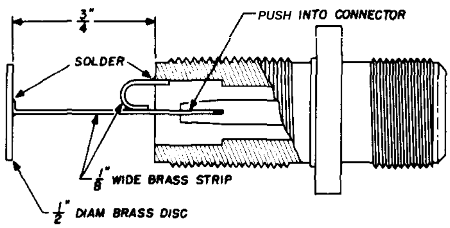
The assembly is clamped together and **slowly** heated over a gas stove. Avoid fast heating because it causes warping and bending that may remain after the unit has cooled down. The grid side of the cavity should face the stove, but the flame should not touch the sensitive fingerstock. As soon as the cavity is hot enough to melt solder, touch the outside of the fingerstock with acid flux and solder. The solder flows around the grid hole quickly and provides a clean solder joint.

When all the grid rings are in place, run solder along the wall of the cavity on the grid side and attach the plate side in as many places as you can easily reach. Let the cavity cool down slowly by reducing the heat gradually. Now invert the cavity and put the plate side toward the stove. Heat it up again gradually until it is warm enough so you can solder the cavity wall (C) to the plate disc (B) with a 200-watt soldering iron. Don't let the cavity get hot enough to melt the grid points you already made.

The next step is to solder the fingerstock to the plate-bypass disc (A). However, before proceeding, hammer the disc around the center until it is slightly saucer shaped. This will result in a compression force when it's screwed against the top of the plate cavity. This will assure that the teflon insulator is compressed throughout. If you punch holes in the disc, this is probably its natural condition. Make sure that the fingerstock is installed on the proper side and in the right direction. The guide for the capacitive output probe may also be soldered in at this time.

The plate cavity can now be slotted for the vanes—preferably with an electric coping saw. Simply cut the eight slots by running the saw blade between the tube holes. The slots should extend about one inch into the cavity for sufficient adjustment range. Put pieces of thin brass in the slots and adjust for 3/4 inch protrusion into the cavity.

fig. 4. Input probe for the 1296-MHz linear.



preliminary adjustment

The plate cavity may now be assembled for preliminary test and adjustment. The teflon insulator between the plate disc (A) and cavity is cut from 0.02-inch teflon sheet. Don't use a different thickness of teflon or the bypass may become resonant. The tube holes should be just large enough to clear the anodes. Be sure the plate ventilation holes are not obstructed. The shoulder washers used to assemble the plate bypass disc to the cavity should be either teflon or ceramic; nylon parts are so lossy at these frequencies that they will melt with only milliwatts of rf.

You can put the 2C39's in and check for mechanical alignment. If everything fits, couple about 20 watts of 1296-MHz rf through at least a 10-dB attenuator (40 feet of RG-58/U) to the output coupler. Put a number 49 lamp with a 1/2 inch diameter loop through the hole provided in the cavity

* Part number 97-380 from Instrument Specialty Company, Little Falls, New Jersey.

wall and tune the probe until lamp lights—indicating resonance. If necessary, adjust the vanes equally until resonance occurs with the probe about 5/16 inch from the bottom of the cavity.

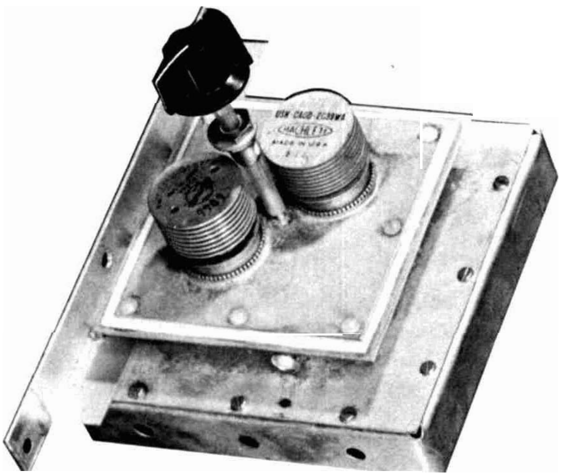
The cavity can now be disassembled and pre-heated so you can solder the vanes in place with a 200-watt soldering iron. Add the remaining parts by working with a soldering iron over the gas stove. Be careful not to overheat the cavity or everything will fall apart. Assemble the cathode disc (F) in the same way as the plate bypass. When inserting tubes for testing under power, adjust the cathode disc so it is saaced uniformlv 3/8 inch from the bottom of the plate cavity.

After checking everything for alignment, disassemble the cavity and have all the parts silverplated by a professional. This is a must to prevent contact-resistance heating, especially at the tube contacts. Don't try to do your own plating—it will only lead to disappointment later.

power supply

The power supply for the amplifier must be capable of delivering 2.0 A at a terminal voltage of at least 900 volts to obtain full power output at a safe plate current. Both efficiency and power gain are dependent on plate voltage; the higher the plate voltage, the more power the amplifier will deliver at

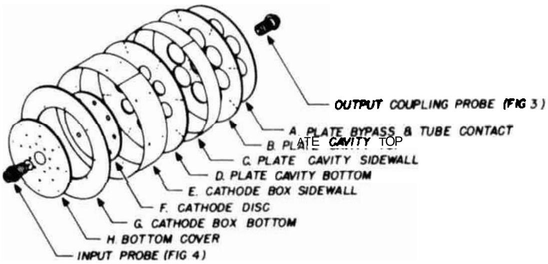
The dual-2C39 1296-MHz 100-watt driver amplifier.



the maximum safe cathode current level of 220 mA per tube (CW). On the other hand, to avoid arcing within the tubes, maximum plate voltage must not exceed 1200 volts. For these reasons, maximum power output will only be obtained safely from a well regulated supply. I use silicon rectifiers and a choke-input circuit with a bleeder; transformer voltage is 1250 volts. Regulation from zero to full load is 150 volts.

The power supply is provided with a 20-A circuit breaker in the 115-V line. This is an additional safety precaution to reduce the

fig. 5. Mechanical 'assembly of the amplifier. The tuning vanes are not shown, but their position is indicated by the dotted lines on parts B and D.



chance of prolonged exposure of the tubes to excessive cathode current. The circuit breaker trips after about 30 seconds operation at maximum power without keying.

The fuses are wired so that drive power is removed when the final-amplifier fuse blows. This avoids excessive grid current in the final. The fuses for both the octet and the driver must be fast-blow ceramic types. Glass fuses (3AG) will explode and form an arc in the fuse holder. This will destroy all the tubes in the octet if one of them develops a grid-to-cathode short. Type 3AB fuses are recommended; they fit the same fuse holder as the 3AG but have a ceramic body. Don't use larger fuses than the values shown in the schematic.

I spent a lot of time trying to develop an automatic overload protection circuit for this amplifier in case of tube arcing. I originally thought that a simple series pass tube with an overcurrent sense circuit would prevent final-amplifier tube damage by removing plate voltage to the driver. However, after building such a circuit, I discovered that once an arc in one of the final tubes is initiated, removing plate voltages from the driver does not stop it; the final-amplifier fuse blows anyway.

The only remaining possibility is to remove all plate power from all tubes. This requires a large tube to pass two amperes of plate current with less than 100 volts drop. I tried it with three 6336A's in parallel; voltage drop was about 75 V. However, constant problems with false triggering of the high-speed circuit finally forced me back to fuses. This is really not such a bad method because arcing is only likely to occur during initial tuneup. The 10-ohm resistor in the high-voltage lead limits peak current to 100 A and is recommended with these tubes.

tuning up

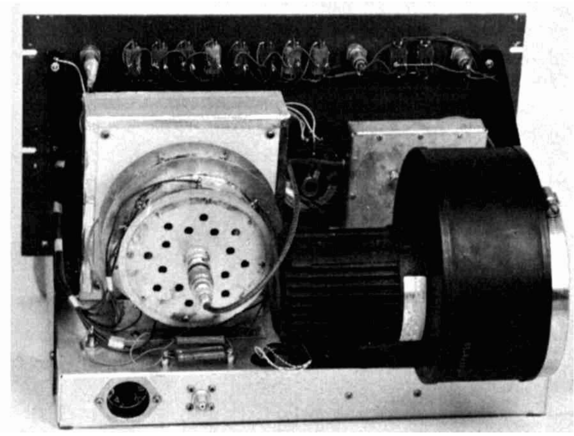
After the driver is tuned up,[1] you can tune up the final. Apply about 50 watts of drive power to the cathode probe (plate voltage disconnected) and adjust the probe for resonance by monitoring cathode current; cathode current should be about 0.5 A. Alternately peak the driver and the cathode probe. If necessary, reduce power so you don't exceed 0.5 A. Check the clearance between the probe and the cathode disc. Resonance should occur with about 1/4 inch clearance. If resonance occurs with the probe too close to the disc, lengthen the probe or increase disc size.

Apply plate voltage and cooling air. Connect the output probe to a dummy load—through an impedance-matching device if possible. Slowly apply drive power. As soon as the final plate current comes up, resonate the output probe. Increase input drive to about 10 watts and tune the driver cavity input and output, cathode probe and output probe for maximum output.

Shut down the amplifier and check the clearances of the input and output probes for safe values. Output clearance shouldn't be less than 3/16 inch and input clearance not less than 1/8 inch.

Reapply power and cooling air and peak up to full power. It helps to key the transmitter with an electronic key to keep the duty cycle down. Unscrew the input and output couplings on the cavity and check for possible damage from overheating. If a fuse blows whenever you hit resonance at full drive, it's an indication that the output coupling is too loose.

You can increase output coupling (if you do not have an impedance matcher) by increasing the clearance slightly between the 6-32 nut and the brass sleeve. If you run out of ceramic fuses while tuning up, shut down and get some more. Do not use slo-blos under any circumstances—they can wipe out all the tubes instantly.



Rear view of tho **complete** high-powor **linear amplifier.** Tho final **is** on tho **left;** tho **dual-2C39** driver **stage** is bohind tho blower on tho right. **The large** box **behind the** final **is** tho air box.

checking tubes

To check the driver tubes, simply turn one or the other off, wait for the filament to cool down and record final cathode current after slightly repeaking the plate circuit of the driver.

The best way to check the final tubes is to remove filament power from one of the driver tubes to prevent overload of the final tubes and set excitation power to the driver to a relatively low fixed value. Then turn off all but one filament in the final. This way you get a relative reading of emission for all tubes in the final.

Admittedly, this is a slow procedure, but it's a lot faster than pulling tubes. It takes about 15 minutes to go through all tubes in the linear. You have to wait until a tube cools down before you can get a reading on the next one and the driver has to be repeaked for every new tube reading. It's a good idea to keep a log of tube condition near the amplifier for reference. If you take tube emission data at frequent intervals, you can locate an arcing tube quite rapidly after the amplifier has been in service awhile. A tube that arcs suffers cathode damage and shows a drop in cathode emission when compared to the last log entry.

tubes

Now a word about tubes. There are a variety of tubes available under different type numbers that will work in the cavity. All are more or less identical and deliver the same power when in good condition. However, the ceramic types, 7289, 2C39B or 3CX100A5, are the most rugged. Some 2C39A tubes are ceramic, but most are glass. Glass tubes have a tendency to develop pin holes and leaks when pushed hard, and are not recommended for high-power work.

Tubes such as the 7211 have higher capacitance and don't tune with the same vane settings. However, this tube can put out more power than the 2C39 family since it can stand more voltage and has twice the cathode. Unfortunately, they are not readily available and are twice as costly new. I've tried them in several cavities; they don't work too well with a 1000-V plate supply but should be quite a tube at 2000V at 250 mA.

safety

This amplifier delivers a substantial amount of microwave power and it can be dangerous if it's not handled properly. Radiation causes internal heating of body parts and eventual destruction. However, there is no danger in the vicinity of the amplifier since the amount of power leakage is low. The danger exists near feeds for parabolic antennas, open connectors and any other **physically small** antenna or radiating device. The radiation only causes a very slight increase of skin temperature and can easily be overlooked.

Most of the radiation is absorbed inside the body where there are no nerves to warn you. The most sensitive part of the body are the eyes since they have no warning mechanism for heat. Never look into feeds under power at close range! The same applies to waveguides and other small antennas. However, it is safe to look into a large dish since the power density is low.

Take a look at the appropriate handbooks to determine safe distances for the antennas you are going to use. For amateur power levels, antennas 20 feet in diameter or larger are no problem at any distance. Smaller antennas, however, have a power-density maximum several hundred feet from the antenna that can be excessive.

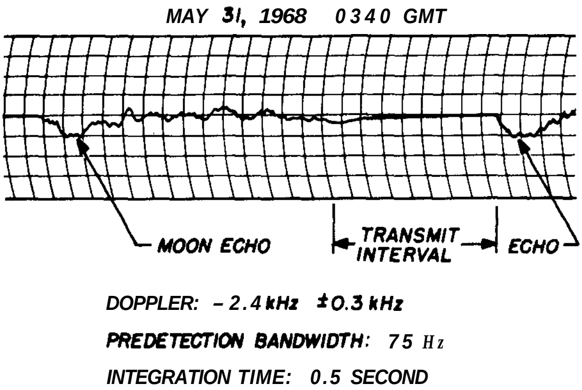
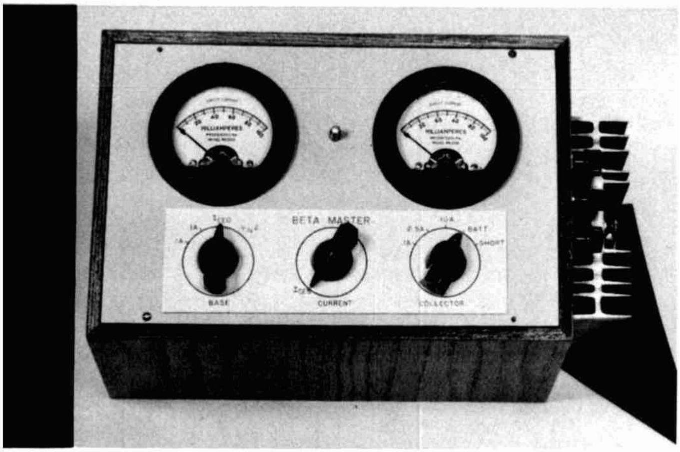


fig. 6. Chart recording of moon echoes received by WB6IOM on 31 May 1968 using this amplifier and a 10-foot dish. Signals peaked about 6 dB over the audibla threshold. Although WB6IOM used a parametric amplifier at the receiver, his feedlina has 1 dB loss, so a good transistor preamp mounted at the feed would do almost as well.

reference

1. P. Laakmann, WB6IOM, "Cavity Amplifier for 1296 Mc," *QST*, January, 1968, p. 17.

ham radio



the beta master

Although designed primarily for matching PNP power transistors, the beta master will also measure many of their parameters

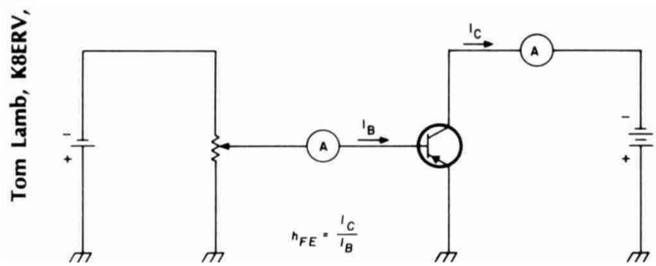
Tom Lamb, K8ERV, 1066 Larchwood Rd., Mansfield, Ohio 44907

Transistors are often used in pairs, such as class-B modulators and dc-to-dc converters. Many semiconductor suppliers do not offer matched pairs, or if they do, they charge for the matching service. The beta master will quickly plot the dc current gain, h_{FE}, of any PNP power transistor up to a collector current of ten amperes.

A simplified circuit of the unit is shown in **fig. 1**. An adjustable known current (I_B) is fed to the transistor base and the collector current (I_C) is measured. The dc current gain, I_C/I_B, is designated h_{FE}. This current gain generally decreases with increasing collector current as shown in **fig.** 2. If all three transistors were measured at 1-1/2 amps, they would all exhibit the same current gain. However, only transistors B and C have nearly the same gain throughout the range, so only B and C are matched.

The complete circuit of the beta master is shown in **fig.** 3. An emitter follower, Q1, provides adjustable base drive of zero to 1

fig. 1. The basic circuit for measuring transistor dc current gain.



amp without a high-wattage potentiometer. The base-meter circuit uses a 100-mA meter that may be shunted to 1 amp; it may also be used as an input voltmeter.

The transistor to be tested is mounted on a heat sink containing a socket. The heat sink can be relatively small because the transistors only pass current long enough for the meters to be read. A 4 × 4-inch aluminum plate, 118-inch thick will do if a finned heat sink is not available. Jacks are provided for testing transistors mounted in equipment, or ones that will not fit the socket.

The collector circuit is similar to the base circuit—essentially a multirange ammeter. Collector current ranges of 0.1, 2-1/2 and 10 amps are provided. Two collector switch positions test for weak batteries and shorted transistors.

The collector supply battery must supply 10 amps for short periods without too much voltage drop. Nickel-cadmium storage batteries are ideal; these are occasionally available on the surplus market. At 10 amps discharge, the voltage drop is very small. Four Eveready 1.5-V E95 alkaline flashlight batteries connected in series-parallel will also work.

construction

The beta master is simple to build. Remember that large currents are being handled, and small voltage drops can cause errors, so use large wire in the emitter and collector circuits. S2 should have heavy duty contacts, or should have two sets of contacts wired in parallel. Q1 is mounted on a small aluminum sink. Either the collector of the transistor or the sink must be insulated from the rest of the circuitry. I bolted the transistor directly to the heat sink and mounted it on the meter terminals with insulating fiber washers.

calibration

Each of the meter shunts is adjusted experimentally by comparison against an external meter. The instrument supplies its own adjustable current. There must not be a transistor in the test socket during calibration and S3 must be depressed to take readings.

First, the R3-R4 combination is adjusted. Plug a 1-amp meter into the base and emitter jacks, negative to the base. Put S1 in position 2 and adjust R1 for either a l-amp or 500-mA reading on the external meter. Then adjust R4 by selection or padding until the same reading is obtained on M1.

Remove the external current meter and connect a low-range voltmeter in its place. With S1 in position 4, set the voltmeter to two volts with Rl. Now adjust or select R3 for a full-scale reading on M1. The base meter circuit is now calibrated.

A power transistor is needed to calibrate the collector circuit. Insert any good transistor in the test socket. Break the collector lead and insert a 10-amp meter. The transistor will act as a variable low-resistance load, adjusted by R1.

When S2 is in the high current positions (2 and 3), collector current flows through the 0.1-ohm shunt, R9. This prevents contact resistance in S2 from causing large errors. R9 can be made from ten 1-ohm resistors in parallel or from a measured length of resistance wire.

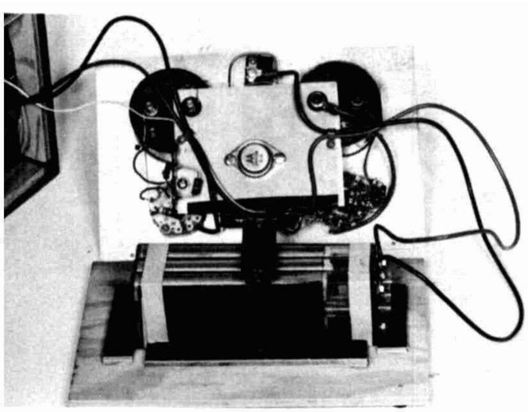
Place S1 and S2 in position 2. Set the 10-A meter at 2-1/2 amps by means of R1. Select or adjust R5 for a full scale reading on M2; R5 will be in the neighborhood of 2 ohms. Other full-scale readings may be calibrated on position 2 if you're not interested in a 2.5-A range.

Now set switch S2 to position 3 and adjust the collector current to 10 amps. If you can't get 10 amps, adjust for 5 amps and calibrate

M2 for 5 amps instead of 10. Select R6 for a full-scale reading on M2; R6 will be about 10 ohms.

Remove the 10-amp meter and reconnect the collector lead. Change S2 to position 5 and jumper the collector and emitter jacks. A shorted transistor will result in this reading obtained on M2; it may be brought to full-scale by reducing R7 to about 30 ohms.

Now switch S2 to position 4. This places about a 10 amp load on the battery and reads the collector voltage on M2. Remove the transistor and jumper. Connect a voltmeter across R8; it must read at least 2 volts, and M2 must read at least 8 amps. Lowering the value of R8 will increase the reading of M2. If you can't get 2 volts, check for weak batteries or voltage drops in the wiring. The voltage loss across R9 can be reduced by lowering the resistance of R9, but is limited when R5 reaches zero when calibrating the 2-1/2-amp scale. If all else fails, use more battery voltage or go to Ni-Cd storage cells.



operation

The **beta master** will check leakage current, dc current gain and transconductance. The tests should be made in the following order.

1. Place the transistor in the test socket and screw it firmly to the heat sink.

2. Place S2 in position 4 (Batt Test) and depress S3. Check for the current found in calibration, indicating that the battery is okay.

3. With S2 in position 5 (Short Test) and R1 at zero current, depress S3. M2 must read less than the value found in calibration, or the transistor is shorted and **must not be tested further.**

4. With S1 in position 3 and S2 in position 3, depress S3 briefly. Should a shorted transistor get past the short test, M2 will pin. A good transistor will give little, if any, reading on the 10-amp scale. Place S2 in position 1 and read the collector leakage current I_{CEO} on the 100-mA scale of M2. This reading is the collector-base diode leakage current multiplied by the transistor gain at the indicated collector current. I_{CEO} is an important indication of transistor quality; a low reading usually indicates a temperature-stable unit.

5. Collector-diode leakage I_{CBO} is indicated on M2 by shorting the base and emitter terminals. This leakage is usually quite small

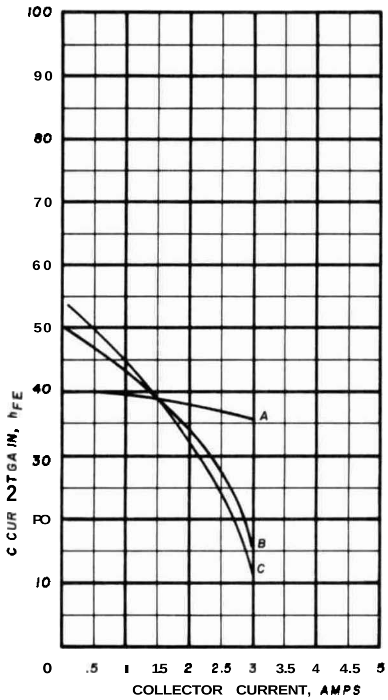


fig. 2. Typical h_{FE} versus I_C curves for three different transistors. From **these curves,** you **can see that transistors** B and C **are fairly** wall **matched.**

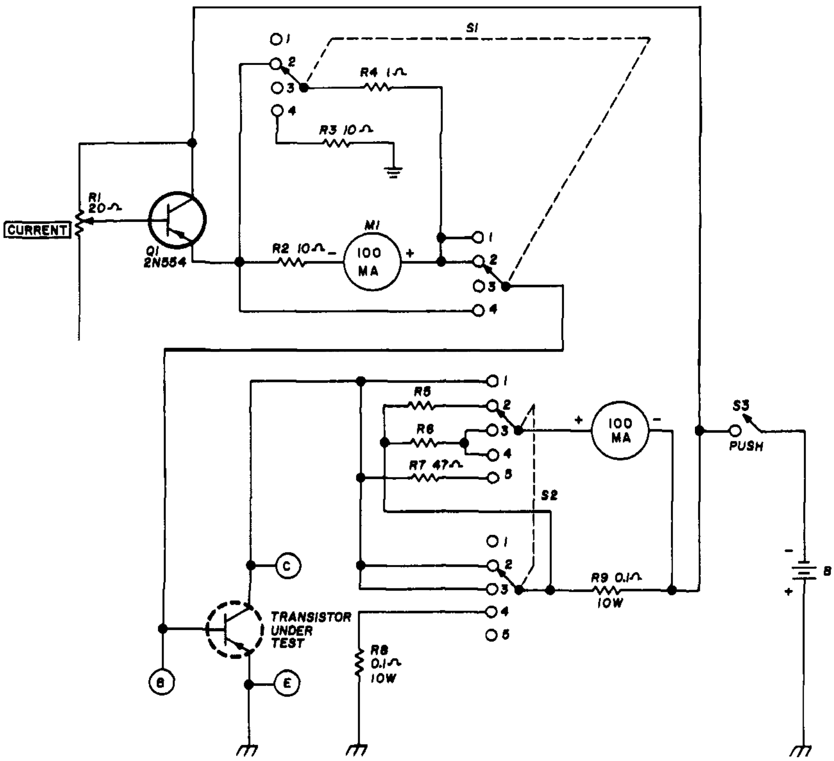


fig. 3. Schomatio **diagram** of tho **beta master. Resistors are all** 1 watt **unless otherwise specified.**

Switch positions:

S1 1 **Base** curront, 100 mA
2 **Base** curront, 1000 mA
3 Colloctor **cutoff current,** I_{CEO}
4 **Base-to-emitter voltage,** 2 V full **scale**

S2 **1** Colloctor **current,** 100 **mA**
2 Colloctor **current,** 2.6 A
3 Colloctor **current,** 10 A
4 **Battery** tort
5 Collector-to-emitter short tort

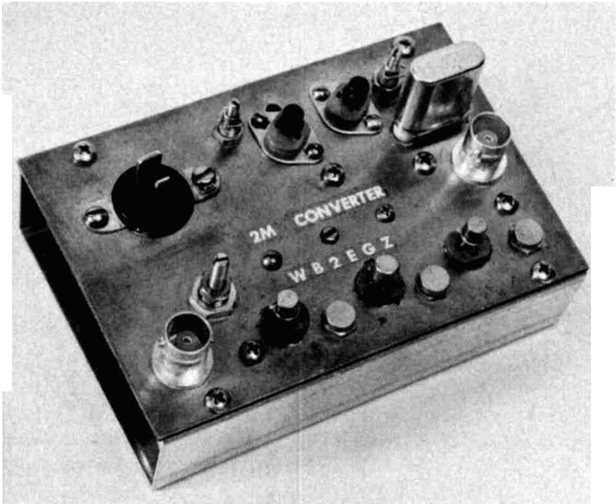
and cannot be accurately measured on a **100**-mA meter.

6. The current gain is measured next. Set S1 to position 1, S2 to position 2, and set the desired base current with R1. At each value of base current, read the collector current on **M2** and record the readings. The collector current should be increased to, or slightly beyond, the rated value for the transistor being tested. Dividing each value of collector current by its corresponding base current gives the current gain at that collector current. Plot each set of values to form a complete curve like those of **fig. 2.** Compare the curves of all transistors of the same type and match up those having the closest characteristics.

7. An input voltage-output curve (transconductance) can also be plotted with the beta master. Just switch S1 to position 4 and run a curve of input voltage vs output current. This curve is useful in determining the correct bias voltage for class-A and -B amplifiers.

The beta master has several other uses besides checking power transistors. The base-to-emitter terminals provide a source of metered, adjustable low voltage—fine for driving tunnel diodes. If M1 is accurately calibrated, these same terminals may be used for calibrating other meters in the 0.1- to l-amp range. The high current drawn by the two-volt meter makes it ideal for testing flashlight batteries under load; connect them to the base and emitter terminals with S1 in position 4 and S3 not depressed. In addition, with S2 in position 5, the collector and emitter terminals make a low resistance ohmmeter out of M2.

ham radio



high-performance mosfet converter-- the two-meter winner

Here's a solid-state two-meter converter that rivals vacuum tubes in cross modulation and excells them in noise figure and gain

Don Nelson, WB2EGZ, 9 Green Ridge Road, Ashland, New Jersey 08034

A successful design is usually more tedium than genius. The development of a solid-state two-meter converter that can outperform the better tube designs has, indeed, been a slow evolution. In recent years, bipolar transistors have been improved to the point where they are competitively priced and have better noise factors than vacuum tubes. However, the conventional transistor has two serious shortcomings: cross modulation and limited dynamic range. Nevertheless, solid-state converters and preamplifiers were designed and loudly praised for reduction of size and power requirements, low noise and even low cost—yet converters using 417A's, 416B's and nuvistors were never really challenged. The tube was still king at rf.

Not more than three years ago, the field effect transistor (FET) became the experimenter's pet. Although the concept of the modern FET predates the invention of the bipolar transistor by twenty years, no one tried to build one until 1958. As with most new inventions, high cost and marginal performance limited much serious amateur experimentation.

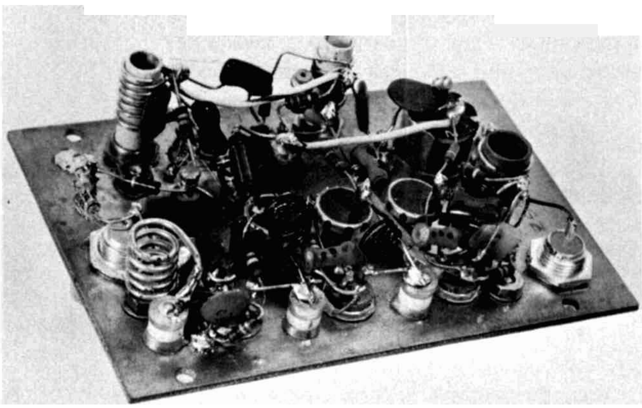
The first FET's priced within amateur reach were junction devices (JFET) that had good noise and cross-modulation characteristics, but suffered from lack of gain. By careful device selection and tricky neutralization schemes, it's possible to get 15-dB gain from a single-stage JFET amplifier at 144 MHz; its more likely that the gain will be on the order of 10 dB.

It's only been in the past few months that the metal-oxide semiconductor FET (MOSFET) has been available at less than $2. Low cost seemed necessary since this device was reputed to be very unreliable when handled. Everyone seemed to know of the static burnout possibility of the MOSFET; very few cared about its advantages.

the mosfet at two meters

My first experiments with the MOSFET were on 2 meters—mainly because 144-MHz circuitry was available from earlier experimental converters. I will discuss an evolution of three converters—all of which had desirable characteristics—with emphasis on the last which is affectionately called, "The Winner."

A relatively uncomplicated design was derived from data-sheet test circuits and W2OKO's two-meter converter which appeared in RCA Ham Tips.[1] A single-stage 3N128 provided rf gain for a 3N141 tetrode mixer (fig. 1). Output from the oscillator-tripler is injected at gate 2 of the 3N141. Without much fanfare, I had a useful converter with lower noise than my two-stage nuvistor converter and considerably less gain. It is interesting to note that the sensitivity was superior to a similar design using 2N3823's; these are JFET's similar to the better known TIS34. Stations up to 120 miles away were heard during a brief band opening during the VHF Sweepstakes in January, 1968.



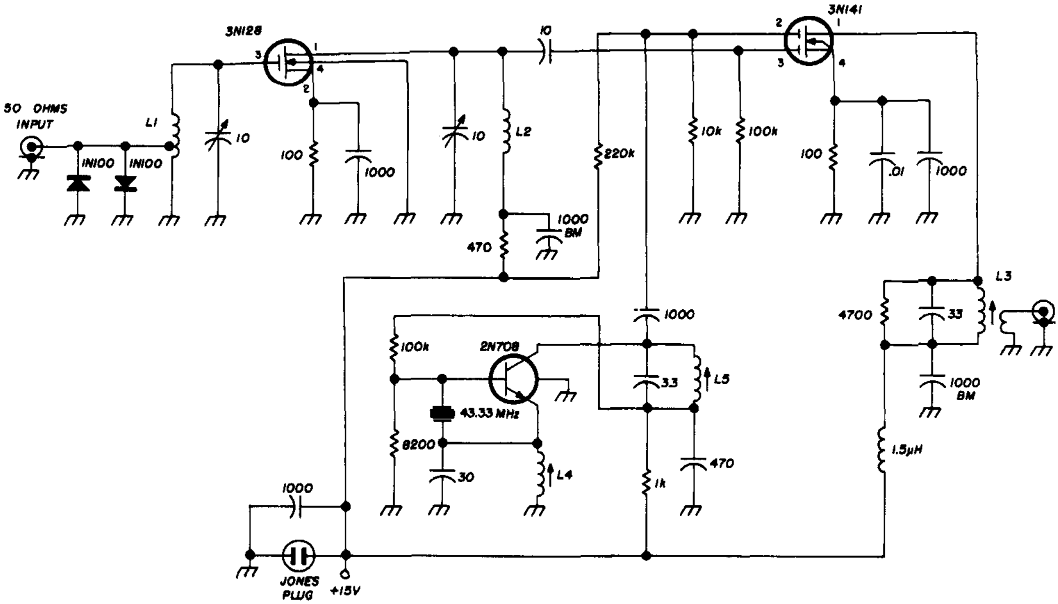
There is one noticeable difference between the JFET and the MOSFET; the feedback capacitance of the 3N128 in the common-source configuration is lower—even lower than a vacuum tube in grounded grid. Typical feedback capacitance of the 3N128 is 0.13 pF. Therefore, neutralization is not needed on two meters. The gain may be increased up to 3 dB by neutralizing, but I didn't try it.

Low feedback is, indeed, important to the MOSFET; the relatively high feedback of the JFET is the primary limitation to its gain. While on the subject of circuit stability, it has been proven using Linvill's equations[2] that the 3N128 and similar MOSFET's are stable at 145 MHz when the input and output impedances are conjugately matched.[3] A less scholarly explanation is that the circuit will "take off like a rocket" if you plug it into a six-meter antenna.

advanced circuitry using MOSFET tetrodes

The 3N128 converter was not the high-gain unit I had hoped to build. I changed the rf amplifier to a 3N140 tetrode (fig. 2), but there was no appreciable increase in gain; the advantages of a MOSFET tetrode as an rf amplifier were not immediately apparent. However, let's discuss the development of the MOSFET tetrode so we can better understand its characteristics.

When a strong signal in or near the band disturbs the signal of interest, it is called cross modulation. Conventional bipolar transistors exhibit the greatest amount of cross modulation. On the other hand, two 3N128's in a cascode circuit have a surprisingly high cross-modulation resistance.[3] Because of this characteristic, the dual-gate MOSFET was developed to incorporate the cascode circuit in a single device. Actually, the 3N140, 3N141 and similar MOSFET tet-



L1 5 turns number 18 space wound ¼" diameter. Tapped 1 turn from ground.

L2 4 turns number 16 space wound ¼" diameter.

L3 Primary: 15 turns number 30 on 3/8" diameter slug-tuned form (J. W. Miller 21A000-2 with red core). Secondary: 3 turns number 26 wound on cold end of primary.

L4 5 turns number 26 wound on 3/8" diameter slug-tuned form (J. W. Miller 21A000-2 with red core).

L5 4 turns number 26 on ¼" diameter slug-tuned form (J. W. Miller 4500-4).

fig. 1. Two-meter converter using mosfet rf amplifier and mixer stages.

rodes integrate two triodes with the drain of the first internally connected to the source of the second. The resulting device has the following desirable characteristics:

1. Improved dynamic range; this results in high sensitivity for weak signals and no overload with strong ones.

2. A second control terminal which is used for agc or mixing service; agc is useful in improving cross modulation characteristics.

3. Good stability because of low feedback capacitance (less than 0.02 pF when gate 2 is bypassed). The socket, if one is used, is the limiting factor since it has much greater capacitance than the transistor. Low feedthrough capacitance is ideal for broadband converter service and little or no neutralization is necessary.

4. Noise figure and power gain are slightly enhanced.

Armed with all this book-knowledge, I left the 3N140 in the converter and loaned it to W2BV for evaluation while I began the third design.

Performance reports on the 3N140 converter show that its noise figure is 2.5 to 2.8 dB. This indicates that the noise of the 3N140 is running well below the 3.5 dB rating of the manufacturer. Despite the less critical oscillator injection requirements of the MOSFET as compared to a JFET, I felt that the 43.3 MHz oscillator-tripler wasn't completely satisfactory. When the converter was used with a Drake R4 receiver, sensitivity was adequate, but lack of injection exaggerated 20-meter feedthrough; a separate ground wire to the converter was helpful in reducing 20-meter interference.

an optimized converter

Finally, I built the two-rf-stage converter shown in **fig.** 3. Greater oscillator injection was provided by a 65-MHz crystal-controlled oscillator and separate doubler. Since this converter has outpaced all other comparable converters, I have called it, "The Winner." Nevertheless, only time will prove its merit.

The first rf stage uses a 3N140 to capitalize on the better features of the MOSFET tetrode. Gate 2 is biased at 1.4 volts for good cross-modulation resistance; drain current is about 7 mA—this is near the point of optimum noise-figure while providing good power gain. If you want, you may be able to optimize circuit performance slightly by varying the bias on gate 2. Please bear in mind that anything higher than 6 volts exceeds the ratings of the transistor. Please remember—remove the transistor while you're changing bias resistors.

The optimum cross-modulation resistance will show an improvement of ten times over a bipolar transistor. However, at this level, gain will not be optimum for the 3N140. I use a gain reduction of 5 dB from optimum

Because of the less critical requirements of the second rf stage, I used a 3N128 triode. The exceptional gain provided by this stage allows very good broadband tuning. In fact, despite Linvill, you may have oscillation problems if you don't use stagger tuning. A lower gain FET, the RCA 40467, was tried successfully in this stage, but had poorer characteristics for broad-banding. The difference in cost—$1.24 versus $1.45—is not enough to consider the compromise in performance.

A few facts about the ubiquitous 3N141 are in order. Althought this device may be slightly inferior to the 3N140 in rf service, it is outstanding as a mixer and exhibits very high resistance to cross modulation, reasonable local oscillator requirements and high gain. As you might anticipate, conversion gain of the dual-gate MOSFET increases within limits, with greater local-oscillator drive.

L1 5 turns number 18 space wound ¼" diameter. Tapped 1 turn from cold end.

L2 4 turns number 18 space wound 'la" diameter.

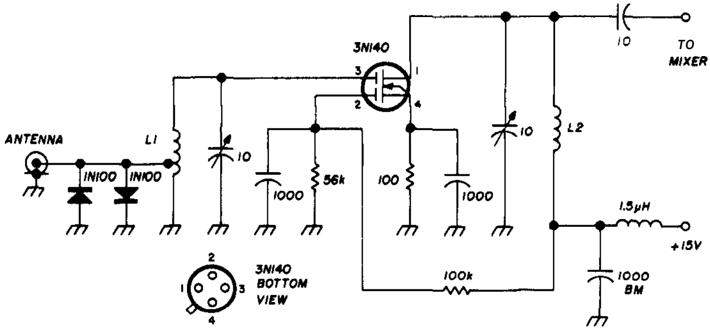


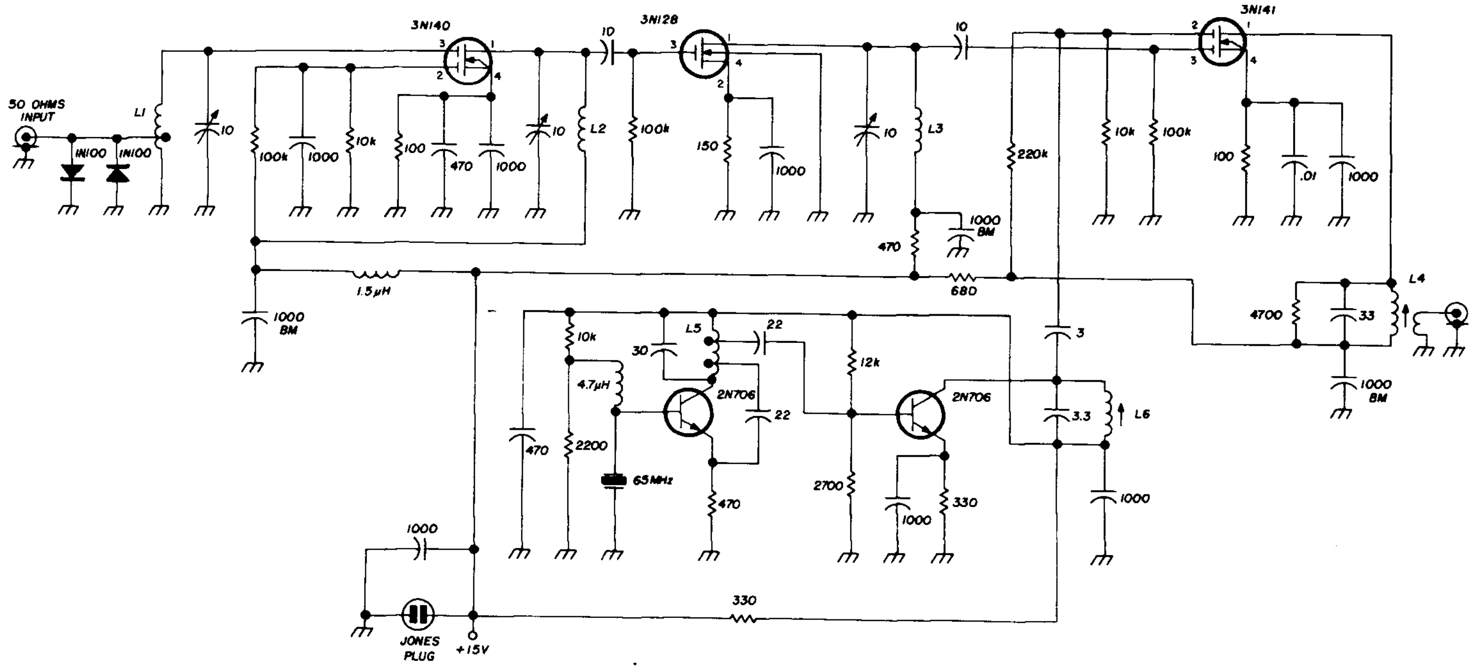
fig. 2. Tetrode mosfet rf amplifier used in the second two-meter converter. This simple change resulted in higher performance.

for the present level of operation. Expect some degradation of noise figure as well as gain.

Another technique for minimizing cross-modulation is to tap-down on the input coil to reduce the voltage swing on the gate. In most cases you'll find it necessary to tap the output coil at the same time to achieve an impedance match. I have illustrated how this may be accomplished in fig. 4, although I haven't optimized my own converter for cross modulation. The compromise is a matter of choice and depends on your proximity to high-power transmitters.

In my case, the dc bias on the second gate was optimized for the LO drive I had.

In using a 65-MHz oscillator and doubler, the number of harmonics is reduced from those present in the 43.3-MHz circuit, so the converter should have fewer spurs. The 65-MHz circuit also provides greater injection. Should cost be a factor, I feel that a 43.3-MHz crystal would probably do a good job in this circuit with only two changes; first, the tank capacitor should be increased to resonate with L5 at 43.3 MHz, and second, the 4.7 μh inductor should be changed to 6.8 μh.

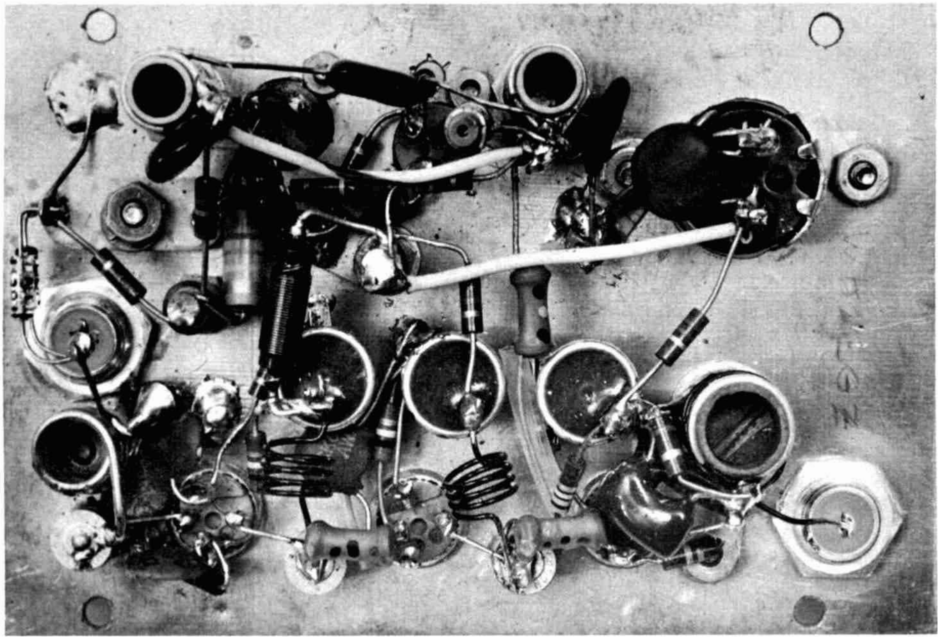


L1	5 turns number 18 space wound 1/4" diameter. Tapped 1 turn from cold end.
L2, L3	4 turns number 22 space wound 1/4" diameter.
L4	Primary: 15 turns number 30 on 3/8" diameter slug-tuned form (J. W. Miller 21A000-2 with red core). Secondary: 3 turns number 26 on cold end of primary. Note: form must be shortened 1/4" and the slug shortened 1/8" to fit into Bud chassis.
L5	5 turns number 20 on 1/4" diameter form (J. W. Miller 4500-2 with red core), tapped at 3/4 and 1-3/4 turns from power-supply end.
L6	4 turns number 26 on 1/4" diameter form (J. W. Miller 4500-4)

10-pF trimmer capacitors are JFD-VAM010 (Allied Radio 43E3475*).

fig. 3. High-performance, low-noise two-meter converter using two mosfet rf amplifier stages—the two meter winner!

* Available from Allied Radio Corporation, 100 N. Western Ave., Chicago, Illinois 60680. Order catalog number 43E3475. $3.95 each plus postage: shipping weight, 2 ounces.



Below-chassis construction of the two-meter winner.

construction

The use of small components results in tidy construction on copper-clad board. I was surprised to find that no shielding is necessary. Circuit board layouts are shown in **fig.** 5 and **6; fig.** 5 may be used for either single rf stage converter while fig. 6 shows the layout for the double rf stage unit. Both layouts are designed to mount on a Bud CB-1626 chassis.*

mosfet handling

Handling the MOSFET requires some care. At a slight sacrifice to gain, I recommend the use of sockets. General applicable rules which should be followed in handling MOSFET's are listed here:

1. Keep MOSFET leads shorted until the device is ready to use. (They are shipped this way.)

2. When cutting leads, hold the lead and case with your fingers to reduce the possibility of electrical and mechanical shock

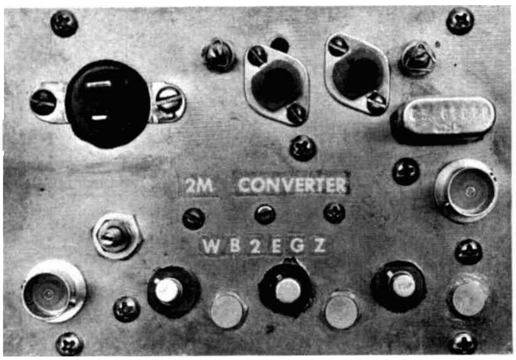
3. Don't solder or change components with MOSFET's in their sockets.

4. Never insert or remove transistors when power is applied.

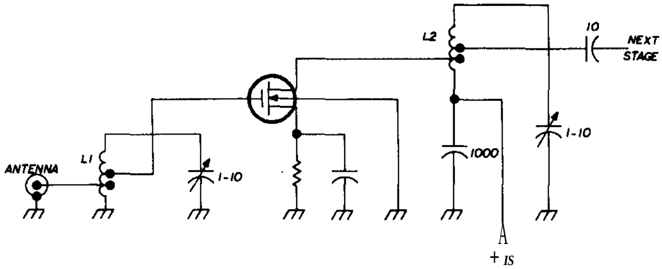
performance

This two-meter semiconcluctor circuit has performance to spare. With a noise figure between 2.5 and 3 dB, it falls short of the miraculously quiet JFET, but the MOSFET

The two-meter winner—extra performance on 144 MHz.



* The Bud CB-1626 chassis is available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog numher 42E7812, $ 55 plus postage; shipping weight, 12 ounces.



L1 5 turns number 18 space wound 1/4" diameter. Antenna tap at 1 turn from cold end. The gate tap is 1½ to 4 turns from the cold end depending on the strength of the interfering signal. Bert rejection at 1½ turns.

L2 4½ turns number 18 space wound on 1/4" diameter. Adjust drain tap for optimum gain. The tap will be approximately the same number of turns from the power-supply end as the gate tap on L1 is from ground. The tap to the next stage is optional. If used, follow the procedure for the gate tap on L1.

fig. 4. Suggested method for tapping down on the tuned circuits to optimize cross-modulation resistance. Although this technique may be used at any frequency, the values given are for the two-meter band.

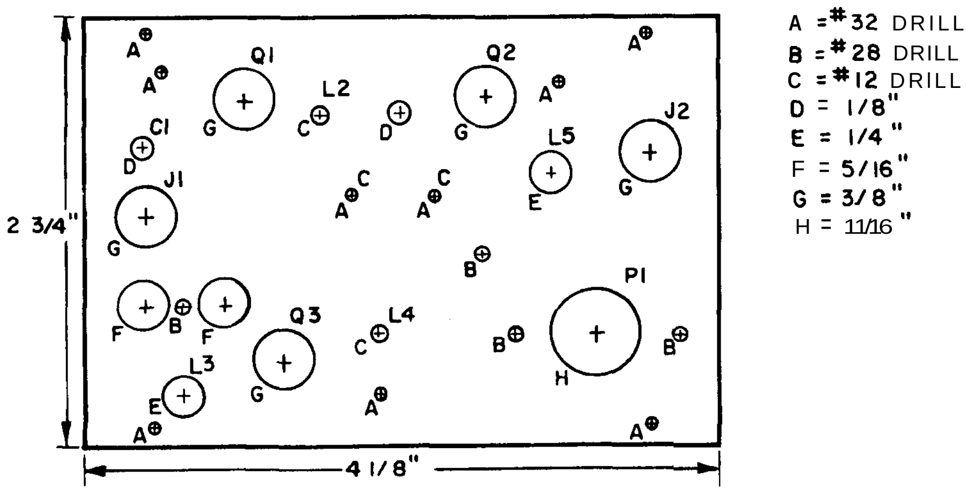
provides better hearing over wider bands. In fact, the gain-bandwidth of the five transistor converter easily surpasses that of two rf stage nuvistor converters. With this amount of gain, the circuit must be stagger tuned unless an agc system is used; 2- to 3-MHz bandwidth appears practical.

It is interesting to note that only three tuned circuits are used in the two-meter amplifiers and no shielding is needed. The absence of neutralization is a mixed blessing—tuning is greatly simplified, but over-all gain may be restricted. In the tetrode, any neutralization is for the socket and **not** the MOSFET. I think the absence of neutralization is a benefit to cross-modulation resistance because high signal levels and high gain may upset stability in a neutralized circuit.

Cross modulation in this converter is about the same as in a vacuum-tube equivalent—a decided improvement over bipolar circuits. Dynamic range, which is about 25 times that of a bipolar transistor, is slightly better than vacuum tubes. The large dynamic range is a direct result of the fact that the MOSFET gate may be operated with either negative or positive bias.

As you will see when you build one, this

fig. 5. Chassis layout for the three-transistor two-meter converters.



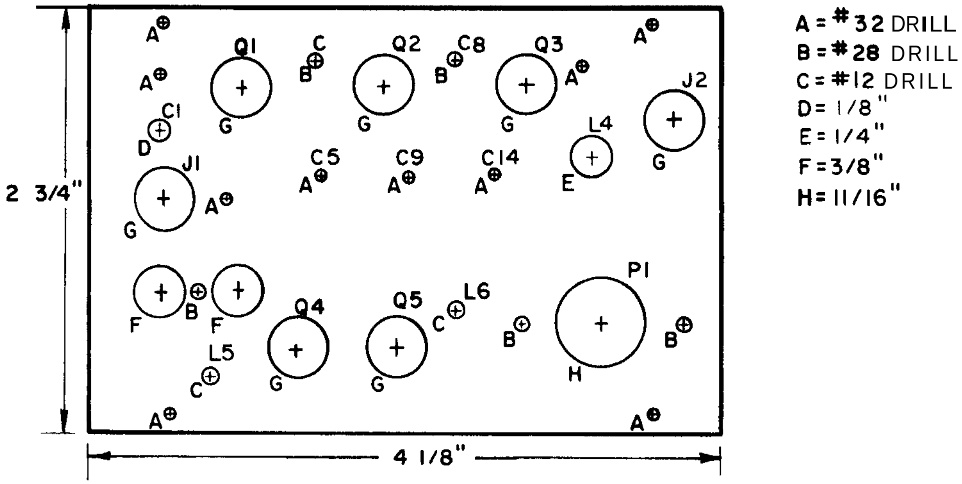


fig. **6.** Chassis layout for the five-transistor two-meter winner.

is quite an impressive converter. If you're opposed to all the gain that is available with this circuit, a single rf stage may be a good choice since it improves the dynamic range slightly. Another interesting change might be two MOSFET tetrode rf stages; in this case, the bias on gate 2 of the second stage should be the same as gate 2 on the first stage for optimum immunity from cross modulation.

Although I am extremely pleased with the operation of this converter, time has not permitted extensive testing under varying operating conditions. If you build one of these circuits, your reports and comments will be appreciated.

acknowledgements

It is next to irnpossible to accomplish a design of this magnitude without the help of many persons, each of whom has experience in some related segment of the project. Mr. Burrell Warnoch, W2BV, provided particularly important test information on the converters. Consultations with Mr. Sy Reich of RCA, Mr. Dick Peterson, WA2CYE, and Mr. Bill Schaefer, WA2EMB, were quite useful in establishing guidelines for the project. My thanks to them all.

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converting vacuum-tube equipment to solid-state

If you have an old piece of vacuum-tube equipment around the shack, here are some ideas for converting it to solid state

John J. Schultz, W2EEY, 4[illegible] Street, Mystic, Connecticut [illegible]355

Unfortunately, you can't simply plug a transistor in place of a vacuum tube. However, in some cases you can replace tubes with easily made solid-state "plug-ins." With this approach, you can often improve the operation of an old piece of tube-type equipment.

If you wanted to, you could probably replace all the vacuum tubes in a receiver or low-power transceiver with plug-in, solid-state modules. The modules wouldn't be universal—you'd have to tailor make one for each tube circuit; a solid-state circuit that would use the components already on the chassis.

It's unlikely that you would ever want to convert an entire tube-type unit to solid-state. However, there may be advantages to converting individual stages or accessories. This is particularly true for vacuum-tube stages that require frequent tube replacement. Circuits that are unstable under mechanical vibration or high temperature conditions are also natural solid-state targets. Also, putting transistors in test equipment or station accessories has several advantages—less heat and easy battery-powered portable operation.

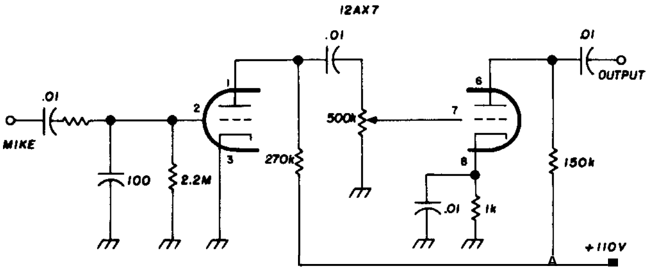


fig. 1. Typical two-stage vacuum-tub* microphone **preamplifier.**

Before you can design a solid-state plug-in, you have to consider several aspects of the stage you're working with: input and output impedances, gain, frequency, bandwidth and the power supply voltages available at the socket.

Two or more transistors may be needed to compensate for the high input and output impedance of tube stages. This also helps in the gain department. Because of the constraints imposed by the components mounted on the chassis, it is rarely possible to replace a single tube with a single transistor. Several years ago the solid-state plug-in was too costly for amateur equipment, but today there are a multitude of low-cost, high-performance transistors available. In many cases several of them cost no more than one tube. Frequency is usually no problem—at least at hf and lower vhf—since many low-cost transistors developed for the television industry exceed the capabilities of comparable vacuum tubes.

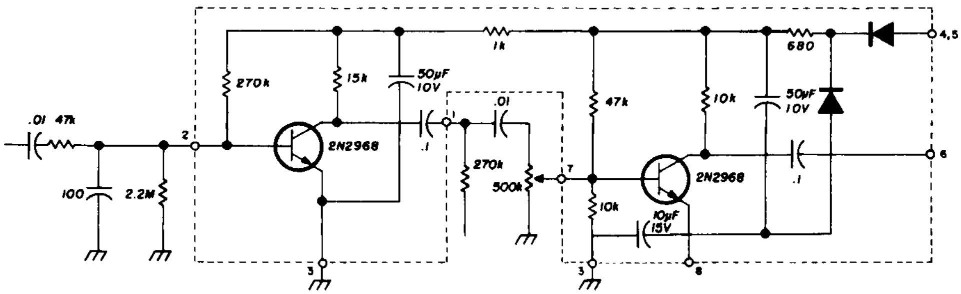
Power for the solid-state plug-in can be taken from the plate or screen supply voltage through a voltage divider. High wattage divider resistors are generally not necessary because of the low current demand of the transistors. Another possibility is the filament supply; just include a rectifier and filter in the plug-in.

In this article I'll describe solid-state conversions for several typical vacuum-tube circuits. They are not applicable to every tube circuit, but if the general methods of conversion are understood, you should be able to develop conversions for your own specific circuits. I have used the original components connected to the tube sockets whenever possible. When this hasn't been possible, only the input, output and supply voltage points have been used.

audio preamplifier

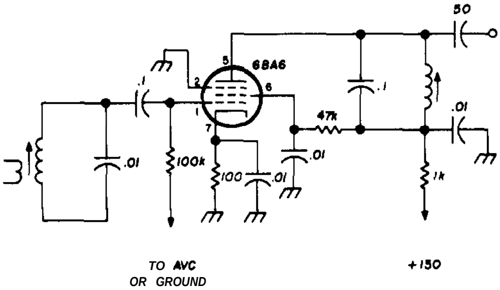
A typical two-stage high-impedance audio preamplifier is shown in **fig. 1.** It's typical of the type used in many modulators. The solid-state conversion is shown in **fig. 2.** A few of the original parts were used including the 2.2-megohm resistor which forms a bias network for the base of the first transistor and the Ik resistor in the emitter lead of the sec-

fig. 2. Tho transistor convarsion of tho vacuum-tub. circuit **shown** in fig. 1. Tho **terminal** numborr corraspond to tho tuba pin **numbers given** in fig. 1.



ond transistor. The gain control is included in the circuit by using a 0.1-μF blocking capacitor. The supply voltage is provided by rectifying the filament voltage. The transistor stages are conventional grounded-emitter circuits. Because of relatively high transistor gain, it was possible to replace each tube section with only one transistor.

fig. 8. Vacuum-tuba i-f amplifier; the grid resistor connection depends upon whether the stage Is rvc controlled or not.



i-f amplifier

Fig. 3 shows a typical i-f stage using the ubiquitous 6BA6. There are two solid-state conversions for this circuit shown in **fig. 4**—one for avc, the other without. Both are usable with i-f frequencies up to 9 MHz. None of the original components were used in this conversion because their values weren't suitable. Voltage to operate the transistors is taken from the screen supply.

To obtain sufficient gain and high input and output impedances, I used two directly-coupled transistor stages. The extra transistor in **fig. 4B** is required to provide avc action without overloading the avc bus. This is because a vacuum-tube avc circuit is not designed to supply any current. The 2N697 performs the avc function by controlling the bias of the 2N293 stage.

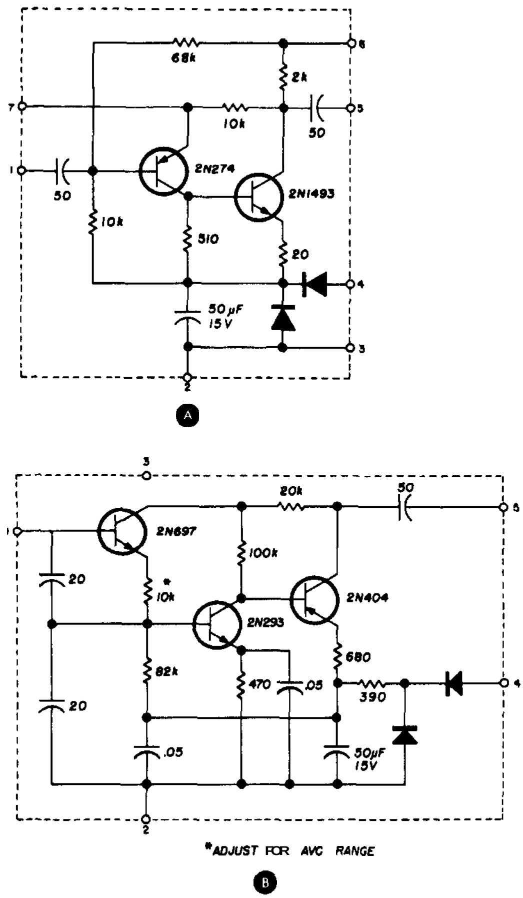
The rf signal is coupled to the base of the 2N293 through a capacitive voltage divider. These capacitors, as well as the output coupling capactor, should be increased to 100 or 200 pF for low-frequency i-f's (455 kHz). In this circuit, the operating voltage is obtained through a rectifier in the filament voltage line.

oscillator circuits

It isn't any problem to convert the simple tuned-plate crystal oscillator shown in fig. 5. The RCA 40080 transistor will operate at frequencies up to about 15 MHz with reasonable output. However, you may have to retune because of the change in output capacitance.

The conversion of a vfo circuit is a little more complicated as illustrated in **fig. *6.*** Two transistors are required to provide the necessary feedback and output. The calibration of the vfo will probably change so you'll have to recalibrate it. If the calibration is too far off, you may have to put new padding capaci-

fig. 4. Solid-stat. convarsion of a vacuum-tuba i-f amplifier. The circuit In B incorporates avc, while A does not.



tors across the tuned circuit to retain the original dial calibration.

limiter circuit

The limiter stage shown in fig. 7 illustrates a circuit which uses a dc feedback loop. The solid-state conversion of this circuit is somewhat limited by the original feedback circuit. The conversion shown in fig. 7B uses three capacitors.

The 0.1-μF input coupling capacitor isolates the ac and dc components present on pin 1 of the tube socket. The ac signal is amplified by the two-stage 2N2968/2N1305 amplifier. The dc feedback voltage controls the collector current of the first 2N2968, thereby producing limiting action.

construction

One simple way of building a plug-in is to mount the components on a piece of Vectorboard which is wired to a plug to fit the tube socket. If you have an old burned-out tube around, you can use the base after breaking the glass envelope. For miniature 7- and 9-pin sockets, Amphenol type CP plugs are available.*

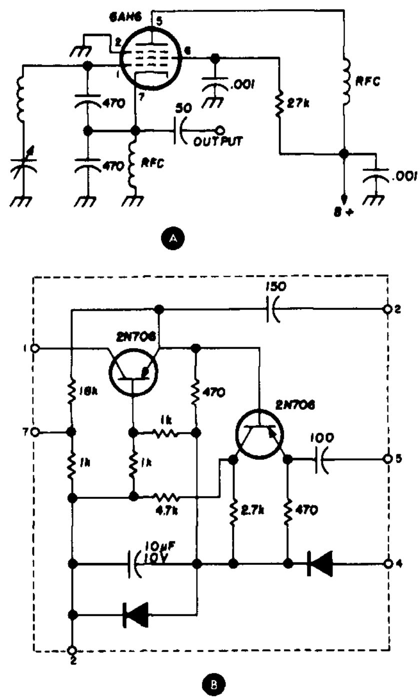
If you want, you can buy complete plug-in assemblies with a plug, perforated phenolic board and a shield can (Vector G2.1-8-4). These units are a little expensive, but they may be worth while if you want to convert a sensitive circuit where shielding is necessary.

*** Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. For the 7-pin miniature plug, order 47E2649, $.99; 9-pin, order 47E2652, $1.38; plus postage; shipping weight 2 ounces.**

summary

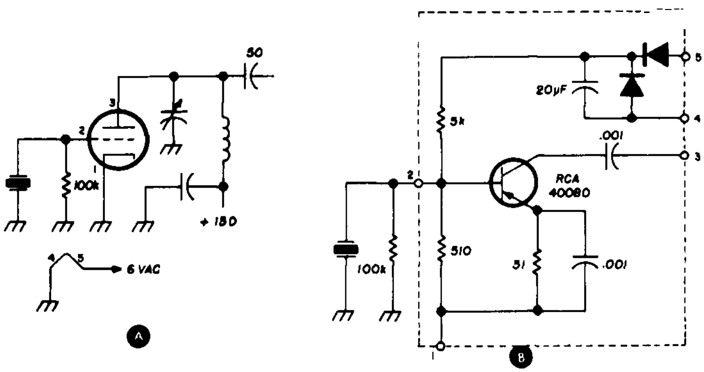
I have tried to present some ideas on how a variety of tube-type circuits may be updated to solid state. It should be pointed out, however, that the conversions shown

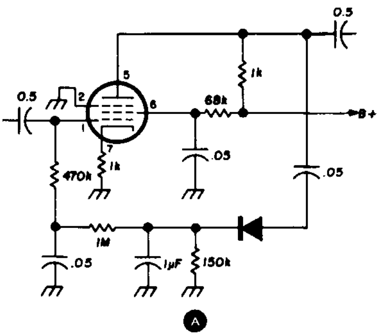
fig. 6. Clapp vfo circuit, vacuum tuba (A) and solid-state (B).



were designed only for the tube circuits shown. They may not necessarily work with the same tube used in other, similar circuits, although they provide a good starting point

fig. 5. Vacuum-tuba and transistor version of a low power crystal-controllad oscillator.





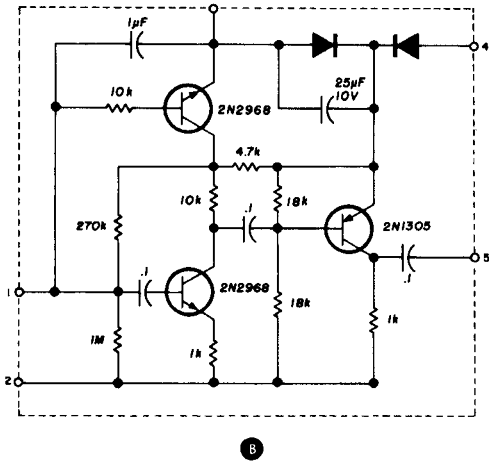


fig. 7. A limiter or compressor circuit (A) and a possible solid-state conversion (B).

when developing a solid-state replacement; simply adjust the component values for optimum performance.

The conversion of vacuum-tube power circuits is also possible, but it's complicated by several factors. First of all, power transistors usually require a medium-voltage (15 to 80 volts) high-current source; this is difficult to arrange in most vacuum-tube units without changing chassis-mounted components. Also, the problem of providing an adequate heat sink makes construction rather difficult. Lastly, the input/output impedance conditions for rf power transistors is quite critical and difficult to achieve without rather complicated circuitry. Although individual transistors are shown here, integrated circuits may be used in vacuum-tube conversions and should be considered if the proper types are available at reasonable prices.

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short circuits

WB2EGZ six-meter converter

Don't ground L3 as shown in **fig. 1** on page 23 of the June issue—it shorts out the power supply! On page 20, ignore the reference to the 2N2708—it's a 2N708 as shown in the schematic in **fig. 1**.

vhf fm

We have been advised by the Lynchburg Amateur Radio Club that they have not published their list of fm nets since March, 1964. The supply of these lists has long since been exhausted. If you have written to them, you'll receive a reply only if return postage was included with your letter.

K4ZAD also advises us that the Motorola Private line is a continuous low-frequency tone system, and not a tone-burst system as indicated in the article on page 90 of the June issue.

amateur aneometer

The μL914 integrated circuit used in the aneometer circuit on page 53 of the June issue is manufactured by Fairchild Semiconductor; sorry we didn't mention it in the article. These devices are available from John Meshna, Jr., P. O. Box 62A, East Lynn, Massachusetts 01904. $1.00 each postpaid airmail. Order 2-914, quad 2-input gate; the 2-914 contains two of the μL914's.

stable transistor vfo's

In **fig. 1**, page 15 of the June issue, the numbers on C1 and C2 should be reversed to agree with the formula on page 20.

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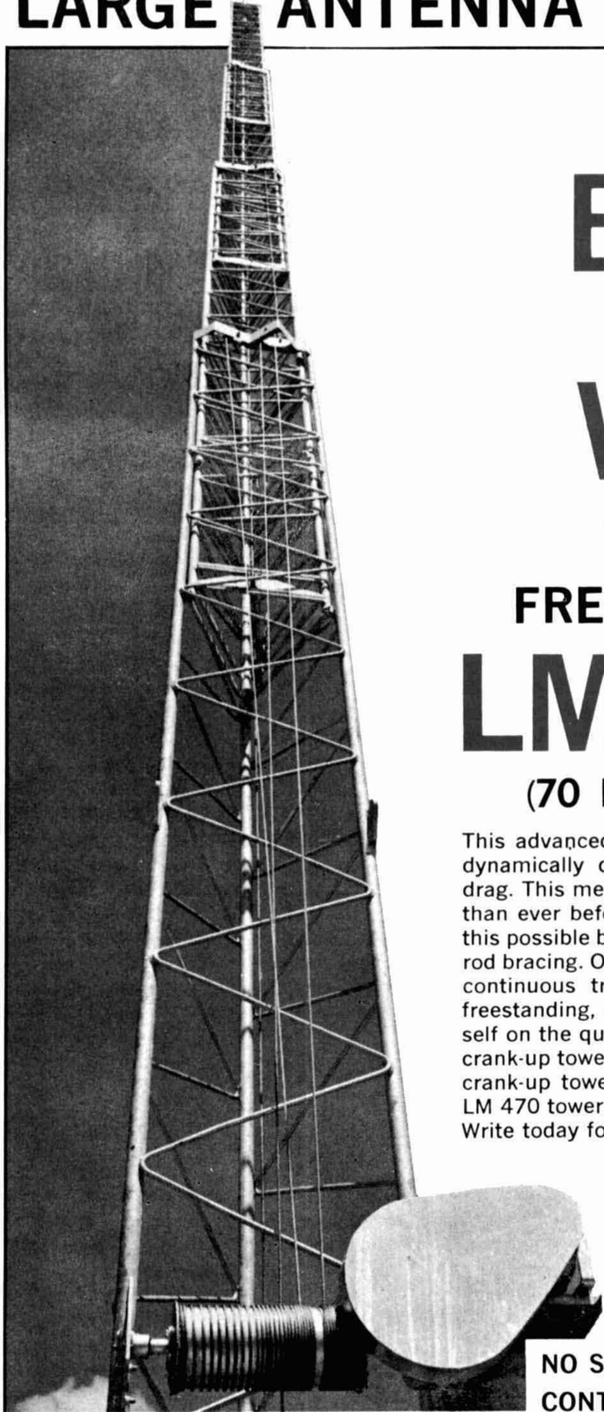

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VRC19 receiver and squelch module before modification to solid state.

solid-state audio amplifier and squelch circuit

Although this audio amplifier and squelch circuit was designed for a solid-state VRC-19 receiver, the same approach may be used with other transistor FM receivers

nk P . W6A JF,6A 850 Donner Avenue S , Ca f .

The high noise levels experienced with a high-gain NBFM receiver during standby periods are always objectionable. Some form of effective noise squelch system is needed. The system shown here was developed in connection with the conversion of a vacuum-tube NBFM receiver to solid-state operation. With this particular receiver, the VRC 19, the squelch can be set to operate properly for signals of 2 microvolts or more. The switching transistor in the squelch unit snaps the audio amplifier on and off very nicely. Although this circuit was developed for the converted VRC-19, it may be used with other solid-state FM receivers.

The VRC-19 receiver has an audio and squelch system in one plug-in module which uses subminiature tubes with a larger miniature tube in the output stage. My task was to redesign the unit for transistors and diodes for use with a 12-volt storage battery.

The required audio output power of one-half watt requires at least 100 milliamperes of current at low voltage rather than 10 or 15 mA at 150 volts encountered with vacuum tubes. A class-A power amplifier running at 200 mA or so at 12 volts, or a class-B system running up to 100 mA at about 9 volts will provide approximately one-half watt output. I decided that an integrated circuit could be used for the complete audio amplifier job. The RCA CA3020 IC contains 7 transistors, 3 diodes and 11 resistors in one 12-lead transistor can for about $2.50. This IC requires about 0.05 volt of drive to produce one-half watt output; this is available through an RC network from the discriminator transformer in the i-f unit of the VRC-19 receiver.

The original three-tube squelch system was replaced by a couple of low cost plastic-cased silicon NPN transistors—GE 2N2711's.

the circuit

The noise voltage appears on the low side of the discriminator input circuit when no signal is present. A surplus 44-millihenry telephone toroid and a .02-μF shunt capacitor select the 5- to 6-kHz noise components to actuate the squelch system. Since voltages in the voice range are also present in this limiter stage when a signal is present, this filter must be used to prevent voice signals from operating the squelch circuit.

If a 44-mH coil and .02-μF capacitor aren't available, an 88-mH toroid and .01-μF capacitor may be used. A 30-mH rf choke and .02- or .03-μF capacitor should also work for this purpose, although a toroid has less external field and lower dc resistance.

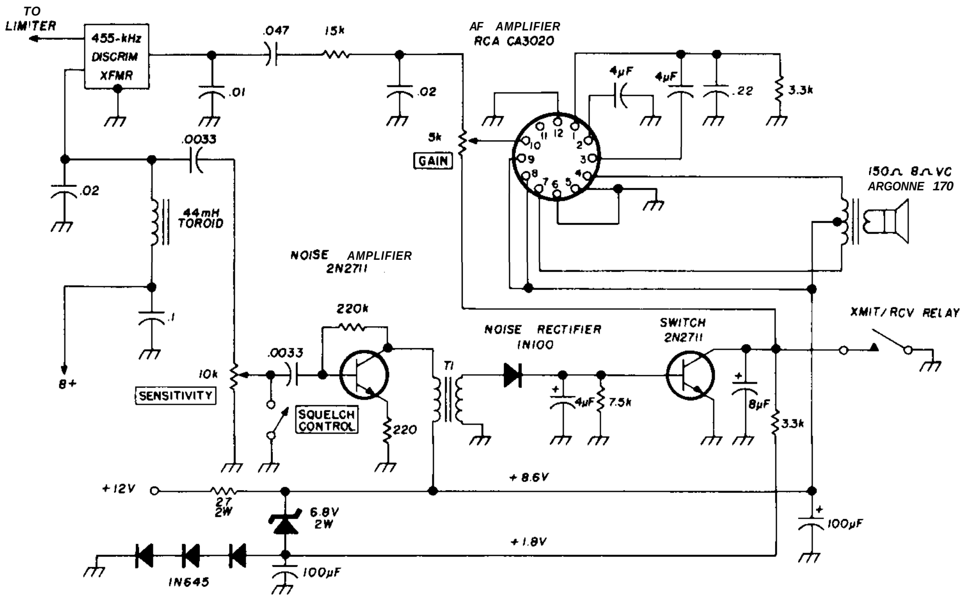


fig. 1. Schematic diagram of the solid-state audio output and squelch circuit used by W6AJF.

Any NPN devices with an h_{fe} of over 30 can he used in the circuit shown here. This circuit picks off a little 5- to 6-kHz "hiss" noise from the last i-f or discriminator transformer and amplifies it with a high-gain audio amplifier stage through a squelch gain control and low-Q tuned circuit. The original half-megohm squelch control in the receiver was changed to a 10,000-ohm potentiometer and the tube circuit rewired to use the new components.

Added low frequency attenuation is provided by the .0033-μF coupling capacitors associated with the squelch sensitivity control. As a result, voice signals are attenuated enough so that only a noise signal will operate the squelch system.

The noise amplifier drives a diode rectifier to provide a dc voltage of about 0.25 volt into the switching transistor which cuts off the input amplifier of the CA3020 IC. As long as a noise voltage is present (no signal), the af amplifier is squelched, and the loudspeaker is quiet. When a signal of very few microvolts is present, the noise voltage disappears, the squelch voltage drops, and the audio amplifier functions normally.

construction

The toroid, the .02-μF capacitor and .0033-μF coupling capacitor are located in the i-f module of the VRC-19. The 47k resistor on the output side of the discriminator transformer is shorted out and the output shunted with a .01-μF capacitor to provide an audio signal at the af amplifier.

The CA3020 is mounted on a piece of perforated 1-1/2 x 4-inch phenolic board with all the squelch and other components. All the original parts, including the output transformer, were removed from the module and a new 5000-ohm audio gain control installed in place of the old half-megohm control. With some rewiring of the power-plug leads, 12 volts is available in place of 150 volts.

A fixed voltage of + 1.8 volts is provided for the CA3020 input stage through the squelch system and af gain control by the voltage drop across three 1N645 200-mA silicon diodes. Each diode provides about 0.6 volt regulated from 1 or 2 mA up to 200 mA. Any small silicon diodes with a current rating of 150 to 500 mA should be satisfactory.

A two-watt 6.8-volt zener diode in series with the three diodes provides the regulated 8.6-volt supply to the CA3020 IC. Since the CA3020 has a maximum rating of 9 volts, it can't be used safely with a 12-volt supply. A regulated supply of under 9 volts output is needed because the IC draws 10 or 15 mA resting and up to 135 mA at full output.

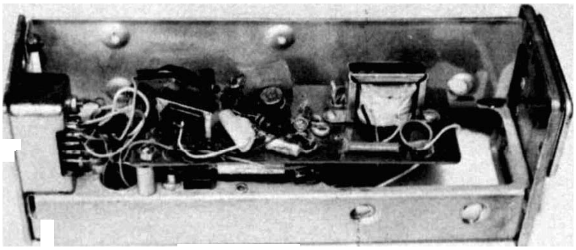
There were some problems with audio "motorboating" when the CA3020 was used with high resistance gain controls and small value bypass capacitors in the regulator and squelch circuits. However, the values shown in the circuit seem to cure these problems.

The CA3020 amplifier and Argonne 170 output transformer seemed to amplify the higher voice frequencies too much. This was overcome by using .22 μF instead of the recommended .01 from terminal 1 of the CA3020 to ground, and by using fairly large shunt capacitors in the discriminator output circuit. The high audio roll-off is greater than that used in the original vacuum-tube amplifier.

The noise rectifier can be nearly any kind of diode, silicon or germanium, since it functions at low voltage levels. The switching transistor base resistance varies from a very high value with squelch cut-off to a few hundred ohms while squelching. This means that the ac impedance of the noise rectifier is also low during operation.

A small audio step-down transformer is needed between the noise amplifier and the rectifier; a small 20k:2k or 10k:1k audio interstage transformer is suitable. The input impedance of the noise amplifier is increased to a suitable value by not bypassing the emitter bias resistor. It may be necessary to increase the value of the 220k base-bias resistor shown in the circuit with other types of NPN transistors. If the voltage drop across the 220-ohm emitter resistor is between 0.25 and 0.5 volt dc, the base bias voltage is ok.

VRC19 receiver af and squelch unit after changing it to solid state. Squelch parts at left and front edges of bakelite board mounted abova old chassis. Af at center and right side.



This squelch circuit should function quite well when connected into any NPN emitter follower of any audio amplifier in which the second audio transistor is coupled to the emitter of the controlled input stage. This type of amplifier has less than unity gain, but exhibits fairly high input impedance when compared to a base input stage with gain.

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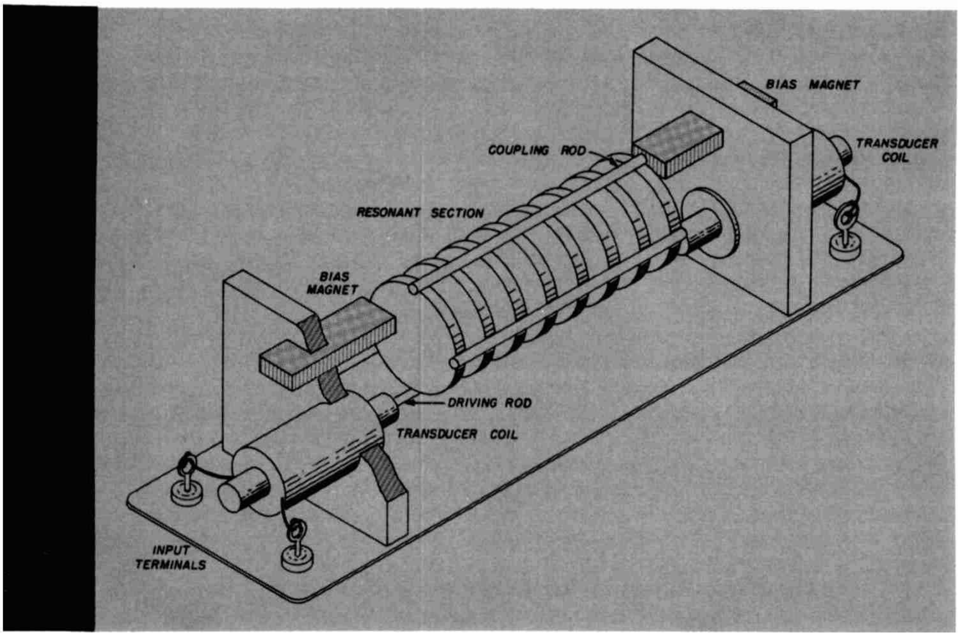


single-sideband filters

The filters required for generating a ssb signal fall into two general categories, crystal and mechanical; here is a description of both

Forest H. Belt, 119 3 k Ave ue, South Plainfield, New Jersey 07080

In a single-sideband transmitter, the signal coming from the balanced modulator is not yet an ssb signal; it still has both sidebands. Although the carrier has been suppressed, it is called a double-sideband suppressed-carrier signal. The job of removing the unwanted sideband is left to a device known simply as a **filter.**

In a communications receiver, incoming signals must be sorted out by the tuned circuits of the rf and i-f sections. These coil-capacitor combinations may allow adjacent signals through almost as well as the desired ones; their response is too broad. Removing those unwanted "side" frequencies is the job of a filter.

In a single-sideband receiver, every conversion the signal goes through generates an unnecessary extra sideband because of the nature of the heterodyning process. To recover the modulation, the ssb detector needs only the original sideband. The job of eliminating the unwanted sideband is turned over to—you guessed it—a filter.

On the schematic diagram of a modern ham receiver or transmitter, the filter is identified merely by a box labeled F1, F2 or FL1, FL2, etc. The filter circuit is almost never shown. Nor is the type of filter indicated. If you look inside the enclosure, you learn very little more. The filter is a sealed "black box" that has been plugged in or wired into the circuit. What's in it remains a mystery.

More important, of course, is what it does. From that standpoint, you can think of a filter as a three- or four-terminal device which certain signals are fed into, and out of which they come in some altered form. It is only a "black box" for all practical purposes. First, then, let's examine what filters do in ham receivers and transmitters; then we can explore what's in them.

shaping the curves

An important characteristic of any communications receiver is its **selectivity.** That is its ability to reject signals on either side of a desired one, while passing **it** freely. The selectivity you need depends on the kind of signal—that is, the modulation it carries.

A CW signal, for example, has no modulation. It therefore has no sidebands on each side of the carrier. It consists of the carrier alone. A receiver with very, very narrow selectivity can pick up the CW carrier (keyed, probably, for code transmission) and avoid interference from other, nearby carriers. One of the best ways to attain such selectivity is with a special filter in the i-f amplifier. The curve in **fig. 1A** graphs the response of an i-f section using a very narrow CW-only filter.

Filter characteristics are rated by their response at two points: 6 dB and 60 dB below maximum. The response in **fig. 1A** is 400 Hz wide at the 6-dB point and 2 kHz wide at the 60-dB point. The ratio between the two bandwidths is called the **shape factor.** The filter in **fig. 1A** has a shape factor of 5. A low shape factor means the skirts are steep, making the filter respond strongly to the desired frequencies—within the passband—and deeply reject signals on either side. A shape factor of 1 is, of course, ideal; the skirts are vertical. In practice, a shape factor of 2 or 3 is acceptable.

Consider the selectivity required for an a-m signal. Since both sidebands are needed for proper demodulation, the receiver must pass a bandwidth of at least 3.5 or 4 kHz—enough for intelligibility. That width passes sideband products for voice signals up to 2 kHz. The filter represented by the response graph in **fig. 1B** is for a-m reception. It has a 6-dB response width of 3.75 kHz and a 60-dB response of 10 kHz. The shape factor is about 2.5—giving fairly steep skirts.

Both curves in **fig. 1A** and **1B** represent responses centered in the i-f passband. For example, if the i-f is 3.395 MHz, the filter in **fig. 1A** responds well to frequencies from 3.3948 to 3.3952 MHz. The 400-Hz response is spread 200 Hz to each side of the i-f center. The i-f section with the filter of **fig. 1B** (3.75 kHz wide) responds well from 3.393125 to 3.396875 MHz. The filter rejects frequencies above and below; nearby signals cannot get through.

For single-sideband reception, the selectivity of the receiver can be narrower than for a-m, since only one of the sidebands is present. A bandwidth of 2.5 kHz is plenty. The curve in fig. **1C** shows the response of the filter in one commercial ssb receiver. Its 6-dB response is 2.1 kHz; 60-dB response is

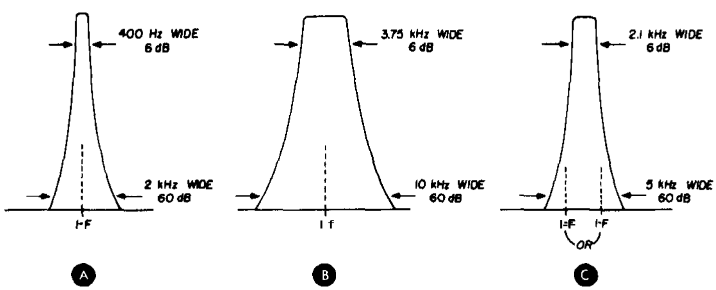


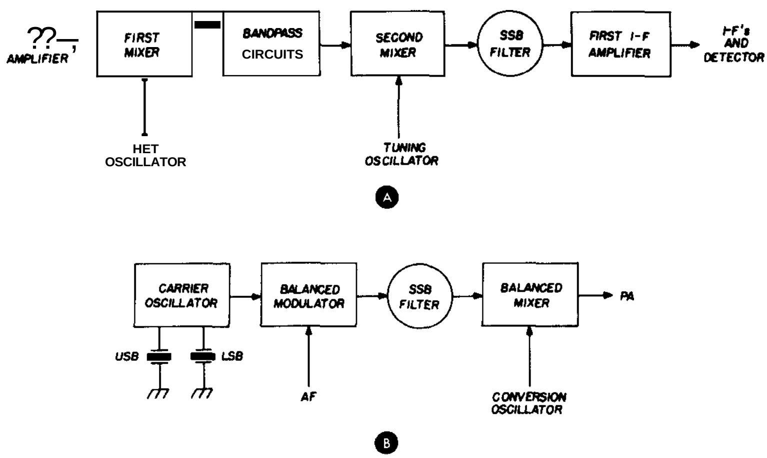
fig. 1. Response curves for typical filters used in ham radio equipment. A for cw reception; B for a-m and C for ssb reception. 2.1-kHz bandswitch is standard.

5 kHz. The shape factor is about 2.4.

There's something else special about the ssb filter in **fig. 1C.** Its response is not centered on the i-f. The "center" of the filter's bandwidth is off to one side or the other of the i-f, placing any i-f signal down on either skirt, below the 6-dB point. Which skirt is chosen depends on which sideband must fall within the bandpass. If the upper sideband must be amplified, the i-f is placed on the lower-frequency skirt of the filter response.

In a superheterodyne ssb receiver, every frequency conversion creates two sidebands from the single-sideband signal. That's because the local oscillator signal beats with the incoming sideband and produces both sum and difference frequencies. Following the i-f amplifier stage, only one sideband is needed for demodulation. The frequency of the filter **(fig. 2A)** is offset from the i-f as already described, to eliminate the extra sideband that has joined the desired one.

fig. 2. Ssb filters in receiver (A) and transmitter (B).

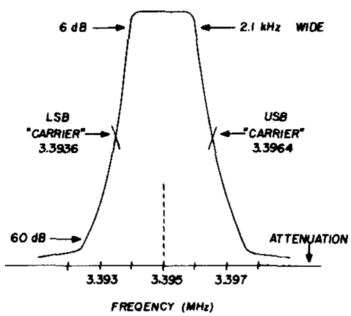


The rejection characteristic blocks the lower sideband.

filters at work

For ssb, a filter like the one in fig. 1C can be used in a transmitter or receiver. In modern transceivers, a single filter is used for both. Let's see how and why.

fig. 3. If the dsb "carrier" (which is suppressed) is placed in either position shown, one sideband is eliminated.



In the ssb transmitter, the chief job of the filter **(fig.** 28) again is to eliminate the unwanted sideband. If well designed, it also removes any vestige of the carrier that might be left by the balanced modulator. Succeeding stages of frequency translation re-create a double sideband, but the two are far enough apart that it is easy to get rid of the unwanted one with ordinary tuned circuits.

In a transmitter, there is also a need to switch from one sideband to the other. With a single filter, this is done by shifting the frequency of the carrier oscillator. Then, the signal that reaches the filter is on the other skirt. The sketch in **fig. 3** gives you some idea how this works. If an upper sideband is desired, the 3.3964-MHz USB carrier-oscillator crystal (fig. **2B**) is activated. Even though the carrier is eliminated by the time the signals reach the filter, the upper and lower sidebands fall on each side of the position shown (**USB "carrier")** on the upper skirt of the filter response curve. The sideband frequencies higher than the "carrier" are attenuated drastically; those below are amplified. (Don't be alarmed that this is called the upper sideband; when the signal passes through a stage of conversion after the filter, the heterodyning process will "flip the sideband over," making it an upper sideband in the transmitter output.)

For a lower sideband, the other crystal is activated, generating the 3.3936-MHz carrier. The sidebands fall on both sides of the position indicated as **LSB "carrier."** The sideband on the upper side is amplified, and that on the lower is attenuated. Again, the ensuing frequency conversion flips the signal over and produces a lower sideband at the output frequency.

commercial circuits

There are about as many ways of using filters in ssb equipment as there are designers who develop the gear. A few of these are diagramed in **fig. 4.** They are merely examples of different ways filters are used in ssb receivers.

The circuit from the Drake R-4B has a variable-selectivity filter, with its center at **5645 kHz.** The filter can be switched to four different bandwidths: 400 **Hz, 1.2** kHz, **2.4** kHz, and **4.8** kHz. The first is for CW; the last is for a-m; the others are for various ssb reception conditions. Not shown is an additional phase control that affects ssb reception.

Another way to obtain variable selectivity with a single filter is shown in the diagram from the Hammarlund HQ-145 receiver. The filter in this case is the simplest imaginable —a series-resonant crystal centered at the i-f. The widest bandpass is obtained with the crystal shorted out—the OFF position of the switch; it is for a-m reception. In posi-

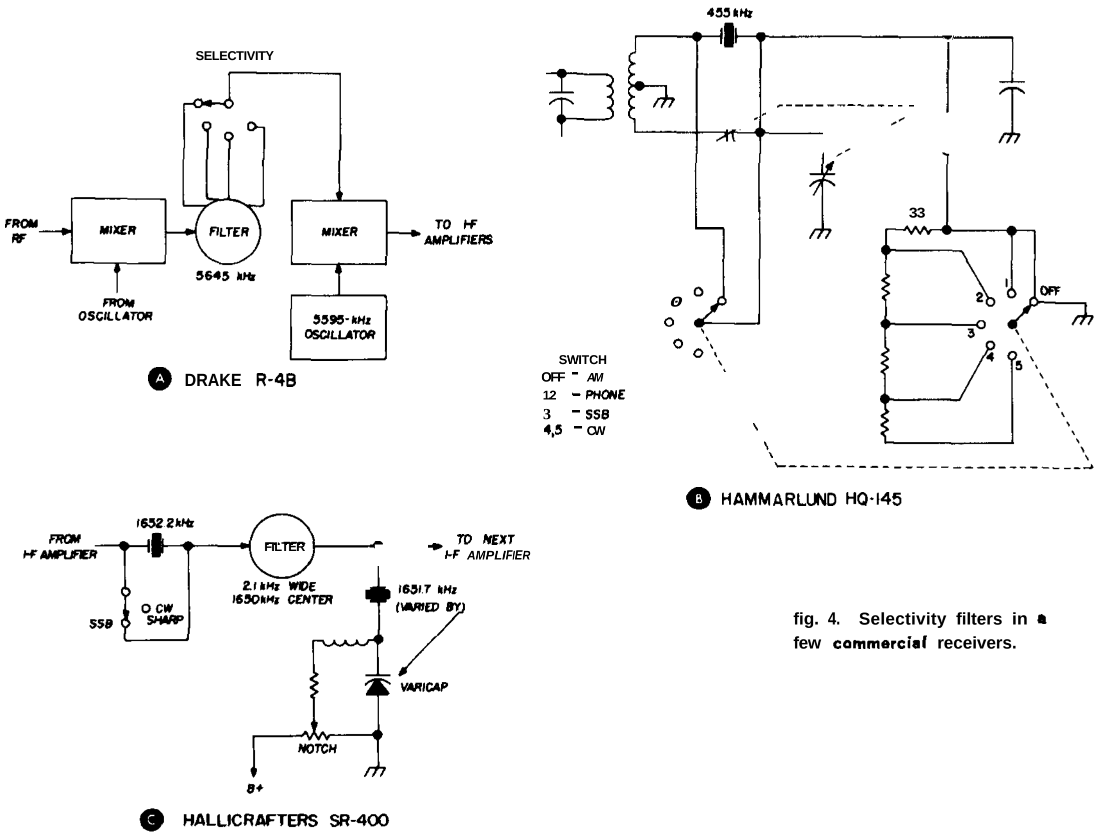


fig. 4. Selectivity filters in a few commercial receivers.

tion 1, the crystal is in the circuit, but the response of the output tuning coil is at its broadest, thus loading down the crystal. In successive positions, additional series resistance is switched in, reducing the loading effect of the output circuit and making the crystal's effect sharper. At the 4 and 5 positions, selectivity is too sharp for ssb reception, but is excellent for "notching out" interference on CW.

A fancier notch-filter circuit, combined with a conventional selectivity filter, is used in the Hallicrafters SR-400. The filter centers at the i-f, 1650 kHz. For CW reception, the CW-SSB switch is opened, inserting a 1652.2-kHz crystal in series with the signal path to the main filter. The relationship between the two frequencies narrows the over-all bandwidth to less than 1 kHz; the series crystal bucks the filter's repsonse near the upper skirt. The notch filter, which places a deep notch or dip in the pass band of the main filter, is a 1651.7-kHz crystal. The notch crystal's frequency is varied by a varicap (voltage-variable capacitor), which permits moving the notch back and forth. A NOTCH potentiometer applies voltage to the varicap to control its effect on the crystal.

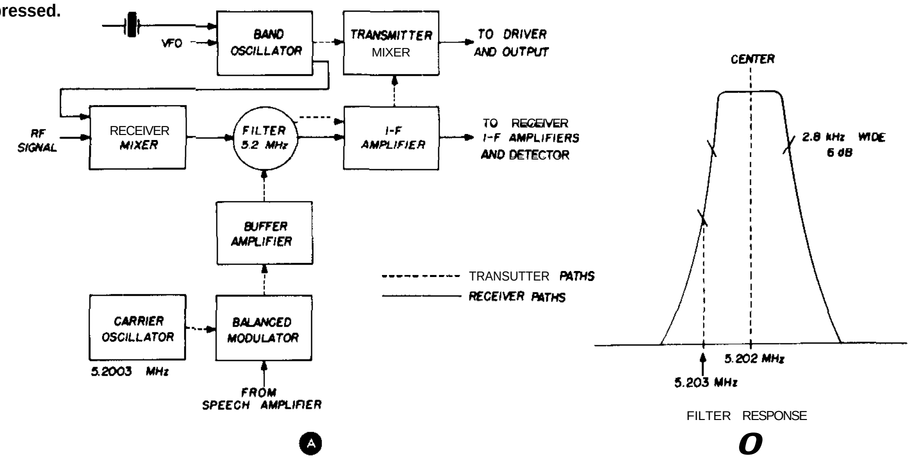
one filter, two jobs

In ssb transceivers, a single filter is frequently used for both transmit and receive. The National 200 transceiver contains a good example of this, diagramed in fig. *5.* The filter is centered on 5.202 MHz. The local oscillator frequency differs from the incoming signal frequency by that amount, thus creating an approximate 5.2-MHz i-f. The specifications for the filter, which is a type called **crystal lattice,** list its 6-dB bandwidth as 2.8 kHz. The solid lines show the signal paths during ssb reception.

During ssb transmission, shown by dashed lines, the same filter and one of the receiver i-f amplifiers are used. Sideband elimination comes from feeding the double-sideband suppressed-carrier signal (produced by the balanced modulator) through the filter. The graph of filter response in **fig. 5** shows how. The carrier oscillator operates at 5.2003 MHz. The two sidebands coming from the balanced modulator are on each side of that frequency. With the filter bandwidth 2.8 kHz wide, and its center at 5.202 MHz, the 6-dB point on either skirt is 1.4 kHz away from center. The 5.2003-MHz carrier is 1.7 kHz below the center of the filter response, placing it below the 6-dB point of the lower-frequency skirt. This position assures additional suppression of any remaining carrier, and complete obliteration of the lower sideband. The upper sideband, on the other hand, is at the peak of the filter response, and passes through unattenuated.

As you can see, the ssb signal is then amplified by an i-f stage before it is applied to the final mixer for translation up to the output frequency.

fig. 5. Using one filter for both receive and transmit. Filter response curve shows how the sideband is suppressed.



inside the black boxes

There's a natural curiosity about what is in a filter. Truly, there need be no mystery. Hams have been building their own filters for many years. Nowadays, the shortage of cheap surplus crystals has slowed down that sort of experimentation; also, commercial units are less costly. Nevertheless, it's nice to know what goes on inside your equipment and what makes it happen.

There are three kinds of filters in ssb rigs: **LC, mechanical,** and **crystal.**

The **LC filter** is, as its name suggests, a coil-capacitor combination. Several high-Q tuned circuits, cascaded, can have a response with very sharp peak and steep sides. At frequencies around 50 or 60 kHz, such filters may suffice. Below that, component size is a problem. Special designs overcome some limitations. For example, Barker & Williamson has a model—the 360—that uses toroid inductance windings and silver-mica capacitors; the center frequency is 18.5 kHz, with a 3-kHz pass band.

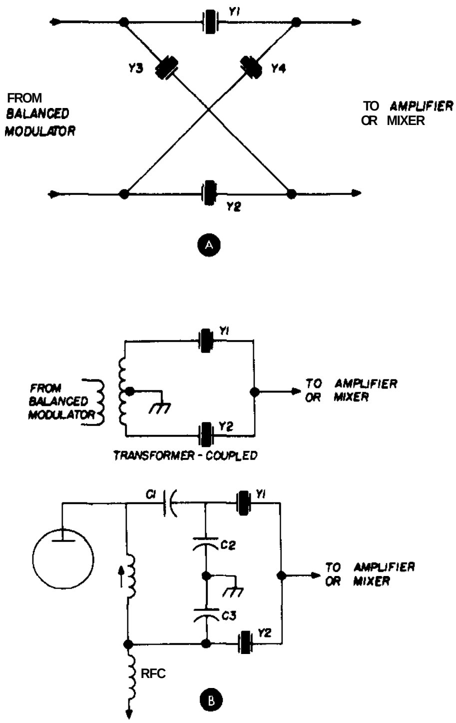
Above 100 kHz, the Q of LC components may not be high enough for practical filters. Commercially available LC filters are usually in the 50-kHz region. The Hammarlund HX-500 transmitter, which generates its primary carrier at 60 kHz, uses an LC sideband filter following the balanced modulator. Burnell's model S-15000 LC-type filter has its steepest slope at 50 kHz.

At 100 kHz and above, the **mechanical filter** becomes practical. Technically, it is electromechanical. A mechanical filter consists of an input coil tuned to the center of the i-f; a magnetostrictive transducer that converts i-f signal energy to mechanical energy; a "stack" of plates, rods or discs which are mechanical resonators coupled together by a coupling rod in their center or along their edges; an output transducer that converts the mechanical vibrations back to energy; and an output tuned circuit to couple the output signal to the next stage. A sketch of a mechanical filter is shown at the beginning of the article.

The best frequency for mechanical filters is around 250 kHz, although models are available from 50 kHz through 600 kHz. Almost never do you find one above 1 MHz. The response shape and center frequency depend on the size and shape of the resonator elements. At frequencies below 100 kHz, the elements are too large for practicality; above 600 kHz, the close physical tolerances that are necessary become too expensive to achieve.

By careful selection of the sizes of resonating elements, the bandwidth of a mechanical filter can be sharply controlled. Also, the very nature of this kind of resonant shaping insures extremely steep response skirts—almost vertical, which means an extremely low shape factor. One of the better-known mechanical filters is the Collins F455Y-31; its center frequency is 455 kHz, its 6-dB bandwidth is 3.1 kHz, and its shape factor better than 2.

fig. 6. Lattice-type crystal filters for ssb. Full lattice, or ring, circuit in A and two forms of half-lattice filters in B.



crystals in lattice networks

From 1 MHz up, ssb filters are most likely to be of the crystal variety. As you've seen already, crystal types are available below that, but most are above 1 MHz. They are costly, but offer excellent performance.

One unit that has become popular in commercial equipment lately is the 9-MHz crystal-lattice filter. Its cost ranges toward $50. For single sideband use, the bandwidth is standard: 2.1 kHz. The 9-MHz filter is used in the Galaxy V Mk2 transceiver, in the Hallicrafters SX-146 receiver and HT-46 transmitter, and in the Gonset 910-series transceivers –that I know of. In the transceivers, a single 9-MHz filter is used for both transmit and receive. One commercially available 9-MHz crystal filter is the McCoy 32B1; it retails for $35.00.*

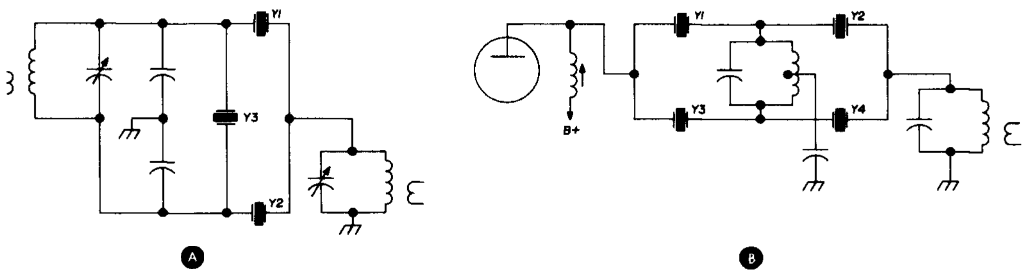
Two basic circuits are the foundation for all crystal filters (other than simple series crystals already described). Both are called lattice **networks,** and are shown schematically in fig. **6.** The one in fig. **6A** is called a full lattice and the two in **6B** are called half-lattice. The full lattice has also been called a ring filter, due to its resemblance in configuration to the well known ring diode modulator. The two half-lattice crystal filters in fig. **6B** differ only in the manner of coupling signal energy to the crystals.

In fig. **6A,** crystals Y1 and Y2 are a matched pair, series-resonant at the center frequency of the filter. They pass along the signals at resonance, but not signals on either side. Broadening the response of the filter are Y3 and Y4. In some filters, they are parallel-resonant at the center of the filter pass band; in others, they are series-resonant at frequencies to either side; in a few, they are chosen to impart special skirt characteristics. The shape factor of most full-lattice crystal filters is down to 2 or better.

The crystals in half-lattice filters, like those in fig. **6B,** are chosen so that the parallel-resonant frequency of Y2 is the same as the series-resonant frequency of Y1. The result is the bandpass curve needed for ssb operation. Occasionally, a capacitor is added across one crystal to warp its frequency for exactly the spacing needed to produce the correct bandwidth. Half-lattice crystal filters exhibit shape factors around 2 or a little higher—adequate, but not so sharp as full-lattice filters.

The differences in coupling in the two half-lattice filters in fig. **6B** deserve brief comment. The half-lattice configuration depends on the crystals being fed in push-pull, with the output taken in parallel. The transformer-coupled circuit does this easily; with a grounded center-tap, the secondary winding feeds the crystals in push-pull. To accomplish the same thing with the single-ended output from a tube is not that easy. With C2 and C3 across the plate coil, and a ground between them, the crystals are effectively fed in push-pull, after all. The rf choke keeps the bottom end of C3 from being grounded through the power-supply capacitors. C1 is merely a dc-blocking capacitor to protect the crystals.

fig. 7. Variants on half-lattice configuration. With a shunt crystal (A), back-to-back for single-ended input (B).



There are several variations on the half-lattice configurations; two of them are shown in fig. 7. Another variation, not shown, is a cascaded series of half-lattice filters. This is seldom used in commercial filters because of the extra coupling transformers needed.

In fig. **7A,** Y1 and Y3 are chosen for the same frequency, the center of the i-f pass

*** McCoy Electronics Company, Mount Holly Springs, Pennsylvania**

band; Y1 is series-resonant, though, while Y3 is parallel-resonant. Y2 is series-resonant at a frequency offset by the bandwidth desired. In one design, for example, Y1 and Y3 are picked for 455 kHz, and Y2 for 453.2; bandwidth is 1.8 kHz. Shape factor is 2.3, but one skirt is much steeper than the other. That's the side the signal is placed on.

The hack-to-back configuration in fig. 7B exhibits a better shape factor than two cascaded half-lattice stages merely cascaded. This is also an excellent way to feed the signal from a single-ended stage. Y1 and Y2 are chosen to match in frequency. Y3 and Y4 match each other, but are 1.5 to 1.7 kHz away from Y1 and Y2 in frequency. This arrangement is used in one commercial ssb filter in the 5-MHz range.

checking filters

There isn't space to go deeply into testing ssb filters. Briefly, though, there are three methods. One is to use an oscilloscope, sweep generator, and marker generator to display an exact curve and examine its bandwidth, shape factor, etc.[1] Another is to use a signal generator and a VTVM with an rf probe; you can plot a response curve in just a few minutes, one point at a time. The third is to test the filter's operation with signals; if the receiver or transmitter cannot be aligned to receive or generate an ssb signal with proper modulation, suspect the filter.

In most cases, you'll be able to do nothing about filter trouble, beyond buying a new one. The sealed black-box nature of ssb filters precludes any repairs. At least, however, you know what should be going on in them.

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ham radio

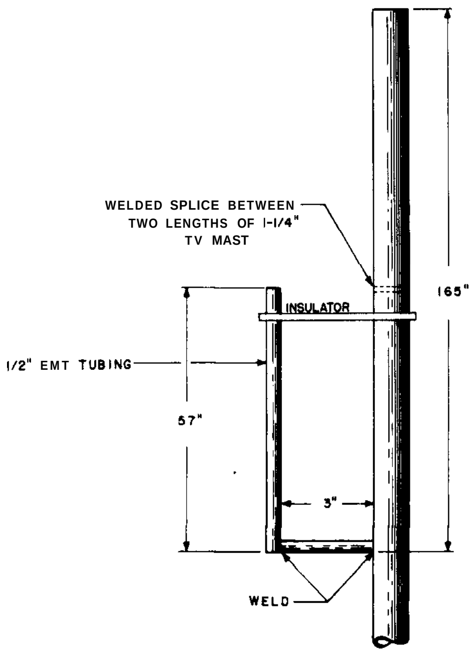
j-pole on six

If you have limited space and are looking for a simple antenna, why not try the J-pole?

Harold E. L rson, K4SDY, 8 -23rd Ave ue North, S Petersburg, Florid a33

As all amateurs know, a good antenna is the heart of the ham shack. You just can't communicate without a good radiator. Some hams buy or build very elaborate systems, while others look for the easy way out. Since

fig. 1. The J-pole antenna for six meters.



I am somewhat confined by power lines, my property line and finances, I looked for an easy way.

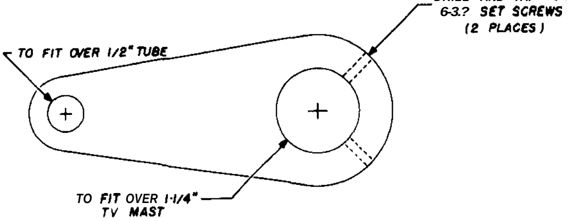
When I wanted to check out my recently completed Knight TR-106 Transceiver, I used a piece of number-8 wire as an antenna. It worked out far enough, 3 miles, to contact Art Bond, K4HQA. Art is "Mr. Six Meters" of St. Petersburg. When he found out what I was using for an antenna, he suggested the J-pole to me. Since completing it, I've worked ground-wave stations up to 40 miles away with no strain. When the band is open, I can get in with the best of them with my 15 watts input.

The J-pole, although a rather simple antenna, has proven to be a very good performer. It doesn't take up much space and it's pretty inexpensive; all you need is two 10-foot sections of 1 114-inch TV mast, 5 feet of thinwall electrical EMT tubing, a small piece of lucite or plexiglass and two strap-type clamps.

The one prime requisite is that the completed J-pole must be one continuous piece of metal. Weld or braze all the joints. The radiator is 165 inches long and is spaced 3 inches from the matching stub as shown in **fig. 1.** The matching stub is 57 inches long. The 1/2-inch tube between the radiator and the stub should be cut so that it mates with the circumference of the adjoining piece. This makes welding easier. Plexiglass is used as a spacer for the top of the matching stub; it should be at least 112-inch thick to accept two set screws. The details are shown in fig. **2.**

The electrical connections to the J-pole are made with pipe straps or water-hose

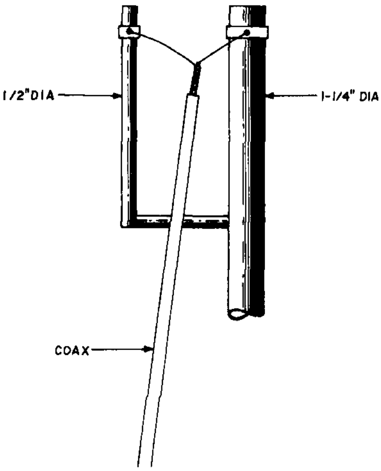
fig. 2. Layout of the insulator.



clamps for ease of connection and adjustment. To provide a match to 72-ohm coax, the connections are about 6 inches up from the horizontal piece; about 4.5 inches up for 52-ohm line. If you want best efficiency, use an SWR bridge to make feedline adjustments. The center conductor is connected to the 1 114-inch mast, and the outside braid is connected to the 112-inch cross piece as shown in **fig.** 3.

Remember to keep the connections as short as possible. The coaxial cable should be fastened to the mast with clamps or tape to

fig. 3. Connecting the feed-line to the J-pole antenna.



hold it in place. When matching the feed-line, the antenna should be at least ten feet above the ground.

Installation is easy; it can be attached to sections of TV mast or a tower. For best operation, the tower or mast should be grounded. In my installation, I drove an inch-and-a-quarter TV mast five feet into the ground. Then two sections of mast were slipped on this piece with the J-pole on top. The top of the antenna is approximately 40 feet off the ground. It is attached to the roof line of the house and guyed once at the 25-foot level. In addition to being a very good antenna—both locally and on skip—the whole assembly is at ground potential, one big lightning rod. In Florida, as in many other areas, this is a definite advantage for safeguarding your home and ham shack.

ham radio

tips for keeping your beam

Ed Marriner, W6BLZ, 528 Colima Street, La Jolla, California 92037

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 cracks. After awhile, the joints of the tubing develop insulating 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

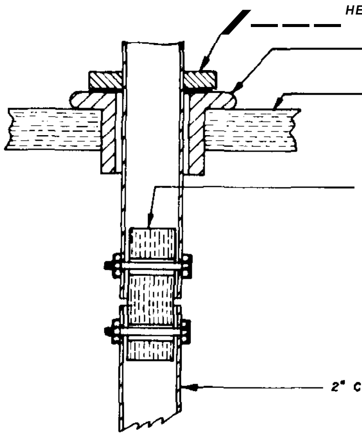
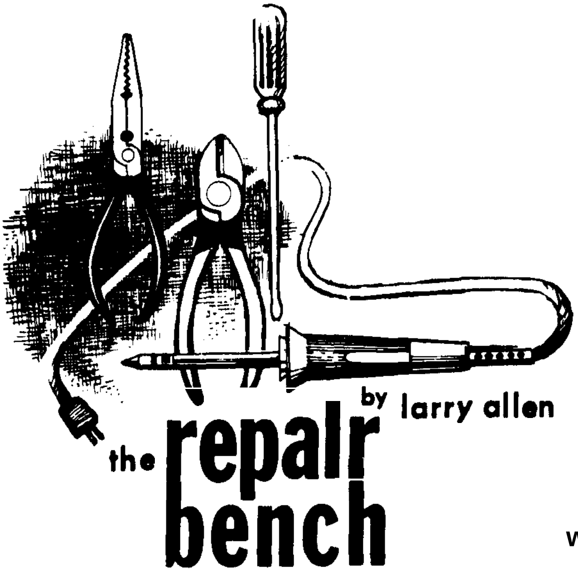


fig. 1. All open ends of the pipe and aluminum should be sealed with corks.

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).

ham radio



high-power troubleshooting –keeping alive

On nights when the bands are open in all the right directions, and you're hearing reports that you're getting into Deep Zamba-Zamba-Land with a five-nine signal, and the XYL comes by the shack with fresh coffee and tells you she thinks you **should** buy that new beam, life is really worth living. Why take a chance of spoiling it? Operating high-power has responsibilities as well as pleasures. The key responsibility is preventing it from doing any harm — to you, to your family, and to visitors.

If you housekeep the shack like you should, and designed and built it well to start with, the last two mentioned are well taken care of. Danger arises when you — lord and master of the domain — get careless during trouble-shooting. Let's face it. When you haul a 2-kW linear up on the bench and open it up, you're dipping into a powerful piece of machinery.

what the dangers are

When testing a high-power transmitter, or its power supply, you have three dangerous voltages to contend with. One is the primary supply. It may be "only" **115** volts of house ac, but it can be the most dangerous voltage in the equipment. Another is the high-voltage dc that supplies the plate of the power amp. Not only is it several thousand volts, but the supply has a powerful current capability— and it's current that does the killing. Third, and just as important, is the rf voltage developed in the final tank and along the antenna feed line. Rf is usually confined to a well shielded cage; but, remember, we're talking about when the unit is on the repair bench — you may take the lid off for some reason.

Each of these three dangerous voltages behaves differently. The primary power, usually 115 or 230 volts ac, has an awesome ability to push current through your body. Once contacted, it can contract your muscles so tightly you can't turn loose, haul your heart up motionless, and hold your lungs powerless to breathe. You can die of asphyxiation as easily as from heart fatigue.

An acquaintaince of mine once survived a tangle with a shorted electric drill on a TV tower. He was bound up so tightly by the current, he couldn't release the trigger; he couldn't even yell. He did manage to grunt, and someone on the ground yanked the plug. lust plain lucky, he was. At that, he blacked out and hung upside-down for a while on the tower, and spent some time in the hospital overcoming shock. Rest assured, you had better respect that common old "everyday" line voltage.

Just as deadly is the high-voltage dc inside the power supply or transmitter. A fellow I had only met once was killed by a 3,000-volt plate supply. He was troubleshooting the transmitter modulator. The trouble was an overload, and he had to cheat the interlocks and hold in the overload relay. He took all the normal precautions for working on live gear; he was a trained broadcast engineer. But, he lost his balance. One hand hit the transmitter cabinet and the elbow of the other arm landed on the 3,000-volt terminal of a coupling transformer. The horrible jolt to his body, heart, and lungs was too much. Revival was impossible.

I'm not trying to scare you, although if that's what it takes to teach you respect for transmitter voltages, then you'd better be scared. Safety in high-power equipment is a serious subject—you bet your life.

The rf voltage in a high-power transmitter has characteristics very different from line ac or power-supply dc. You don't even have to touch the wire or terminal carrying high-energy rf; it can jump out and burn you badly if you even get close.

If you're in good health, the rf voltage from a ham transmitter might do no more than give you one of the nastiest burns you can imagine. But high-power rf has been known to kill, and you shouldn't take the chance. Stay away from it. That isn't always easy, for reasons I'll explain, but you should learn how.

general precautions

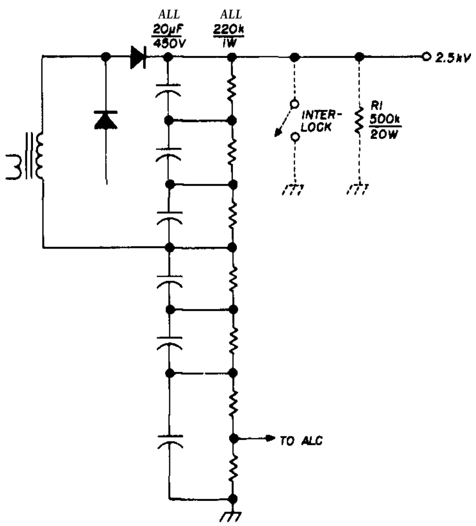
The most obvious step to protect yourself from high voltages is to keep the equipment **off** while it's on the bench.

"That sounds nuts," you say. "How can you troubleshoot it?"

Well, in most cases you can. Practice using resistance measurements to guide you to the trouble. You can do that with the unit turned off. There aren't really so many parts and dc paths in the average high-power transmitter or linear. It might turn out to be faster than going to all the trouble of wiring up the high-power unit for operation on your bench.

If you're convinced you have to fire up the equipment on the bench, first make sure there are no bare wires trailing anywhere. Use exactly the neat installation-type wiring you'd use if you were installing it permanently. The transmitting antenna lead must not be exposed. To "haywire" a test setup is courting danger. The primary power wiring must be through a cable and plug just the same as in a permanent installation; alligator-clipped connections won't do. Be especially careful of the wire carrying the high voltage from the power supply to the transmitter. It **must** be one solid piece, and should be the highest-quality ignition wire you can find.

fig. 1. Protective bleeder resistors in a typical high-voltage power supply. R1 is an added safeguard, and tho switch is an Interlock feature to discharge the power supply if the cage is opened for service.



Try to do your troubleshooting without cheating the interlocks. They're included to automatically disconnect primary power from the high-voltage supply when the lid or door is opened. Sure, you may have to open up the unit to reach test points. But it's dangerous to cheat interlocks. Again, the old resistance-measuring technique may save your life. When you have the interlocks cheated, and power is applied to all circuits, the open equipment is at its most dangerous. The primary ac, the dc high voltage, and the powerful rf energy are all right there waiting to zap you the instant you get careless.

Even with the set turned off, you're still not safe. The better equipment has built-in safeguards against some of the voltage hazards that lurk in a dead transmitter, but they can't always be trusted. (Suspicion is a useful companion where high voltages are concerned.)

For example, filter capacitors in the high-voltage section can hold a body-jolting charge for days and weeks after a set is turned off. To prevent this, bleeder resistors are almost always included. Fig. 1 shows a high-voltage doubler circuit, with bleeders. What happens if one of them opens? The charge won't drain off. Safer transmitters include another resistor as bleeder for the whole supply (R1 in fig. 1). The safest units further include a switch that shorts out the high-voltage dc output whenever the protective shields are removed or the housing lid is opened. This is in case the primary interlocks are cheated, purposely or accidentally. The discharge switch also protects in case bleeders aren't doing their job.

Still, the watchword is: don't trust **them.** All these devices are fine, unless for some unusual reason they aren't working. It only takes once to ruin you. So . . . use a grounding stick, ALWAYS. This is something you can make. A long bakelite or fiberglass rod is best **(fig. 2).** Screw a small hook—stainless steel or aluminum—into one end. Attach a 2-foot piece of wire to the hook, making very sure of good electrical connection. To the other end of the wire, which can be heavily-insulated ignition wire, solder a good, heavy-duty clip.

When you open up a transmitter or power supply, clip the grounding wire to a bare chassis spot (paint insulates). Hold the rod at its free end. Then touch the hook to all bare wires or terminals that you have any slight suspicion may have carried dc voltage. Finally, hang the hook over the main high-voltage terminal—usually at the power-supply output. Leave it there until you are finished troubleshooting. That gives you protection just in case you or someone else accidentally breaks an interlock.

Another precaution, often overlooked on the repair bench, is grounding. A ground from the chassis to a cold-water pipe or to an 8-foot copper ground rod is as important on the bench (maybe more so) as at the operating console. Without it, the chassis may become electrically hot with respect to ground, and offer a dangerous situation. The chassis might also, when you fire up the transmitter, take on some of the rf energy; that, too, is dangerous. Make a point of grounding the chassis carefully.

the live ones

Sooner or later you're going to insist on troubleshooting a transmitter live. It may even be necessary. You'll be tampering with dangerous stuff, though, and should act accordingly. There are some protectice measures you can take.

First of all, READ THE INSTRUCTION BOOKLET. If you won't or can't, you have no business working on dangerous high-powered equipment. The manufacturer of your transmitter or power supply is well aware of the danger points, and will likely have printed cautions in the manual. Study them. Compare the schematic diagram with what you see in the chassis. If you have any doubts about where a test point is, trace it down with the unit turned off and with all the earlier precautions. Then you won't have to probe around later hunting for it. A live transmitter is no place to "learn" where the test points are.

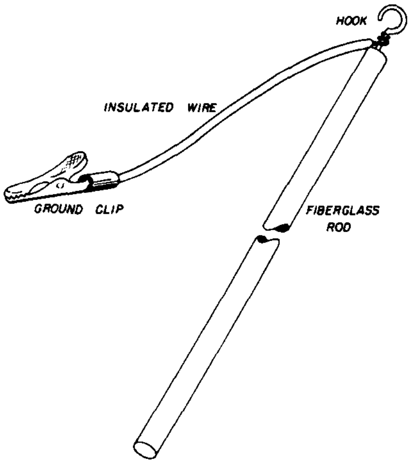
Next, rehearse each test you're going to make. Sound silly? Nothing's silly that might save your neck. A written list of what you're going to do and what you expect to find will prevent that "hunting around" that breeds carelessness. A dress-rehearsal run-through, with the set turned off, will give you the confidence to go straight ahead with your plan when the power is on. You can't afford many mistakes—maybe none.

Once you start, the old ploy about keeping one hand in your pocket is still a good idea. It's based on the theory that you won't complete a circuit to the chassis and therefore won't get bitten by voltage. That theory

doesn't always work, because you can contact two points of voltage with one arm or hand. However, it can help keep the current from passing through your rib cage, and might very well save your life even if you get a nasty jolt.

Standing on a rubber mat is another good idea, especially if your floor is concrete. Again, it may not prevent your getting shocked, but it could keep the effects to a minimum.

fig. 2. Grounding stick for discharging high-voltage circuits. Heavy insulation on wire keeps it from contacting any circuit but the intended one.



In tuning a high-power transmitter, you may have to change taps on a coil. To make each change, shut down the transmitter, hook the grounding stick to the coil (other end of the wire clipped to chassis, of course); make the change; remove the stick; then fire up the transmitter again. This takes a little extra time, but not enough to matter when your health is at stake.

When messing around with tuning, keep away from that rf. It can reach out in some of the most unexpected ways. I saw a guy "testing for rf" (he said) by bringing his screwdriver blade near different parts of the plate coil. He judged by the little rf arc he could draw. He laid the screwdriver out of reach, so he tried it with a pencil. The burns he got on three fingers and a thumb took three weeks to heal. He forgot the pencil-lead ran all the way up through the pencil. Good thing it was only a 250-watt transmitter. The only safe place for your hands in regard to a radiating transmitter is away from anything that even looks like a coil.

You have to watch how you handle tools, too. Not only can you damage your transmitter, you can damage yourself. Metal tools in a powerful rf field will overheat quickly, even though many inches away from direct contact. Keep them out of live transmitters. If you have an adjustment to make inside a radiating transmitter (not a very good design, and not likely in commercial units), use a plastic tool; plenty are available.

the bench itself

This is another factor worth consideration if you expect to work on a high-power transmitter. The bench should be wood; metal just isn't a good idea. Receptacles to fit various power plugs are important; you don't want to get tangled up in any jury-rigged power connections.

The all-important ground connection mentioned earlier is most handy if it terminates right there at the bench; that makes it quick to connect to. The wire lead to the pipe or the ground rod should be as short as possible and of the heaviest wire you can round up.

A well-protected antenna connection is a good idea, if you can't manage a dummy load. Coaxial cable is always best, but not feasible at all frequencies or with all transmitters. Just make the terminations at the rear of the bench space, so you don't accidentally touch them while the transmitter is radiating.

Finally, the bench should have an accessible switchbox that shuts off everything. You never know just when you might want to close down the whole operation suddenly, or need someone else to. Make it easy. One precaution: whoever can turn it off for you could also turn it on when you least expected or needed it. Add extra safety by having a hidden switch in series with the main one; when you want the bench to stay off, open them both.

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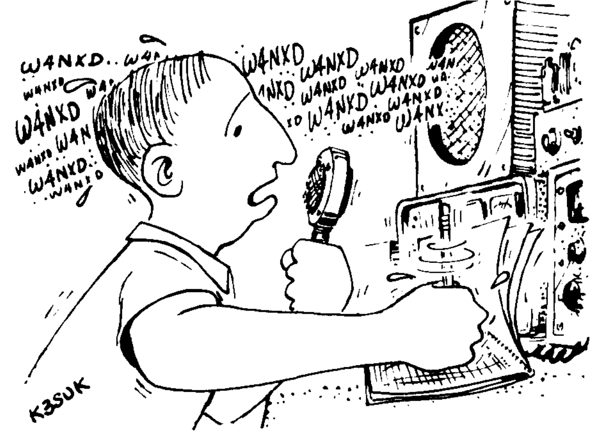

how to be DX

J. Michael Blasi, W4NXD, 1490 Enota Avenue, N. W., Gainesville, Georgia 30501

Several weeks ago I decided to see if my transceiver still worked on ssb since I'm usually on CW. I cranked her up on the high end of 20 meters, made sure the final tubes weren't turning pink and gave a listen.

The high end of 20 meters, in the phone band, is a bit different from the lower portion. The stations are only two deep and every *2* kHz there's a net that always tells me they are running emergency traffic from a boy scout on the Island of Gamua to his 112-year old grandmother who thinks he went to the store for bread. This may be a bit exaggerated, but I think you get the idea of the general conditions. Generally speaking, however, I feel these people do a lot of good.

I found 14.336 kHz clear, so I asked if the frequency was being used. It was. A tremendous signal came on, told me he was W8UMR and that he was net control for the Independent County Hunters Net—did I want to check in7

Since I'm a cautious person by nature, I inquired what might be the purpose of this organization. I was told they tried to get rare counties on the air so those hams looking for awards could work them. This seemed reasonable to me, and I didn't think I'd get too much traffic for grandmothers. I gave my call, said that I was in Hall County, Georgia and sat back to drink a cup of coffee while I listened. This was not to be—about twenty stations wanted to know who was in Hall County.

This really threw me; I didn't know I was in a rare county, but since most of the local hams operate two-meter fm, I guess I am probably the only one who operates on 20 meters. The net control asked me to QRZ the frequency to see who needed Hall County, I did. The result was unbelievable; everybody was calling "W4NXD." Compared to this, Don Miller seemed to be as popular as a W8 calling CQ DX on 20 CW.

Using my lightning reflexes and years of hot-shot operating, I panicked. There must have been fifty stations on the frequency calling me; well, at least 30; would you believe 107 Seriously, it made Field Day seem like ten meters.

As I said before, my years of operating came through as I pushed the sweaty mike button and said, "QRZ W1 onlyl" There, right in my ears were about a dozen WI's calling me. I sifted them out, wrote the info in my log like a jack rabbit and called QRZ for W2's only.

My only trouble came when I got to the W9's (they always do it). There were so many I actually had to say, "QRZ W9's in Wisconsin only." In about 40 minutes I worked close to one hundred stations. Yes, little me in Hall County had been a rare one. I say had, because as of now, Hall County is no longer rare. I took care of that. All total, I worked many more stations over the next few days, but nothing like the first time I checked into that net.

So, if you'd like a quick dance in the spotlight, just check in and see if you're a rare one. Enjoy it while they need you, because nothing is sadder than a once rare county.

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amateur radio in space -- a complete bibliography

Nick Marshall, W6OLO, NASTAR, Post Office Box T, Syosset, L. I., New York 11791

With the rapid advances which are being made in electronics technology, and the great interest in amateur space experiments such as OSCAR and MOONRAY, a great number of space-oriented amateur radio articles have appeared in the past few years. The following bibliography is a result of a literature search that was conducted by NASTAR—Nassau College Amateur Satellite Tracking Society. NASTAR is an independent, non-profit amateur group which is currently working on MOONRAY, an amateur-band lunar translator.

Since the first OSCAR was put into orbit around the earth in 1961, there has been continual progress: OSCAR II in 1962 and OSCAR's III and IV in 1965. OSCAR V, built by a group of Australian amateurs, is due for launch in late 1968. MOONRAY, a five-pound translator operating in the 420-MHz band, is in the proposal stage. It is proposed that this package, with an isotope power supply, will accompany the third manned lunar landing.

The following bibliography is a collection of articles covering the amateur radio space effort. It has been made as complete as possible, and covers the major U.S. amateur radio magazines as well as other publications. The bibliography is broken down into four separate lists. One for each of the magazines QST, CQ, and 73, and a miscellaneous category for articles appearing in other magazines. The articles in each group are listed chronologically with the most recent ones first.

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ham radio

grid-current measurement in grounded-grid amplifiers

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 9407

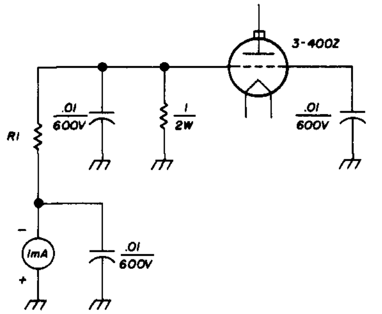
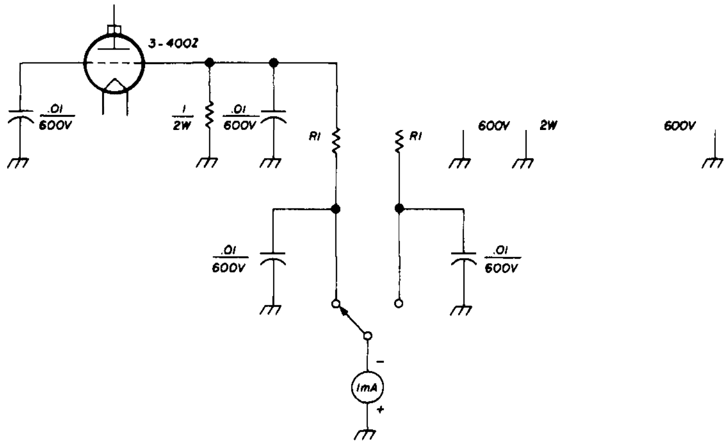


fig. 1. Measuring the grid currant of a single grounded-grid stage.

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. **1.** The control grid is grounded through a l-ohm composition resistor that is bypassed with a .01-µF 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. 2.** The maximum current which can be measured is determined by the internal resistance of the milliammeter

fig. 2. Measuring the grid current of a pair of groundad-grid amplifier tubes with one meter.



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 Triplett 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 the milliampere 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 milliampere 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 l-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 vbltage drop across the l-ohm resistor and larger values of series-multiplier resistance are needed.

ham radio

code practice - - the rf way

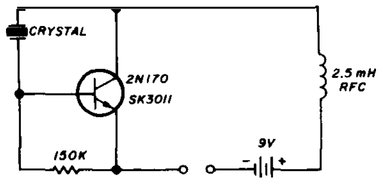
There are many transistorized audio oscillators available for code practice, but almost all of them fail to give a realistic "on-the-air" quality to the signals. The rf oscillator described here provides a realistic signal and is copied through the receiver just as any CW signal. It is tunable of course, and if the receiver has a bfo, the pitch can be varied as well. Furthermore, construction is even simpler than the audio type code-practice oscillator—so much so that a printed-circuit board would accomplish nothing but to complicate construction!

the circuit

The circuit is a simplified crystal-controlled Pierce oscillator. Since it's designed for one of the lower frequencies, the inexpensive 2N170 works very well. If you want to go to higher frequencies, other transistors can be chosen for operation on these bands. PNP types can be used by merely reversing the battery polarity.

construction

First, get the parts together—you should find most of them in your junk box. Then cut out the 2x3-inch perforated board and drill the mounting holes for the crystal socket at one end. Mount the transistor, rf choke and 150k-ohm resistor on top of the board and push the leads through the holes to the bottom of the board. Attach two soldering lugs for key connections and the battery clip. Then, solder all the necessary connections; use a heat sink on the transistor leads. If you make the physical layout like the schematic, most of the connections can be made with the component leads.



operation

Attach your key to the soldering lugs, put a crystal in the socket, hook up the battery and tune in the signal on your receiver. With a little practice, you can tell just how close the little oscillator has to be to the antenna lead to provide adequate signal strength in the receiver

George Haymans, WA4NED

two-band novice superhet

An easy-to-build superhet receiver for 80 and 40 that provides high performance at low cost

Darrell Thorpe, 3110 N. 83rd Street, Scottsdale, Arizona

For the prospective novice, a first receiver can be a rather difficult choice, especially when he looks at the large price tags. He really doesn't have much choice—it's either "shell out" for the fancy factory-wired job or resort to a simple regenerative receiver. However, the red-hot superhet described here can open up a whole new receiver era for the beginning ham. This simple, easily built superhet can be built at a price that compares to the simple regenerative receiver, yet it gives big-set performance.

Total cost is only $73 if you use all new parts. Sensitivity is around 1 microvolt; this is high performance when compared to the usual 10-μV sensitivity of low-cost ready-made receivers. The outstanding performance of this receiver is achieved by using the latest solid-state devices and toroid cores. In addition, this top performer covers both the 80- and 40-meter bands without a lot of complex coil winding or handswitching.

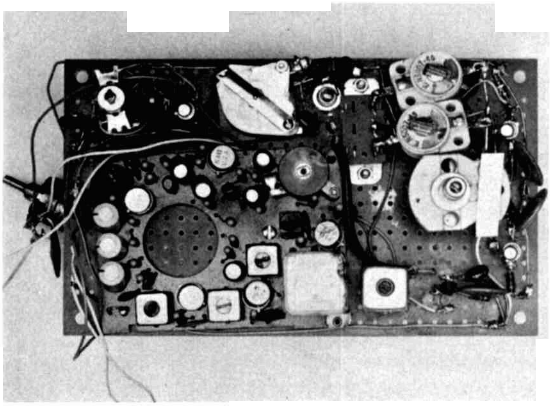
the circuit

The heart of the two-band novice receiver is an eight-transistor broadcast-band receiver that sells at discount stores in most cities for under $5.00. This receiver, with only a couple of minor changes, provides a three-stage i-f amplifier strip and detector and supplies plenty of audio to drive a built-in or external speaker.

This is not only a compact ready-made i-f and audio package—you could barely buy the eight transistors for the price of the entire radio! Mentally add up the cost of the i-f transformers, speaker, capacitors and resistors plus the transistors, and you'll begin to see what a real bargain this ready-built module is.

Next comes a late innovation from the semiconductor industry—a field-effect transistor (FET) in a mixer circuit. The FET is used for its superior mixer performance. Similarly, toroid cores are used because they provide superior coils. In the oscillator, the toroid coil, which you'll find is about the easiest coil you have ever wound, provides very good stability, and a two-stage oscillator circuit maintains it. At the antenna, the toroid coil provides high-Q for better selectivity. Another advantage of the toroid is the small amount of space required as compared to the usual 80-meter tuned circuits

Construction of tha two-band novice superhet. The transistor broadcast radio and volume control are to the left; the oscillator is in the upper right and the FET mixer, in the lower right.



construction

Start hy removing the case from the transistor radio. Look for two leads that go from the circuit board to the earphone jack and then to the speaker.

1. Disconnect the leads at the phone jack.

2. Remove the speaker and phone jack from the radio and mount them in the new cabinet (Radio Shack Perfhox 270-097).[1]

A mounting hole for the phone jack is already provided in the cabinet. Drill three holes for the speaker-mounting screws. Note that the nuts grip the edge of the speaker frame.

3. Disconnect and remove the antenna loopstick after noting where the leads of the small winding are connected. One lead is connected to the base of the converter transistor and the other to the bias resistors for this transistor (see fig. 1). Solder a 2-inch length of insulated wire to each of these points; they will be connected to a new i-f transformer (T1) later.

4. Note that the collector lead from this same transistor goes to one winding on the oscillator coil. Use a short piece of insulated wire to short out this winding. This kills the oscillator and converter stage in the broadcast set and becomes an additional i-f amplifier. This additional stage contributes to the excellent sensitivity and selectivity of the receiver.

5. Clip or unsolder the wires going to the volume control. It's not necessary to remove the control, but make sure the "on-off" switch is always "on"; the power switch for the novice superhet is on the new volume control. Connect three 8-inch pieces of wire to the circuit hoard where the three leads from the volume control were connected. Be sure to note the center-tap connection. Solder the other ends of these wires to the new 5000-ohm volume control.

6. Clip the battery leads in half. These leads

1. **Radio Shack, 730 Commonwealth Avenue, Boston, Massachusetts 02215.**

will be connected to the terminals on the perforated board.

These are the only changes to the broadcast radio, so it can be mounted on the perforated board. Flea clips are mounted in the perforated board near the radio for audio and power connections.

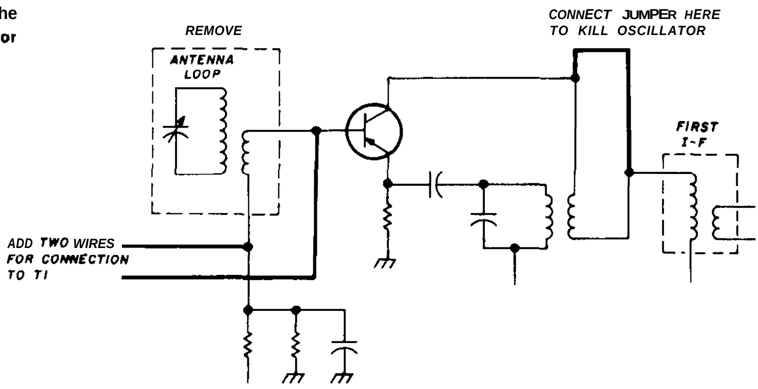
It is a good idea at this time to drill the remaining holes in the cabinet that are needed for the vernier dial, the volume control and bfo pitch control. The holes for the dial and the bfo pitch control are rather critical since they dictate the position of the tuning capacitor and bfo oscillator coil. In addition, drill a small hole in the bottom of the cabinet below the dial so you can tighten the setscrew in the dial.

needed of the 455-kHz i-f. The new i-f transformer you added (T1) may need a little peaking; however, new transformers are very close to 455 kHz.

Start with the 80-meter band first. It's the easiest to adjust because you should be somewhere within the band regardless of where the tuning and trimmer capacitors are set. Set the main tuning capacitor to minimum capacitance, the antenna trimmer capacitor at about 3/4 of maximum capacitance and the bandswitch to the 80-meter position. Adjust the 80-meter trimmer until you hear phone signals coming through.

These phone stations will probably be on ssb, so they'll sound like a bunch of quacking ducks. As you tune the main tuning capacitor towards maximum capacitance, you will tune down through the novice and general-class CW bands to 3.5 MHz. The 80-meter CW bands will be spread over the entire range of the capacitor for easy tuning.

fig. 1. Modifications to the transistor broadcast radio for use in the novice superhet.



Position the oscillator tuning capacitor on the perforated board so it lines up with the hole for the dial. Enlarge two of the holes in the perforated board to accept the stator lugs extending from the tuning capacitor. Put the tuning capacitor on the board and push two flea clips through the board right next to the capacitor lugs and solder.

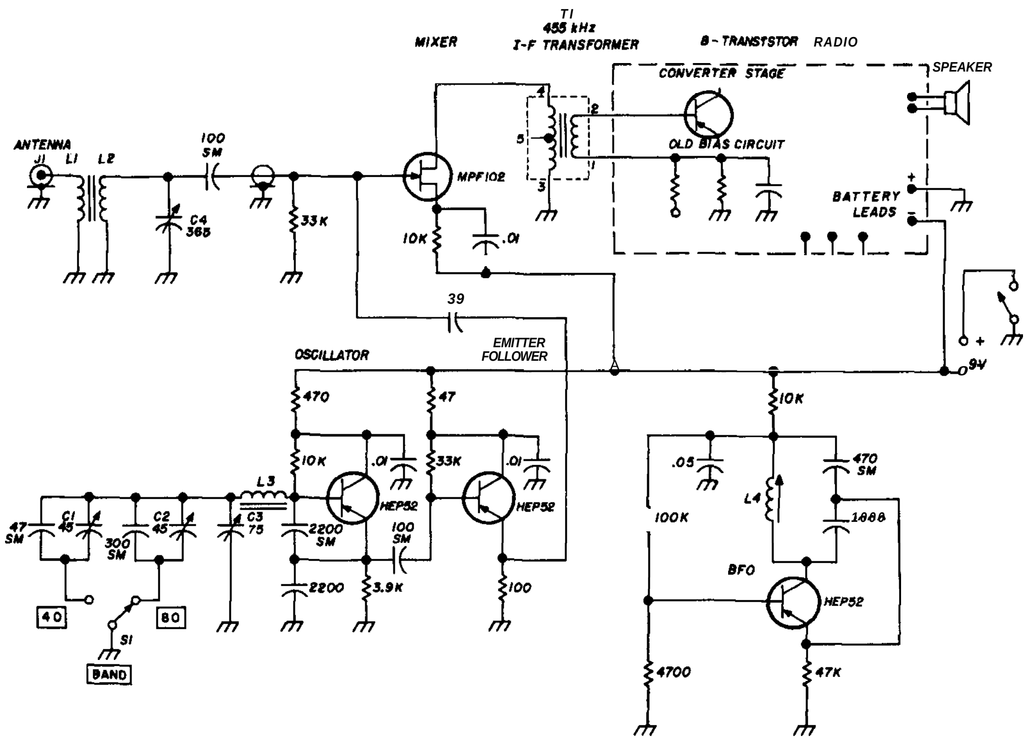
Proceed with the wiring of the oscillator, mixer and bfo as shown in **fig.** 2. The oscillator trimmer capacitors are spaced away from the board so they clear the oscillator coil. Notice that the bfo doesn't need any direct connection other than power.

alignment

Since the broadcast receiver is already aligned, there's practically no alignment

Since this receiver was designed primarily for novice use, the 75-pF tuning capacitor will tune the entire CW band and part of the phone band. If you want to cover the entire 3.5 to 4 MHz range, use a 100-pF tuning capacitor; however, the bandspread will be reduced. The only adjustment of the bfo is the pitch control; this is tuned about mid-range until a beat note is heard.

The 40-meter alignment is a little more tedious. Set the oscillator tuning capacitor to mid-range, the antenna capactor to about



C1, C2 7-45 **pF** trimmer **(Centralab 825BN)**
C3 **75-pF** variable **(Hammarlund** MAPC **75B)**
C4 **365-pF** variable (Radio Shack' 272 1341)
J1 Phono jack
L1 5 **turns** number 24 enameled **wire** wound between turns on around end of **L2**
L2 18 turns numbmr 24 enameled wire wound on ferrite core **(Amidon** Associates[a] **T44-15)**
L3 29 turns number 24 enameled wire wound an **ferrite** core **(Amidon Associates T50-2)**
L4 bfo coil (Antenna loopstick) **(Lafayette[3] 32H4106)**
R1 5000-ohm volume control with switch **(Lafayette 32H7363)**
S1 SPDT slide switch (Lafayette **34H3704)**

1 cabinet (Radio Shack **270-097)**
1 tuning dial (Lafayette **99H6030)**
2 small knobs **(Lafayette 32H2405)**

fig. 2 Schematic diagram of the novice superheterodyne receiver for 40 and 80 **meters.** The transistors are inexpensive Motorola types.

1/4 of full capacitance and the bandswitch to 40. Adjust the 40-meter trimmer to receive 40-meter phone stations between 7.2 and 7.3 MHz. The 40-meter position provides several megahertz of tuning range, so the 40-meter band is only a small portion of the dial. Therefore, there's not nearly as much bandspread on 40 as on 80, but even so, this receiver has proven quite useful on 40.

final assembly

Note the position of the vernier tuning dial when the setscrew is aligned with the hole in the bottom of the cabinet. Then adjust the capacitor for this setting. On mine, the dial read 70; therefore, I set the tuning capacitor to 25% of maximum capacitance. Then, when the assembly is installed in the cabinet and the setscrew is tightened, minimum capacitance will occur when the dial is advanced to 100.

Now, mount the volume control and bfo coil and install the completed assembly in the cabinet. Tighten the setscrew on the tuning dial and install the knobs.

2. Arnidon Associates, 12033 Otsego Street, North Hollywood, California 91607. $1.25 will cover cost of cores and postage.
3. Lafayette Radio Electronics, 111 Jericho Turnpike, Syosset, L. I., New York 11791.

about power supplies

A 9-volt power supply is recommended for this receiver. Don't use a supply that exceeds 10 volts or you may damage the transistors or electrolytic capacitors in the broadcast radio.

I have been using a battery pack with good results. For prolonged operation, six D-size flashlight batteries are recommended, preferably heavy-duty alkaline units. Alkaline batteries give up to ten times more service even under continuous operation. However, the standard D cells give highly satisfactory service. For short-term portable use, six AA batteries are okay, but alkaline cells are recommended for longer life. These battery packs can be charged many times and provide low-cost operation. Small 9-volt transistor-radio batteries will not hold up.

There have been a number of 9-volt power supplies described in different articles. Any of them will do the job. However, if you don't already have a power supply, the battery charger/battery pack may be the simplest scheme because large filter capacitors are not needed for the charger power supply.

antennas and operation

A good 80- or 40-meter antenna will give the best results, although highly satisfactory reception can be obtained with 30 to 40 feet of wire strung around the room. A good ground connection (water pipe, conduit, etc.) is also recommended, especially for battery operation. When the receiver is operated from an ac supply, sufficient grounding is obtained back through the power supply, and an external ground is not too important.

Operation is very simple since there are so few controls. Set the bandswitch to the desired band and peak up the antenna trimmer. Adjust the bfo pitch for a pleasing tone. You may find it necessary to repeak the antenna trimmer for weak signals as you go from one band edge to the other, but you don't have to peak it every time the dial is moved.

With a little practice, ssb voice signals can be tuned in. No switch is provided for the bfo because there are so few a-m phone stations. To listen to a-m, simply tune the bfo frequency out of range.

ham radio

propagation predictions for august

High frequency propagation forecasts for the month of August plus a discussion of scatter and short skip

Victor R. Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, California 94602

Propagation of high-frequency signals over long distances during August will likely be somewhat improved over July. Higher F-layer maximum usable frequencies (MUF's), lower absorption and lower noise levels may be expected as the sun slowly progresses southward.

Summertime conditions will still prevail, however, in the temperate north latitudes. The MUF's will slowly rise during the day after their initial jump at dawn. Peak MUF's will occur near sunset, and MUF's will decrease gradually during the evening to a minimum just before dawn.

Sporadic-E activity will decline during the month. The Perseids meteor shower will liven the month for the vhf operators, and some trans-equatorial forward scatter (TE) may reappear during disturbed conditions.

Propagation predictions consist of two parts: first, determination of the path MUF (is the MUF above the operating frequency?), and second, determination of the path LUHF (lowest useable high frequency) is the operating frequency above the LUHF?). Generally speaking, the path MUF is not dependent on system parameters, while the path LUHF is. The system parameters are transmitter power, antenna gain and noise levels. These predictions assume 100 watts CW output and unity signal-to-atmospheric noise ratio in a 6 kHz bandwidth. Antenna gains over an isotropic radiator are: −6 dB for 3.5 MHz, 0 dB for 7.0 MHz, 6 dB for 14.0

MHz, and 10 dB for 21.0 MHz.

the muf chart

The MUF data is presented as a time chart of median monthly 4000-km MUF values against local time and latitude at the control point. This is essentially the form used by the Central Radio Propagation Laboratory (CRPL)* of the National Bureau of Standards until 1963.

You may consider, as a starter, that this pattern is fixed in space relative to the sun, and the earth rotates underneath. Therefore, the time chart is the equivalent to a contour map of MUF vs latitude and relative longitude. However, there is a sizeable variation of MUF with longitude for the same local time, due to the offset of the magnetic pole from the geographic pole.

Originally, CRPL divided the world into three prediction zones (E, I and W) to take this difference into account. Present day ITS ionospheric predictions are a series of MUF contour maps presenting the median monthly MUF contours vs longitude and latitude for 12 hours of the day.

When the world-wide distribution of MUF is required, the ITS ionospheric predictions** should be consulted. However, I think the time chart presented here is adequate for American amateurs who want to predict the times of band openings beyond 2500 miles in various directions.

noise level

The noise levels are assumed to be set by atmospheric noise, values of which are published in charts[1] as a function of season, geographic location, frequency and time of day. While last month's forecast of maximum distance assumed a fixed noise level for each band, this month's forecast assumes an averaged and smoothed noise level which varies symmetrically with time of day. Noise level at 10 AM is assumed to be the same as that at 2 PM local **standard** time.

While there is some error in the smoothing process, it is insignificant compared to the variation of noise with location. If you are located near a thunderstorm area, your working range, particularly on 3.5 MHz, will be decreased. If you are located on the West Coast, your working range may be increased over that forecast.

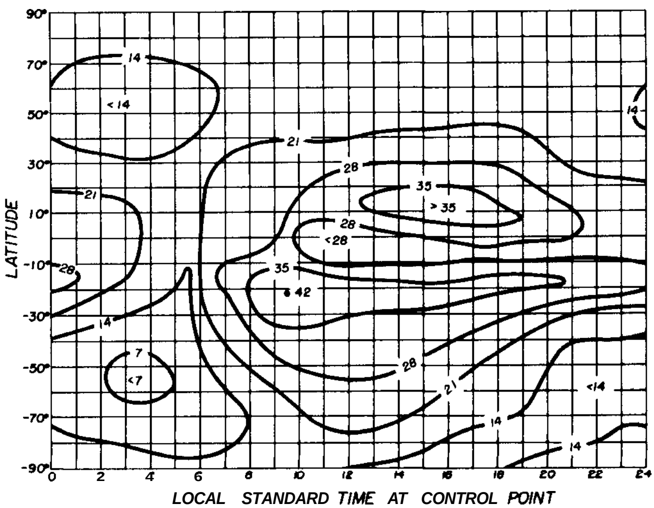
scatter propagation

While the MUF is not dependent on system parameters for regular on-path F-layer propagation, various scattering means will allow communication between well-equipped stations at frequencies **above** the

* Now the Institute for Telecommunication Sciences (ITS) of the Environmental Science Services Administration (ESSA).

** Available from the Superintendent of Documents, U. S. Government Printing Office, Washington, D. C., 20402, for $2.75 per year.

fig. 1. Maximum usable frequency curves for August 1968 based on 75° W longitude.



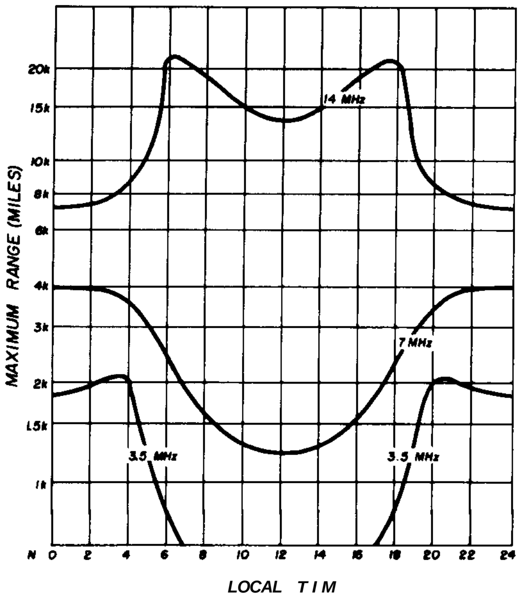


fig. 2. Maximum range to **the** north from **38°** N **lati-**tude due to absorption.

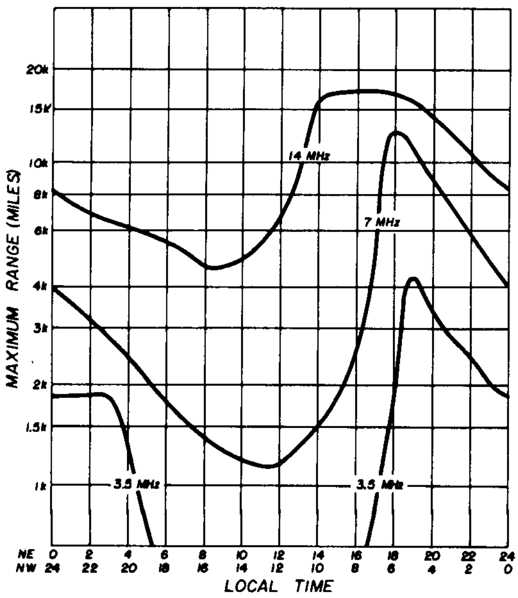


fig. **3.** Maximum **range** to the north-east (top time **scale)** end to the north-west (lower time scale) from 38° N latltuda due to absorption.

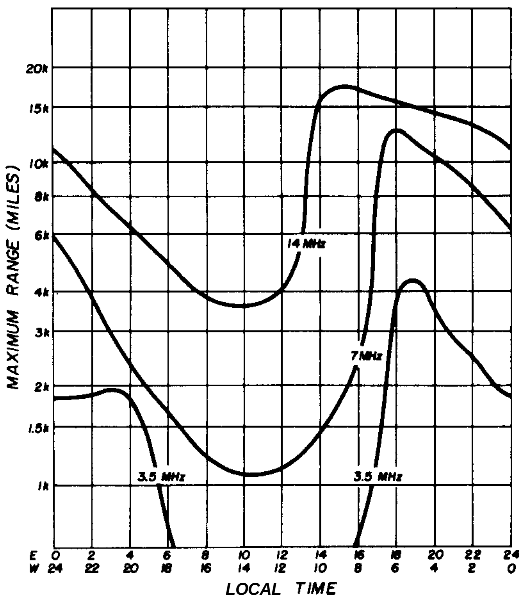


fig. 4. Maximum range to the east (top time scale) and to the wart (lower time scale) from **38°** N **lati-**tude due to absorption.

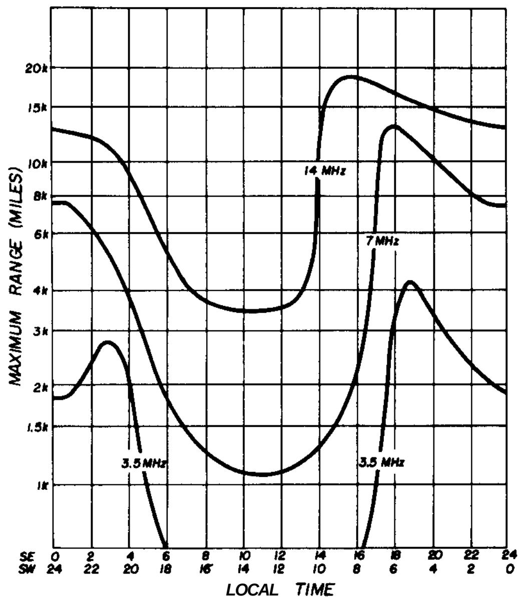


fig. 5. Maximum range to the south-east (top time scale) and to the south-west (lower time **scale)** from **38°** N latitude due to absorption.

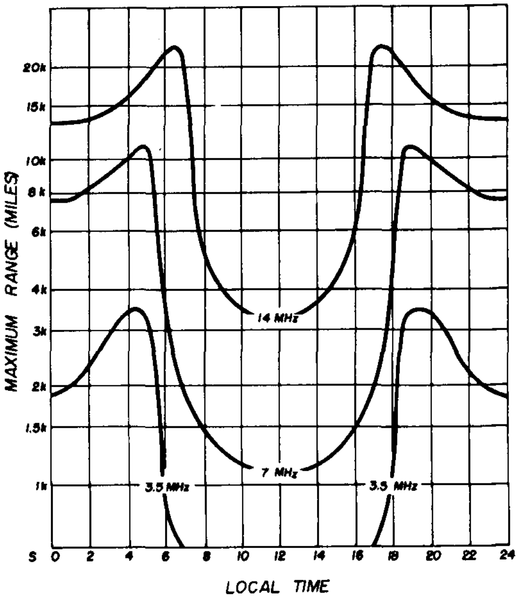


fig. 6. Maximum range to the south from 38° N latitude due to absorption.

path MUF between them.

One of the scattering means is side scatter. This is ground scatter from an area well to the side of the direct path, where the MUF's to both transmitter and receiver are both above the transmitter frequency. Using this mode, stations in Europe may communicate with stations in the States by way of ground reflections in Northwest Africa. Both stations simply point their beams at Northwest Africa.

direct path

The direct-path MUF for a path less than 2500 miles is lower than the MUF shown in fig. 1. **Fig.** 7 shows the ratio of the F2-layer ordinary-wave path MUF for path lengths between zero and 2500 miles to the MUF shown in **fig. 1.** The actual direct path MUF will be due to the extraordinary-wave whose MUF is between 0.1 and 1.4 MHz above the ordinary-wave MUF and depends on the orientation of the path to the earth's magnetic field and the path length.2

More on ordinary and extraordinary waves in a later column.

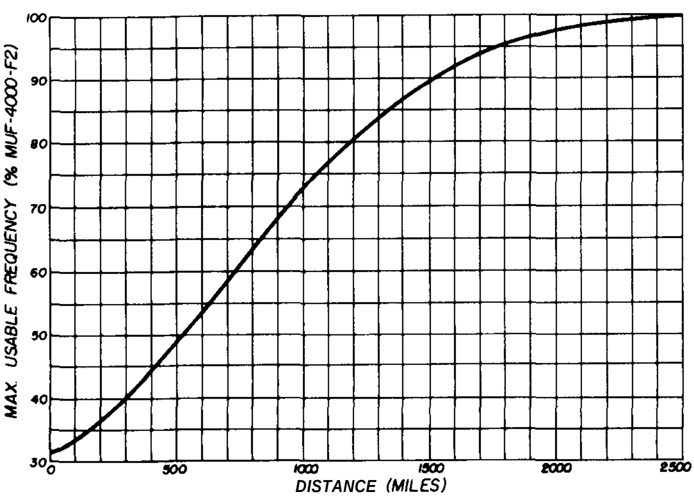
F2-layer MUF for paths under 600 miles in length will seldom be above 28 MHz, but stations this close may be worked by a variety of scattering modes on 28 MHz. The strongest mode is usually ground back-scatter from a commonly-illuminated area 1200 to 2500 miles away.

Ionospheric forward scatter and meteor scatter will furnish some communication out to distances of 1200 miles for well-equipped stations. Round-the-world propagation will furnish some signal inside the normal skip zone.

propagation summary for august

80 and 40 *meters.* Summertime noise levels and absorption will limit DX possibilities. European and Asian broadcast stations will become more troublesome because of decreased absorption over the pole as the equinox approaches.

fig. 7. Maximum usable frequency for distances under 2500 miles as a percentage of tho MUF plotted in fig. 1 on page 73.



20 meters. Twenty will be best for evening and early morning hours. The MUF will remain above 14 MHz in the Northern Hemisphere except for the predawn minimum period. As a consequence, non-amateur operations from various parts of Northern Asia will continue to clutter the band during evening hours.

15 meters. Fifteen will be the band for DX during daylight and early evening hours although transpolar openings will be spotty.

10 meters. Ten may appear to have died with the decrease of sporadic-E and sporadic-E assisted propagation, but openings should continue to the Southern Hemisphere.

6 meters. There are a few good sporadic-E openings left in August. During the Perseids meteor showers, widespread meteor-induced sporadic-E is expected. After the Perseids, some TE and even Aurora may occur during disturbed conditions.

2 meters. August is a good month for tropospheric openings but is more noted for the Perseids meteor shower. Most useful schedules will be between midnight and 11 AM local standard time at the path midpoint. Distances of 600 to 1200 miles are easily covered by well-equipped stations using schedules and fast break-in. Distances of 1200 to 1500 miles are more difficult to cover, but the more adventuresome may be scheduling out to 1500 miles.

The more adventuresome may also be trying 220 MHz for meteor scatter. What is really needed, in my opinion, is an amateur "Janet" type system with high-speed teletype, interrogation and storage for making the maximum utilization of each burst.

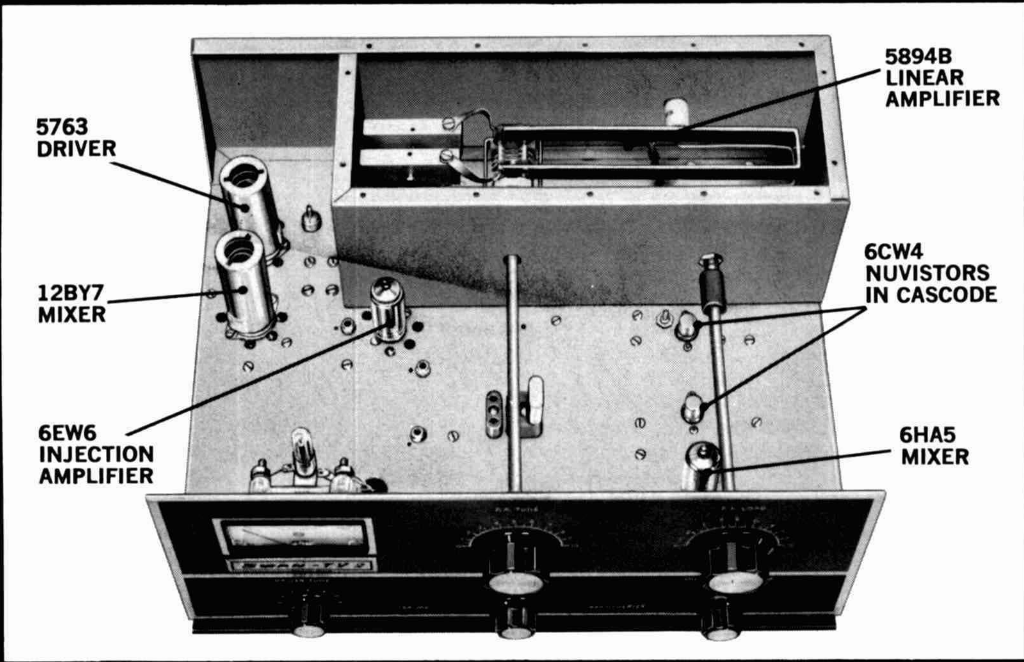
references

1. CCIR Report 322 (1964), World *Distribution* and *Characteristics* of *Atmospheric* Noise.
2. K. Davies, Ionospheric Radio Propagation, National Bureau of Standards Monograph 80, U. 5. Government Printing Office, Washington, D. C., 1965, p. 181.

ham radio

2 METER SINGLE SIDEBAND

144-148 mc 240 WATTS P.E.P. INPUT



THE NEW SWAN TV-2 TRANSVERTER

A receiving and transmitting converter for the 2 meter band, designed to operate with Swan Transceivers, models 250, 350, 350-C, 400, 500, and 500C.

SPECIFICATIONS:

14 mc intermediate frequency is standard. Thus, when operating the Transceiver from 14 to 14.5 mc, the Transverter functions from 144 to 144.5 mc. Additional crystals may be purchased and switched in for other portions of the 2 meter band, such as 144.5-145, and 145 to 145.5 mc. Three crystal positions are available.

Alternately, the TV-2 Transverter may be ordered for an I.F. in the 21, 28 or 50 mc bands, if desired. Of course, for use with a Swan 250 six meter transceiver, the Transverter must be ordered for 50 mc. Otherwise, the standard 14 mc I.F. is recommended since bandspread and frequency readout will then be optimum. The Transverter can easily be adjusted in the field for a different I.F. range, if required.

A 5894 B Power Amplifier provides a PEP input rating of 240 watts with voice modulation. CW input rating is 180 watts, and AM input is 75 watts.

Receiver noise figure is better than 3 db, provided by a pair of 6CW4 nuvistors in cascode.

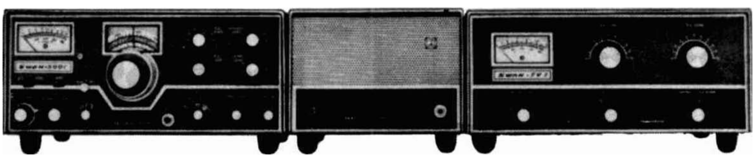
Only a Swan Transceiver and Swan AC power supply, Model 117-XC, are required. The power supply plugs into the Transverter, and the Transverter in turn plugs into the Transceiver. Internal connections automatically reduce the power input to the Transceiver to the required level.

Tube complement: 58948 Pwr. Amp., 5763 Driver, 12BY7 Transmit Mixer, 2N706 crystal osc., 6EW6 Injection Amp., 6CW4 1st rec. amp., 6CW4 2nd rec. amp. in cascode, 6HA5 rec. mixer.

The Swan TV-2 may also be operated with other transceivers when proper interconnections and voltages are provided. A separate Swan 117-XC power supply will most likely be required.

Dimensions: 13 in. wide, 5½ in. high, by 11 in. deep. Weight: 13 lbs.

$265



MODEL 250 ...... $325
MODEL 350C 420
MODEL 500C 520

MODEL 117-XC
AC POWER SUPPLY .. $105

MODEL TV-2
144 mc TRANSVERTER

SWAN
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the ham notebook

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 maintainance 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 mis-alignment 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 reduces 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.

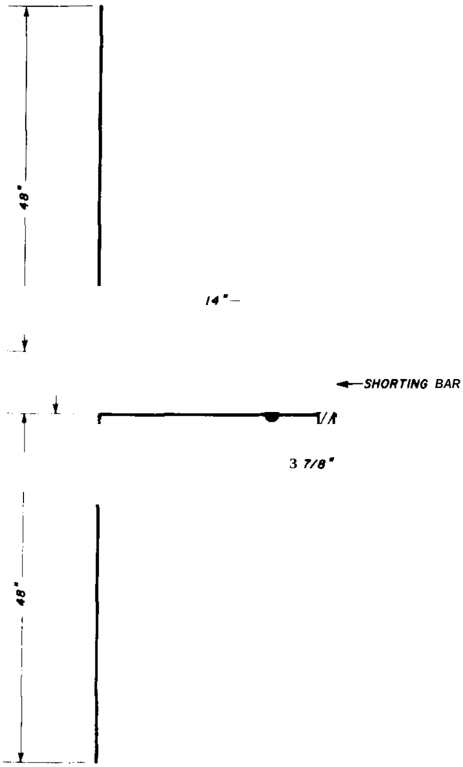
To keep tuning slugs put, use the wax drippings from a lighted candle, preferably a small birthday-cake type. If you use paint on the slugs, the paint may find its way into the threads and prevent the slug from turning — permanently!

D. E. Hausman, VE3BUE

simple 2-meter antenna

There are many commercial two-meter antennas available for home use, but here's a simple one you can put 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

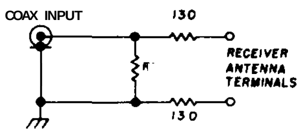


receiver impedance matching

Here's an idea if your receiver is designed for a 300- to 500-ohm transmission line and you want to use coax. For maximum performance, an impedance-matching device is a must to take advantage of everything the antenna delivers. The circuit of fig. 2 will do just that with little loss. A small perforated circuit board, 2 x 2 inches, will handle the coaxial jack and the three resistors; number-14 wire will support the board when it's connected to the antenna screw terminals on the back of the receiver.

John Devison, **WØZFN**

fig. 2. R1 is 51 ohms for 50-ohm coax, 75 ohms for 75-ohm coax; these are standard 5% values.



zener diodes

The zener voltage (rated voltage) of any zener diode is somewhat dependent on the temperature of the devices. This temperature dependence is more noticeable in higher voltage units. You will observe a voltage increase during the first few seconds after it is turned on; this is caused by heat generated by current passing through it.

The voltage rating of zener diodes is determined at some specified test current. Smaller currents will cause zener voltage to be slightly less than its rated value. Good engineering practice is to operate l-watt zeners at about ½ watt or less. This provides voltage temperature stability plus a safety factor for over-voltage conditions.

Several zeners of about equal voltage may be connected in series to obtain a higher wattage unit than would be possible from a single higher voltage unit. When two zeners are put in series, the zener voltage becomes the sum of the two voltages and the power rating also becomes the sum of the two units. (Example: a 12-volt and a 10-volt 1-watt zener are placed in a series; the result is a 22-volt, 2-watt unit.)

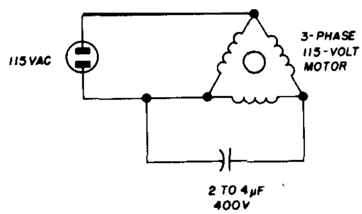
Don't connect zeners in parallel, even if they have the same voltage rating, because one of the units will take all the load.

M. Weinschenker, **K3DPJ**

three-phase motors

A three-phase blower motor can be run very easily from single-phase ac by connecting it as shown. 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**

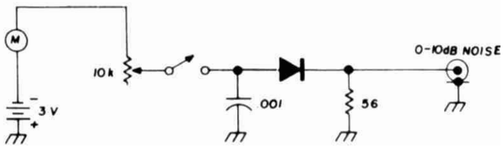


using noise generators

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. 1**. A silicon diode should be used (a 1N21 or 1N23) and held in place with a common fuseclip. The time you spend in constructing 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 mast 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 over-all 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.

• J. A. Huie, K2PEY, "A VHF Noise Generalor," QST, February. 1964, p. 23.
H. Olson, W6GXN, "The Noise Diode Caprr," QST, February. 1964, p. 28.

Bob Brown, K2ZSQ/W9HBF

new products

tornpkins tunaverters

An interesting new line of rf converters has been announced by Tompkins Radio Products. Six new models are available—several for the amateur bands. These converters feature a fixed output frequency of 1500 kHz in the standard a-m broadcast band. Tuning is done at the converter rather than at the receiver; a calibrated dial with 6 to 1 reduction is provided for this purpose. One crystal position is also included so that the unit can be changed to fixed-frequency operation with the flip of a switch.

Models for amateur frequencies include the 273X for 26.9-30 MHz, the 504X for 50-54 MHz and the 1450X for 144-150 MHz. Other models are available for police, fire, marine and aircraft bands. The unit is powered by a self-contained 9-volt battery. The X-line converters are supplied complete with a mobile-mounting bracket and are guaranteed for one year.

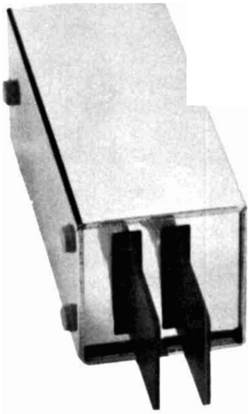
Accessories offered hy the manufacturer include crystals for your exact listening frequency and an adapter for coupling to the loop or external antenna of portable and home receivers. The price of the X-line converters without accessories is $32.95, postpaid from Herbert Salch and Company, Woodsboro H, Texas 78393.

westcom noise blanker

The new Westcom noise blanker is designed for the vhf operator who is looking for maximum noise suppression for reception of weak DX and scatter signals. This device uses four high-gain FET amplifiers with two stages of noise clipping to efficiently remove ignition noise and other forms of pulse interference.

Automobile ignitions, power distribution circuits, motors, radar and other sources of high-energy pulse interference cause severe problems in vhf receiving systems. The noise pulses excite tuned amplifier circuits in the receiver and produce mixer and i-f amplifier overload. In most communication receivers, the noise limiter is placed at the output of the i-f amplifier strip. However, by the time a strong noise pulse has passed through the i-f amplifier, it has been amplified and stretched to the point where it totally masks weak signals. To complicate matters, ssb receivers with fast attack-slow recovery agc systems will hang up on the amplified noise pulse and reduce receiver gain. Loss of receiver gain at this point will cause even moderate signals to disappear into the background noise. To distinguish weak signals in the presence of noise pulses it's desriable to remove the interference before it is introduced to the receiver.

The Westcom noise blanker is connected in the coaxial cable between the vhf converter and the receiver, and provides an effective means of suppressing noise at the converter output frequency. Since the noise pulses are eliminated before the pulse amplifying and stretching circuits in the receiver, the signal-to-noise ratio is improved significantly. The Westcom noise blanker is available with i-f ranges of 40, 20 or 10 meters. Other i-f ranges are available on special order. $29.95 from Westcom Engineering Company, P. O. Box 1504, San Diego, California 92119.



the permaflex key

The James Research Company has just introduced a new key for electronic keyers that is a unique departure from standard key design. This new switch mechanism is completely enclosed and combines the twin lever for electronic keys with a straight hand key in a pivotless two-paddle design. With this key, CW operators have an instant choice of automatic, semi-automatic, or straight hand keying in one compact unit.

Although the **permaflex key** was designed specifically to key the low currents associated with integrated-circuit keyers, it will handle transmitter currents up to 8 amps. The contacts are gold-diffused pure silver for protection from corrosion and are adjustable from zero to 0.06 inches. Paddle weight is adjustable from 5 to 50 grams. The paddles are made of rugged fiberglass epoxy with gold-plated copper conductors to ensure a long-term, low resistance, anti-corrosion keyer.

The cabinet is 16-guage chrome-plated steel, with an inner chassis that can be quickly removed for contact gap and tension adjustment. Tho low-mass paddle design and durometer rubber feet permit accurate high-speed keying with a minimum of "walking". $19.95 postpaid from the James Research Company, 11 Schermerhorn Street, Brooklyn, New York 11201.

mosley cubical quads

Mosley Electronics has introduced a new line of cubical quad antennas for single and multi-band operation. These quads feature well-insulated aluminum spreaders for greater strength, a light-weight, low wind load spreader mount that eliminates wind-resistant webbing at the hub and the time-tested Mosley boom-to-mast clamp with stainless steel U-bolts for greater quad stability.

These new cubical quads are designed so that the voltage and current minima and maxima are located between the corners. This eliminates possible signal loss due to spreaders intersecting with the wire elements at these critical points. These quads incorporate gamma matching with a single 52-ohm coaxial feed line to the antenna. The Mosley cubical quads maintain an SWR of 2.2:1 or better with exceptionally flat response across the full bandwidth.

Four models are available: the MCQ-10 for 10 meters, the MCQ-15 for 15 meters, the MCQ-20 for 20 meters and the MCQ-3B three-band quad for 10, 15 and 20. For more information on these new cubical quad antennas, write to Mosley Electronics, Inc., 4610 N. Lindbergh Boulevard, Bridgeton, Missouri 63042.

the man behind the mike

If you're interested in professional broadcast announcing, this brand-new book offers the practical help needed by most aspiring announcers and station and broadcast personnel. Many of the pointers contained in this book will help broadcasters suffering with growing pains to solve the lack of readily available professional talent; both seasoned veterans and neophytes will gain by reading this helpful book.

The Man Behind the Mike is based on many years experience in training beginners and practicing broadcasters, and offers practical and helpful guidance on every phase of announcing. It is well suited for everyone in broadcasting—managers and program directors will find a wealth of material to guide them in their daily role as well as a host of management and programming ideas. $9.95 from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

pin-point tv troubles in 10 minutes

If your family ever calls upon you to fix the television set, here's the book you need if you want to get the job over with in a hurry. It offers the type of help not usually found in books of this type. It contains large photos of different picture-tube troubles that are keyed to trouble-finding charts which identify over 700 possible defects. With this aid, you can pin-point almost any trouble in a tv set in a matter of minutes.

In addition, this book describes and illustrates methods for checking component performance. Also included are explanations of circuits used in the majority of tv receiver5 manufactured since 1953. For problems that require special troubleshooting methods, suggested procedures are outlined which will lead you to the trouble in a hurry. In addition, the beginning of each section gives information on circuit peculiarities, methods for improving set performance, service tests and adjustments and checking individual components. $4.95 paper; $6.95 hardbound from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

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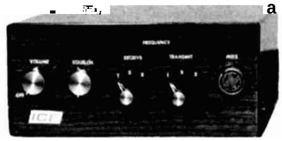

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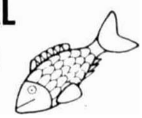

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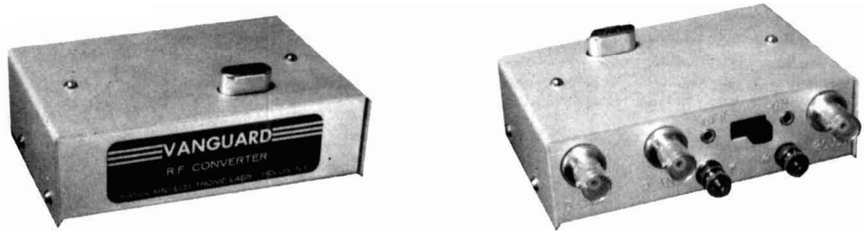

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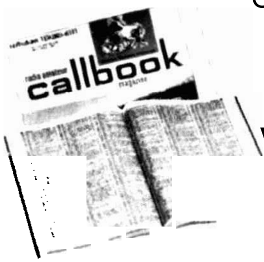

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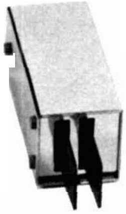

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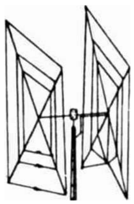

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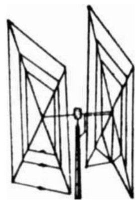

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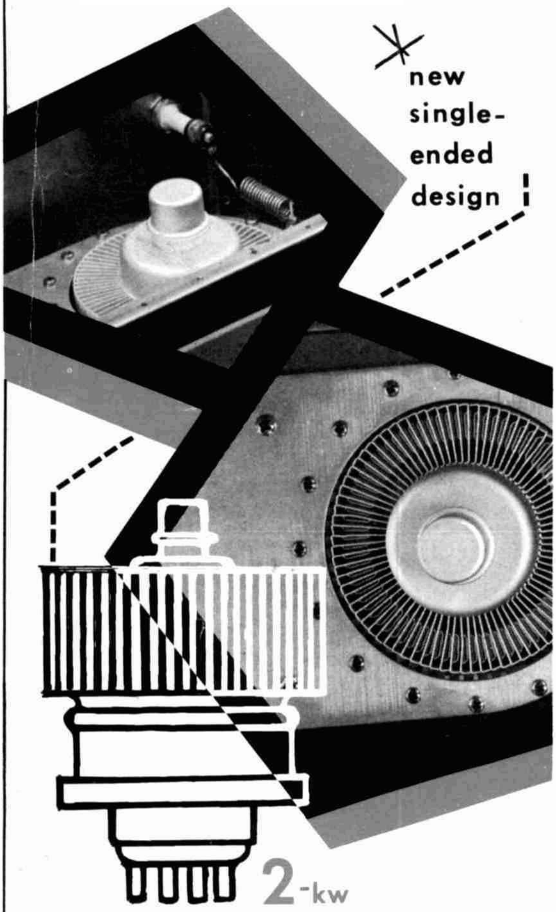

strip-line amplifier for 432 mhz

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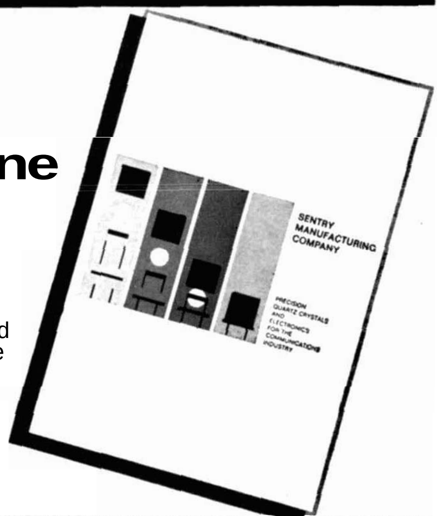

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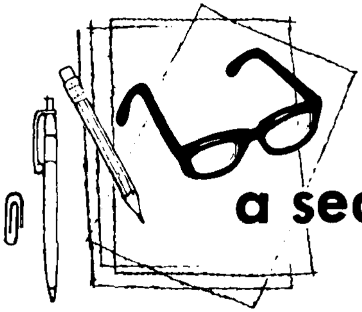

a second look

by jim fisk

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

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

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

Some of the other recommendations proposed for putting more transmitters into the same amount of space include: more "splitting" to reduce the originally-assigned bandwidth to make room for other users, narrowing the 6-MHz now allowed for TV stations so that adjacent and in-between frequencies can be used for other purposes, and tightening of standards for both receivers and transmitters to conserve bandwidth. This last proposal could have some interesting effects on amateur operation—for one thing, improved standards governing receiver susceptibility could reduce TVI problems. On the other hand, improved standards might carry a "type-acceptance" clause that could eliminate our traditional privilege of building our own equipment.

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

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

Jim Fisk, W1DTY
Editor

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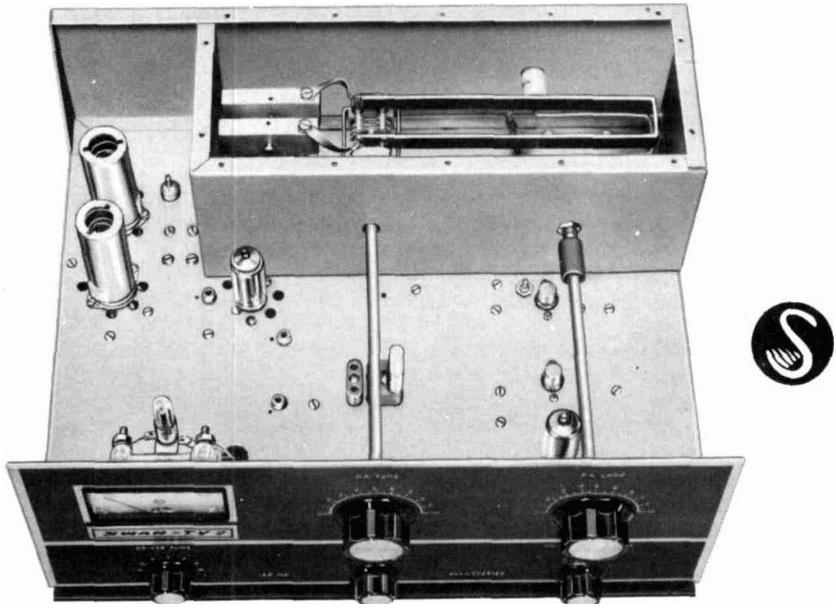

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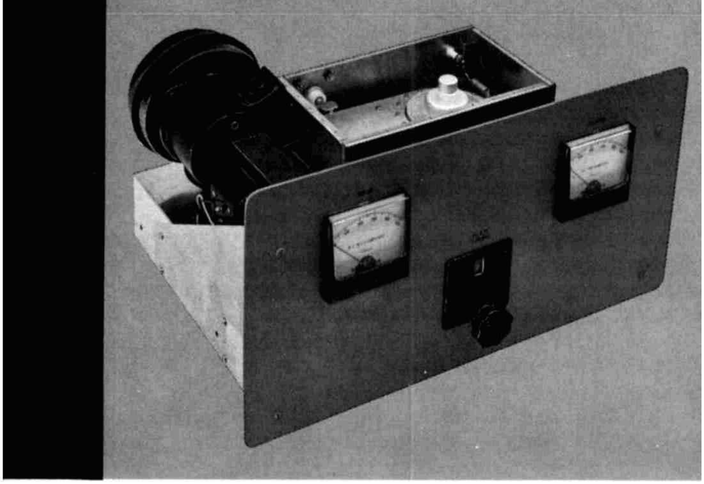


2-kw pep amplifier for 432 mhz

The combination of a new ceramic zero-bias triode and unique circuitry promises higher power and greater efficiency in high-power rf amplifiers for the vhf/uhf region

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

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

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

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

Since no operational data for the 3CX-

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

strip-line plate circuit

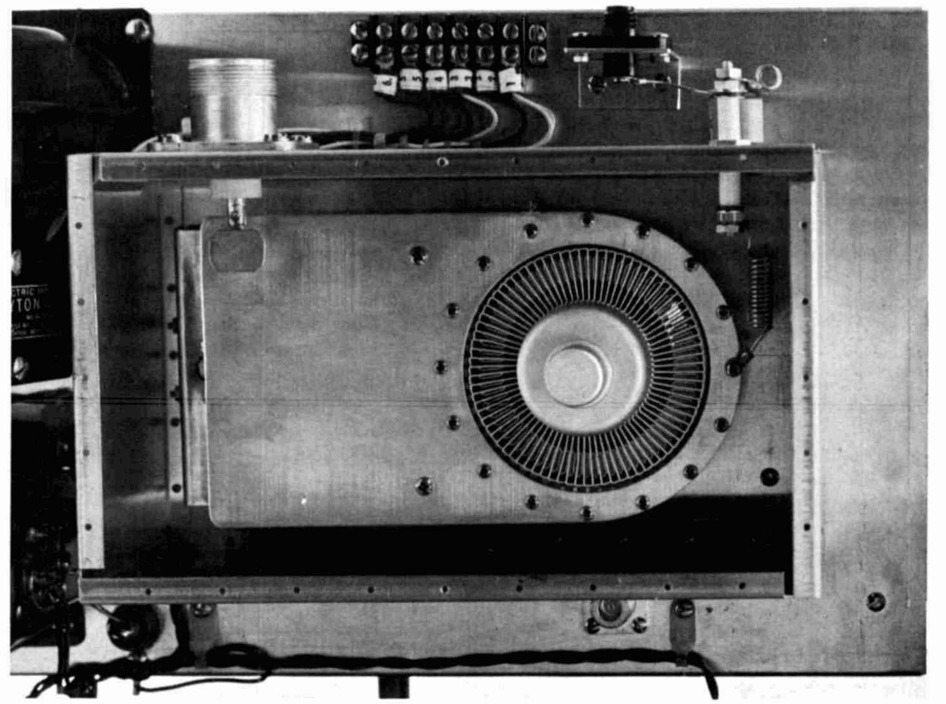
In the interest of simplicity, it was decided to try the simple strip-line plate tank circuit shown in **fig.** 2. Since the output capacitance of the 3CX1000A7 is only about 15 pF in the grounded-grid configuration, a half-wavelength plate line was chosen. Even so, a large portion of the line is swallowed up inside the tube because of its internal capacitance and rather large physical size. The half-wavelength line eased coupling problems and promised better operating efficiency than a shorter line.

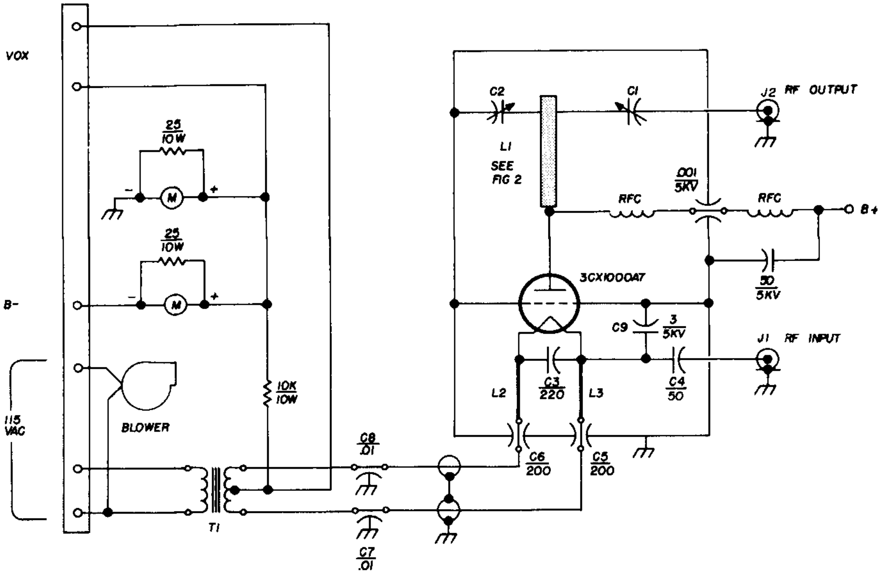
The input capacitance of the 3CX1000A7 is approximately 30 pF. We thought this seemingly large value would pose problems at 432 MHz. We hoped that excessive driving power wouldn't be expended in the charging currents flowing through the input capacitance and that the power gain of the stage wouldn't be excessively low. Both of these difficulties were overcome by properly designed tank circuits.

The plate circuit strip line consists of a 1/8-inch-thick copper plate, rounded at one end and placed in an aluminum cavity box. The rounded end of the strip line encircles the anode of the 3CX1000A7 and is connected to it with a matching copper collar, fig. 3, with flexible finger stock soldered to its inner circumference.

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

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





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

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

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

grid circuit

The grid of the 3C1000A7 is at nominal ground potential since the Eimac SK-870 air-system socket grounds the multiple grid terminals to the chassis. Even so, the grid structure within the tube is above ground at 432 MHz by virtue of the small but discrete cumulative inductance of the socket, grid terminals and grid cone assembly within the tube. A portion of the driving voltage appears across this cumulative grid inductance, and makes the circuit degenerative and reduces the input-to-output isolation of the tube.

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

filament circuit

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

amplifier construction

Amplifier construction is relatively easy and simple. The complete assembly is mounted on a 10 x 17 x 3-inch aluminum chassis, supported behind a 13-inch relay-rack panel. A space is left between the chassis and panel to accommodate the gear mechanism of the plate-circuit counter dial.

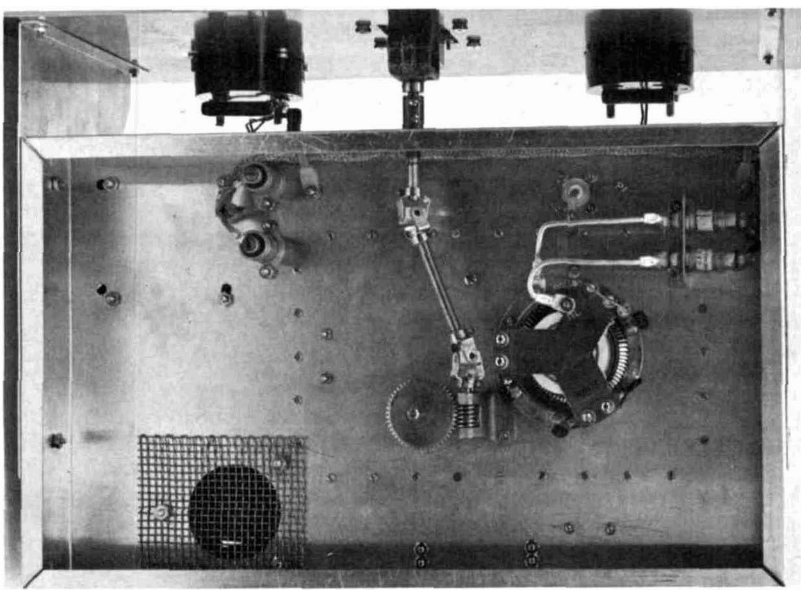
The plate-circuit assembly box is a standard aluminum enclosure measuring 6 x 4 x 10 inches, with removable 6 x 10-inch sides. The box is firmly bolted to the chassis by the lips of one of the open sides and the other side serves as the top of the enclosure.

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

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

*** Eimac CF-300. Available from Allied Radio Corporation, 100 N. Wrrtcrn Avcnuc, Chicago, Illinois 60680. Ordrr catalog numbcr 47E2087. $5.80 plus postage; shipping weight 12 ounces.**

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



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

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

The moving part of the hinge is jumpered with a wide strap of flashing copper. The hinge provides mechanical stability—electrical conductivity through the hinge joint is provided by the flashing copper. The capacitor is varied by means of an eccentric arm driven by an under-chassis worm drive from the counter dial. A '14-inch diameter teflon rod insulates the drive system from the hinged capacitor plate.

The filament line is made of two 4-inch lengths of number-10 copper wire spaced 3/4-inch apart. They are bent back to reach the filament bypass capacitors mounted on an aluminum bracket in the corner of the chassis. The filament circuit tunes quite broadly and is relatively uncritical.

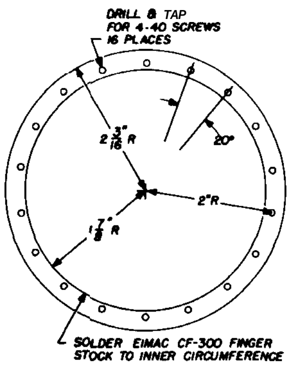
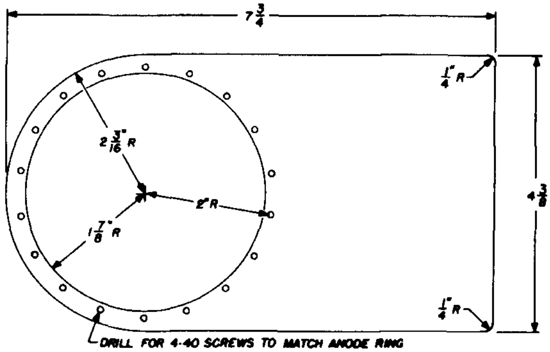


fig. 3. Anode ring for the strip-line amplifier. Make from 1/8" copper.

The rf drive point is attached to one leg of the line, near the midpoint, and may be juggled a bit to establish the lowest drive level after the amplifier is in operation. This is best accomplished with a directional wattmeter in the drive line. In the interest of conserving drive power, RG-225/U teflon coaxial cable is used to couple the driver to the amplifier.

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

fig. 2. Construction of the plate line. Make from 1/8" copper sheet and silver plate.



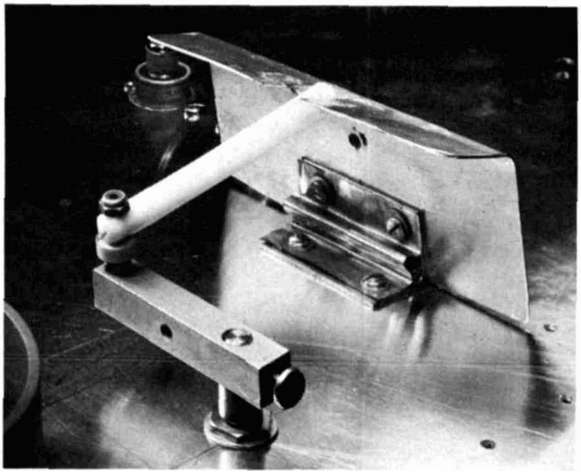
A small value of cathode bias—17 volts—is applied to the 3CX1000A7 to reduce the zero-signal plate current. In lieu of an expensive zener diode, a homebrew version was made by placing 13 bargain-counter 3-ampere stud rectifiers in series and using their summed forward voltage drop as zener bias. The rectifiers are rated at 50 volts PIV and provide enormous carrying capacity at minimum price. The diodes were mounted on a '14-inch aluminum plate with mica insulating washers and placed in the power supply unit. Alternatively, two 1N4561 50-watt zener diodes may be connected in series to provide a bias of 11.2 volts at a substantially higher cost.

The output circuit is capacitively coupled to the plate line by means of a small, semi-variable capacitor made of a copper tab supported by the center pin of the coaxial antenna connector. The output circuit of the amplifier is designed for heavy-duty RG-17/U coaxial line. The mounting holes of the coaxial connector arc made oversize; antenna coupling may he adjusted by loosening the bolts holding the connector to the chassis and moving it about in the mounting hole.

amplifier adjustment

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

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



For initial adjustment, recluced plate voltage, 2000 volts or so, and reduced rf drive are applied to the amplifier. The plate circuit is resonated for maximum output, and the antcnna coupling capacitor (C7) is adjusted for best power transfer. Coupling exists between the input and output circuits, and while the amplifier remains stable, grid current varies abruptly with plate-circuit tuning and loading. Grid current should run about 40 percent of the loaded plate current at all times.

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

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

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

operation

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

A word of caution: the 3CX1000A7 has not been rated for operation at this frequency nor for plate potentials above 3.5 kV. Experimental operations of this magnitude at this frequency exceed the warranty of the tube. In addition, it should be noted that full grid drive should not be applied without the plate circuit loaded and high voltage applied under any circumstances—otherwise the grid structure may be damaged. The data derived is based on an experimental test, and the Application Engineering department of the Eimac Division of Varian should be consulted before using this information for final equipment design.

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

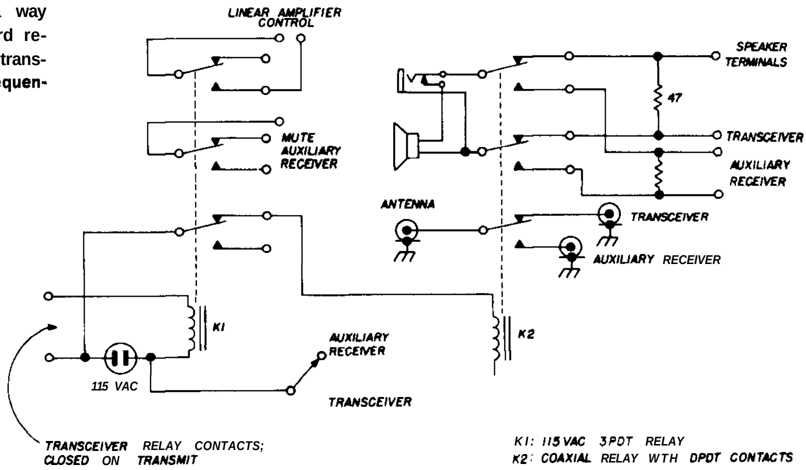
ham radio

using an outboard receiver with a transceiver

There are many times when it's helpful to have an outboard receiver available for listening on frequencies other than your transmit frequency. Remote VFO's accomplish this in most cases, but not always. A good example occurs when working DX stations on 80 meters. With your transceiver tuned to 3850 kHz, the preselector isn't peaked up on the European section of the band even if you have a remote VFO. Also, with a separate receiver, you can monitor other bands for activity. This is particularly useful when you're waiting for ten or fifteen meters to open up in the morning.

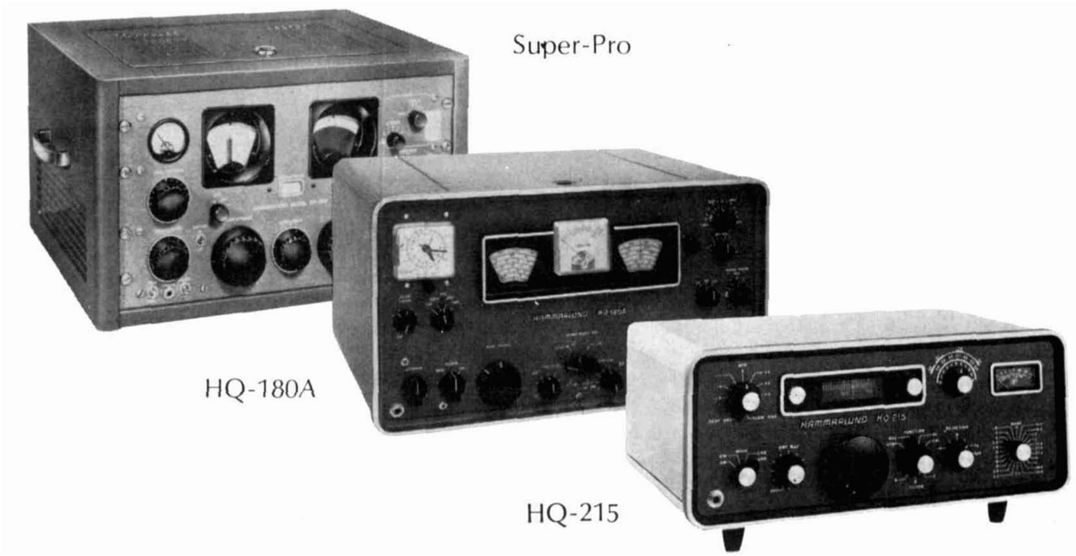
With the circuit shown in **fig. 1**, an outboard receiver may be switched onto the antenna and speaker at the flick of a switch. When the transceiver is switched to transmit, the speaker and antenna are automatically connected to the transceiver. Relay K1 is a 3PDT relay and K2 is a 115-Vac coaxial 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.

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



Jim Fisk, W1DTY

Super-Pro & Sons



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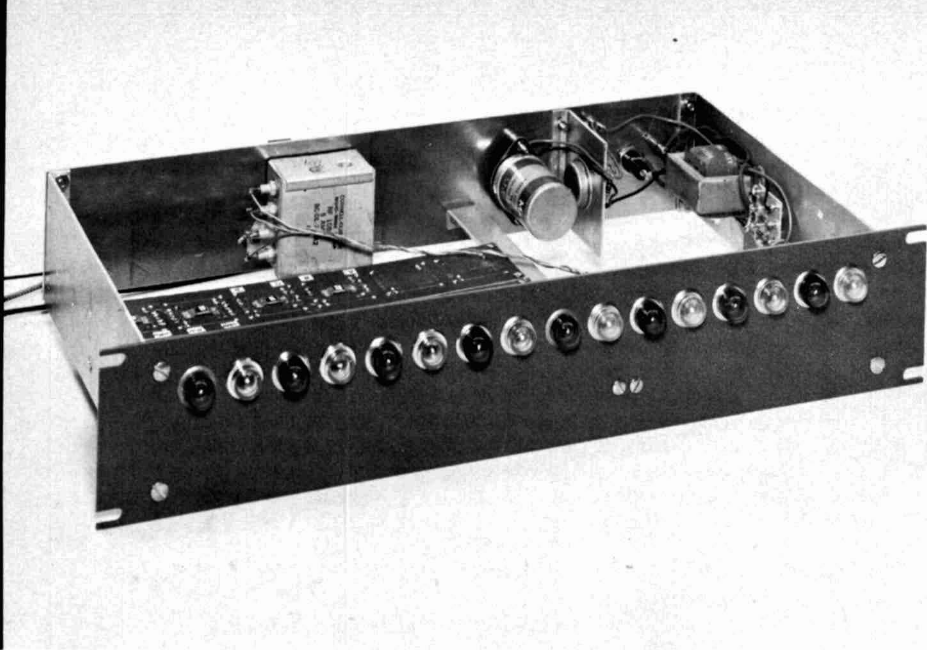
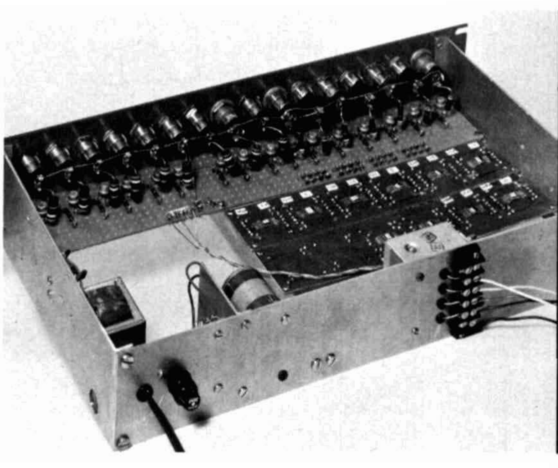


digital wind direction indicator

If you like to turn your beam into the wind when the wind starts blowing, why not try this simple system for remote wind-direction indication

Hank Olson, W6GXN, P. O. Box 339, Menlo Park, lifornia 9 25

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



indicator of relatively unusual design using integrated circuits.

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

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

digital shaft encoder

The heart of the indicator is the digital shaft encoder illustrated in **fig. 1.** This encoder is a very simple one that will indicate only sixteen discrete directions (points of the compass). This example is a binary-coded disc; note that there are four tracks on it. These four tracks cut the information-carrying region of any direction sector into four boxes. Since any box can be white or black, every direction sector has a different pattern of black and white boxes. If you doubt this, go through tnem and check. These black and white patterns can easily be converted to voltages by lamps and cadmium sulphide (CdS) photocells and transmitted to a remote readout. The wires necessary would be six; one common, four for CdS cells, and one for the lamp-supply voltage.

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

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

fig. 1. Binary-coded direction wheel.

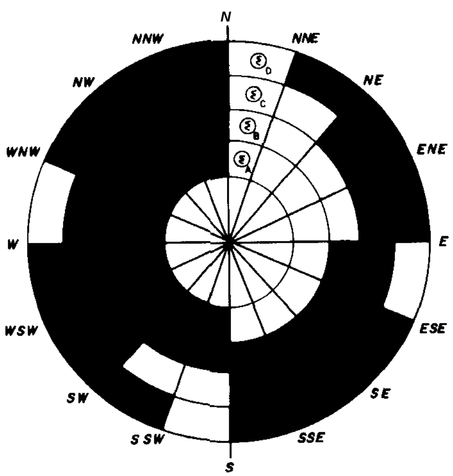
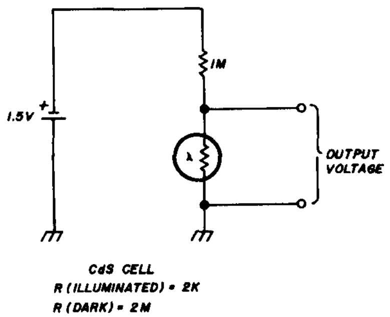


fig. 2. Cadmium-sulfide cell circuit.



digital decoding

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

table 1. Output from a binary-coded disc.

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

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

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

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

The 930 series of DTL integrated circuits come as close to an industry standard as any digital IC made. This series is made by Fairchild, Motorola, Raytheon, Sylvania and Texas Instruments, as well as several others. With such industry-wide acceptance of the 930 series, and with so many companies grinding them out to mil specs, it is only natural that a large quantity of rejects are available through the surplus electronics emporiums. Such is the case in the San Francisco Bay Area, where 930 DTL series units are produced, and where one local surplus store offers them at ten for a dollar—mixed, unmarked, and of unknown worth. I built the wind-direction indicator shown in the photos from 930's and 946's that were gleaned from $5.00 worth of such offerings.

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

fig. 3. Gray-scale direction wheel.

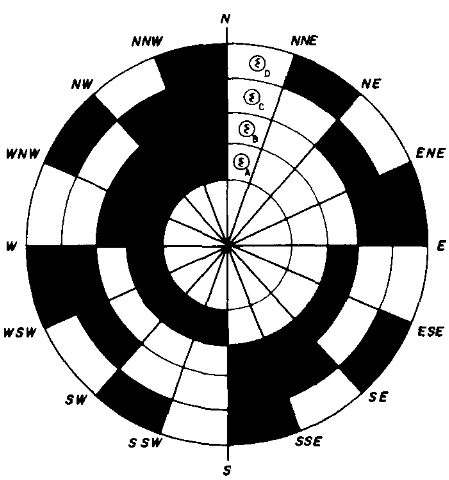
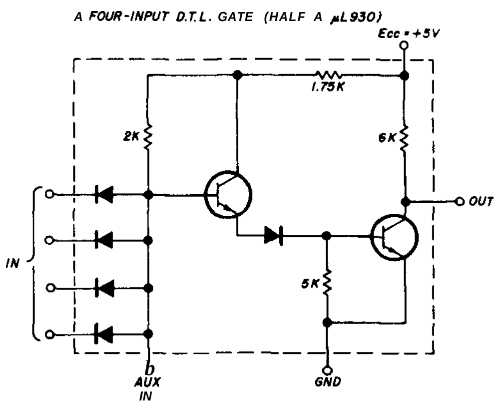


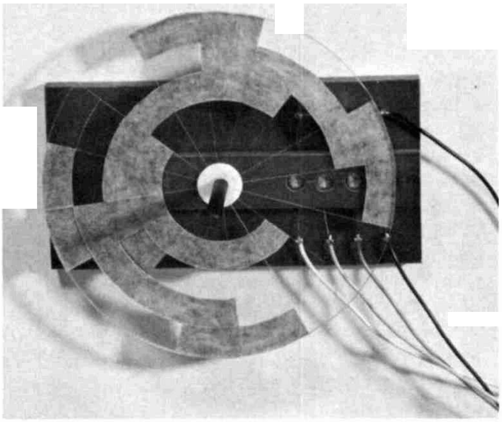
fig. 4. A four-input DTL gate using one-half of a Fairchild μL930.



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

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

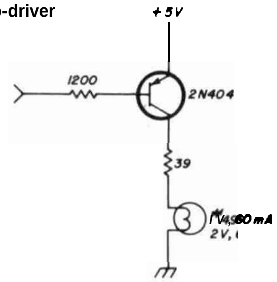
The way these DTL gates are used is the crux of this system. To understand it we must take a look at the basic DTL gate. The circuit of a 930 four-input gate is shown in **fig. 4.** With all four inputs open-circuited, or connected to +1.5 volts, the output level will be nearly zero since both transistors are conducting. If any one or more inputs are grounded, both transistors stop conducting, and the output level goes up to at least 3 volts. We can sum this up by saying that the output is low when all the inputs are "high," and the output is "high" when any input is "low."



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

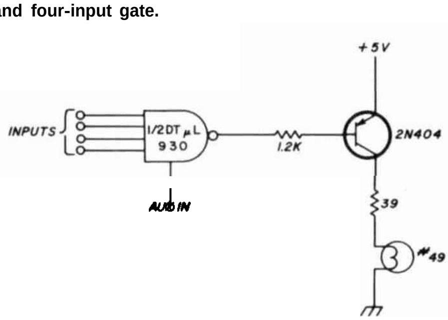
The only exclusive state for the gate is when all the inputs are high, the "and" condition. That is, input number 1 is high and input number **2** is high and input number 3 is high and input number **4** is high. This is called **NAND** logic (short for Negative **AND)** since the output has a logic state opposite from the inputs. Since the exclusive state will be the one that lights the indicator, a "low" gate output is used for the turn-on signal. The circuit of **fig. 5** will turn the lamp on when driven by a "low" gate output.

fig. 5. Lamp-driver circuit.



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

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



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

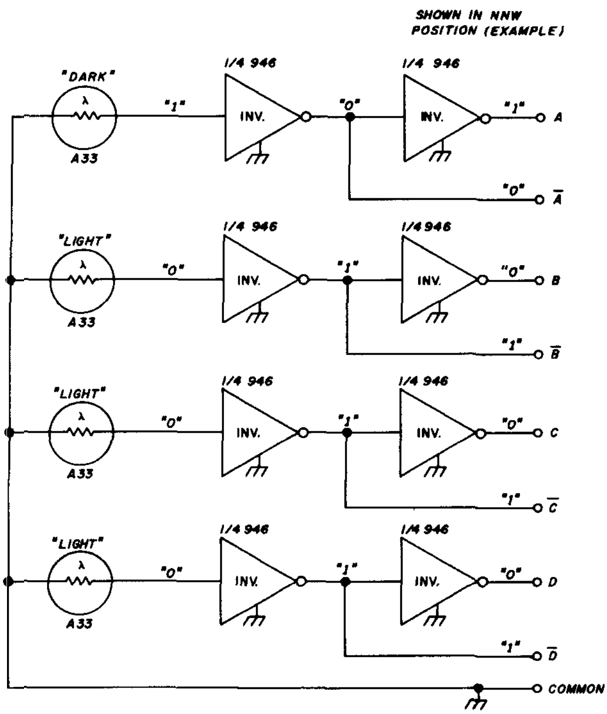


fig. 7. Circuit for generating a logic output from the gray-scale direction wheel.

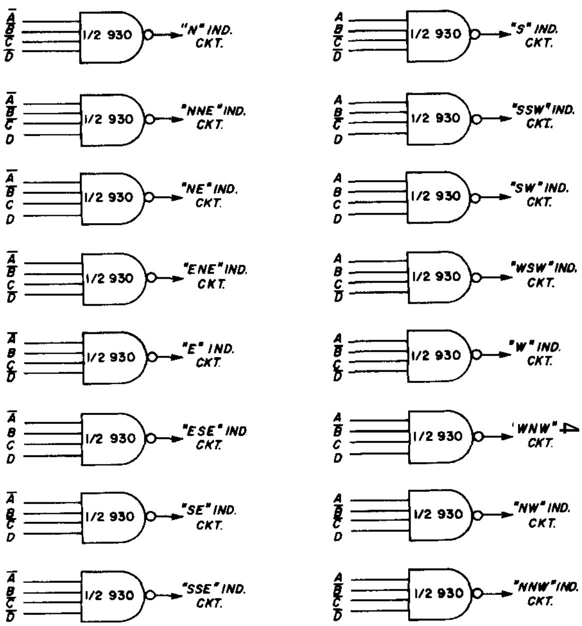


fig. 8. Decoding the logic output from the gray-scale direction wheel.

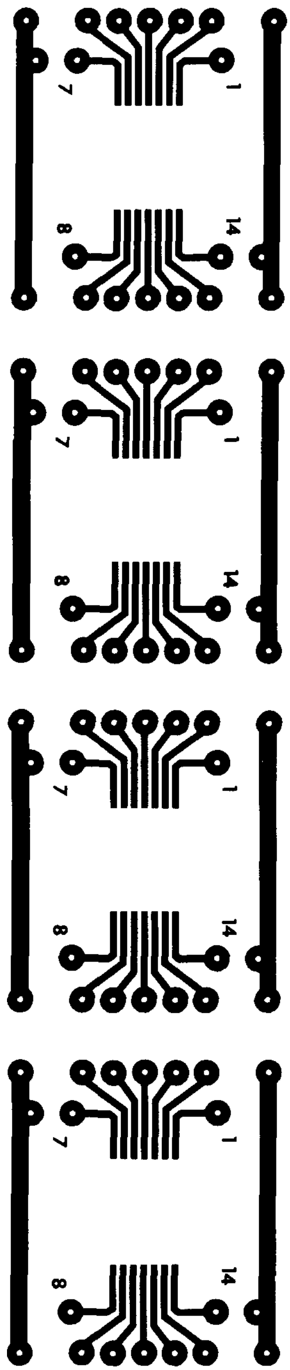


fig. 9. Layout of the etched circuit boards used to mount the integrated circuits. Wiring between IC's is done with small insulated wire.

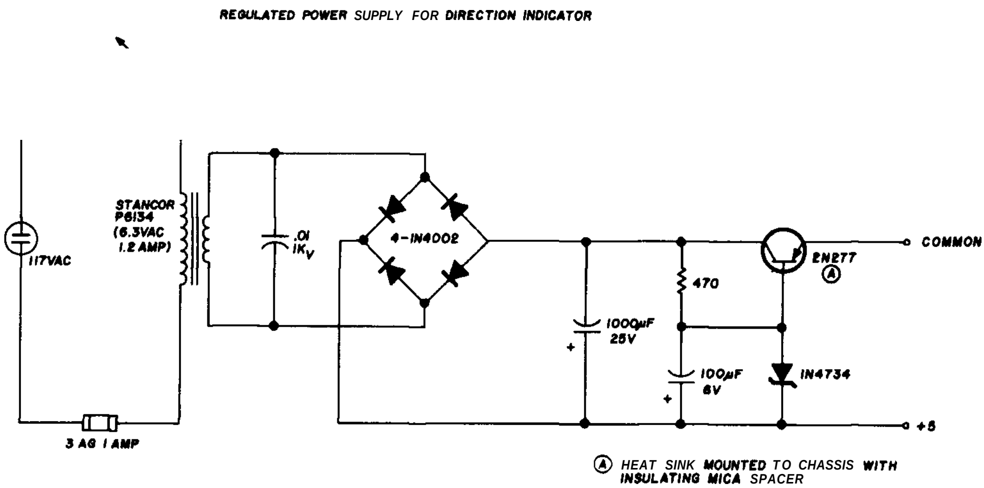


fig. 10. Regulated power supply for the direction indicator.

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

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

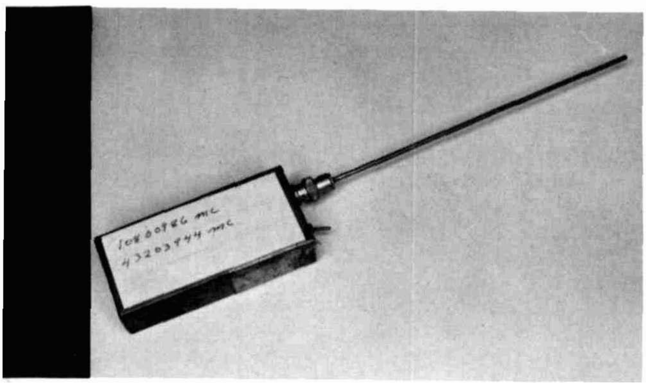
construction

The details of construction are shown in the photos. The indicator unit is built around a 3-1/2-inch rack panel with odd bits of scrap aluminum to form a box-type arrangement. The power supply is located in the right-rear corner. The indicator circuits are built on perforated Vector board (64AA18) in the front of the unit right next to the indicator lamps. The IC's are mounted on two six-section strips of etched-circuit board at the left center. The layout of these etched-circuit boards is shown in **fig.** *9.* Note that an rf-interference filter was put in each of the four input signal lines to prevent false-triggering the DTL circuits with your transmitter. The particular filters I used are simple toroid-L, coaxial-C filters from the surplus store, but any good rf lowpass filter should do as well.

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

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

ham radio



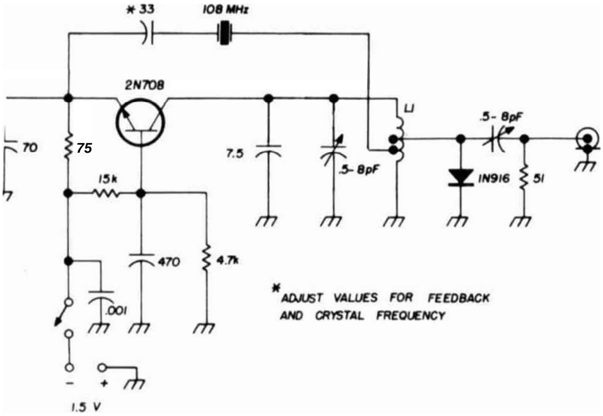
calibrated signal source for 432 and 1296 mhz

Del Cro ell, e IL, 4Mo rga St.
unta Vi Califorin 94 040

Find your way on 1296 and 432 MHz

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

The signal source is simply a 108-MHz crystal-controlled oscillator using a single 2N708 transistor with a 1N916 diode connected from the output tap on L1 to ground for generating harmonics When the output is displayed on a Hewett-Packard spectrum analyzer, the twentieth harmonic is still quite large. The spectrum chart in **fig.** 2 shows the output before and after the diode was in-fig. 1. The calibrated uhf signal source. L1 is 5 turns number 20, space wound on 1/4" diameter: tapped 1 turn end 2 1/2 turns from ground end.



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

The signal source can be built into most any type of package—I used a home-made sheet-brass box 2 1/2 x 5 x 1 inch. Make all the leads as short as possible. A 6-'14-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 1kHz at 432 MHz. Drift is very slight—with a stable BFO, I can only detect approximately 200-Hz drift during a 10- to 15-minute period. This is apparently due to slight voltage drops in the flashlight cells.

Construction of the calibrated signal source.

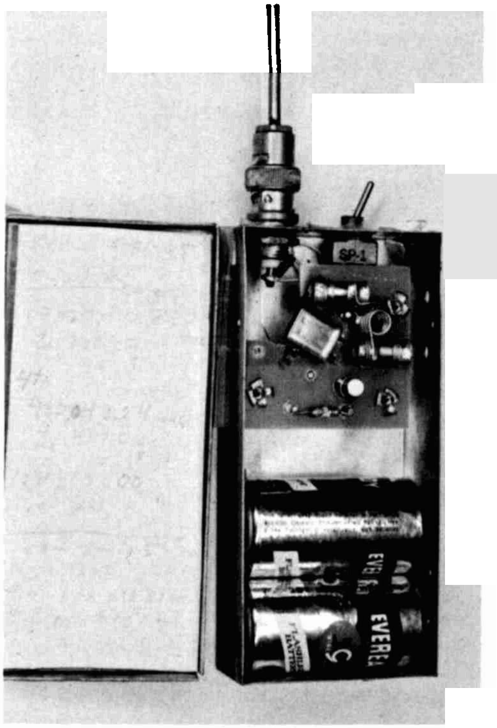
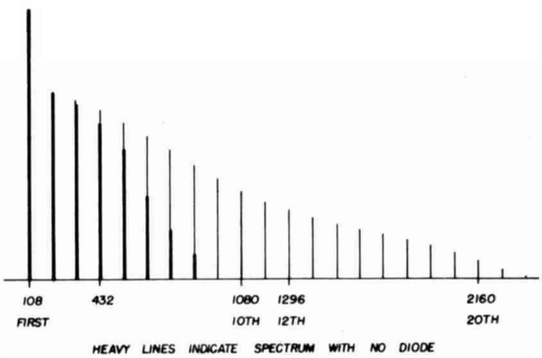


fig. 2. Spectrum output of the signal source as measured with a Hewlett-Packard 8551 analyzer. Heavy lines indicate spectrum without diode.



ham radio

frequency translation in ssb transmitters

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

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

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

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

the simplest system

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

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

To translate the desired sideband upward to an operating frequency, a heterodyne mixer is used. A variable-frequency oscillator (VFO) furnishes a signal that beats with the sideband from the filter and produces a sideband at the desired frequency. In the process, the simple mixer can't avoid also producing a carrier at the VFO frequency and an image sideband (as far from the VFO carrier as the desired sideband, but on the opposite side).

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

The reasons for going through this process may not be obvious. First of all, the isolated sideband can't be raised in frequency by simple frequency multipliers, as in non-ssb transmitters, because they would lose their identity completely. In the second place, a constant carrier frequency in the balanced modulator means that the resulting sidebands can always be fed to the same filter. If there were a lot of different frequencies, a different filter would be needed for each one. It's much easier to heterodyne or **translate** the fixed frequency up to the various desired ones.

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

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

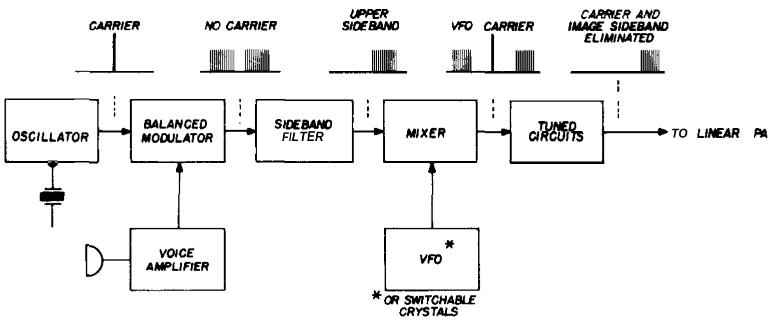


fig. 1. Simplest means of translating a sideband from a carrier-generated frequency to an operating frequency.

double heterodyning

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

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

The first frequency translation takes place in the first mixer. The VFO is tunable from 2.5 to 2.7 MHz; a frequency of 2.6 MHz (2600 kHz) is chosen for the example. Again, because of the heterodyne process, two sidebands are produced, but they are far apart. The desired one encompasses the sideband frequencies from 3050.65 to 3053.55 kHz; the other is an "image," with frequencies from 2146.45 to 2149.35 kHz. The desired sideband is still a lower sideband even though it has an "upper" position with respect to the image. The desired sideband is the lower sideband of 3055 kHz (2600 + 455 kHz).

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

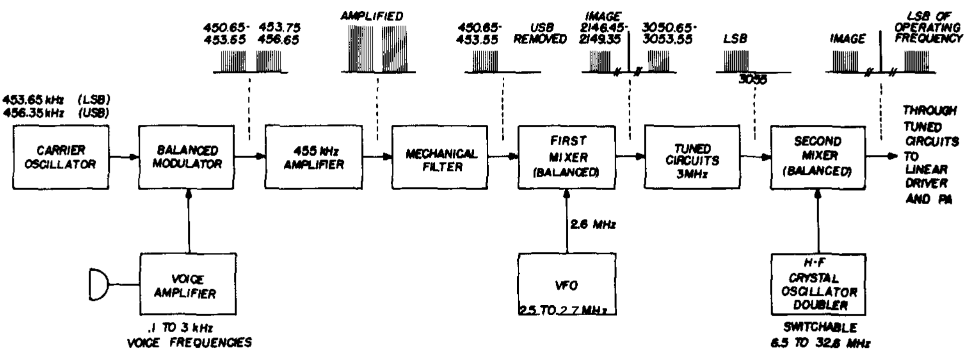
The 3055-kHz sideband must still be raised to the operating frequency. The second translation is handled much like the first. A switchable crystal oscillator supplies an rf signal for the second balanced mixer. Beating with the sideband signals that were produced in the first mixer, the rf signal develops a single-sideband signal in one of the ham bands. The band depends on the crystal selected in the hf oscillator, and the exact frequency depends on the setting of the VFO. An example will show you how this works.

Suppose you want to produce the lower sideband of 14.25 MHz. You set the switch of the hf oscillator to the crystal that places the output frequency in that vicinity. The crystal for this happens to have a frequency of 8.6775; but its oscillator is a doubler, so its operating frequency is 17.3550 MHz. The lower sideband of 14.25 MHz lies from 14.2470 to 14.2499 MHz. The sideband signal fed to the second mixer must therefore be the difference between those frequencies and the hf crystal frequency; that sideband covers from 3.1051 to 3.1080 MHz.

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

On the front of the transmitter, the hf-crystal switch would point to the 14.2-MHz sector of the 20-meter band, and the VFO dial would indicate 50 kHz. The combined readings would signify an operating frequency of 14.25 MHz. The transmitter output would be the lower sideband of that frequency.

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



frequency synthesis

Developing bands of frequencies by one translation and developing the frequencies within that band by another are excellent reasons for using double and triple heterodyne systems. Frequencies can be spread out wider than with any other system. Bands can even be sectored, and the VFO range used to cover only a portion of each ham band—thus spreading the frequencies even wider and making it that much easier to tune a particular operating frequency.

There's another way this can be done—by a method called frequency synthesis. The chief principle behind synthesis is illustrated in the transmitter diagramed in fig. 3. For simplicity, the frequencies are marked without reference to the sidebands; you know that what comes through the 9-MHz amplifier is actually one sideband or the other of 9 MHz. The same is true of the frequencies following the mixer.

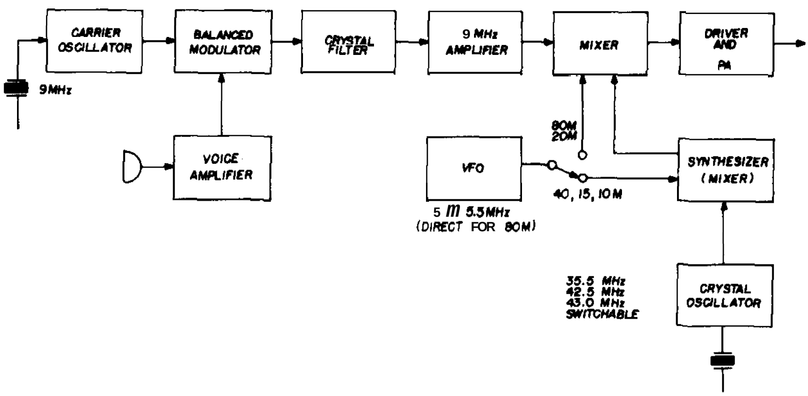


fig. 3. The basic setup for fraquency synthesis. The translating signal itself is heterodyned up before it is mixed with the sideband for the translation process.

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

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

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

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

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

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

triple heterodyning

From the words, you can figure out that

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

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

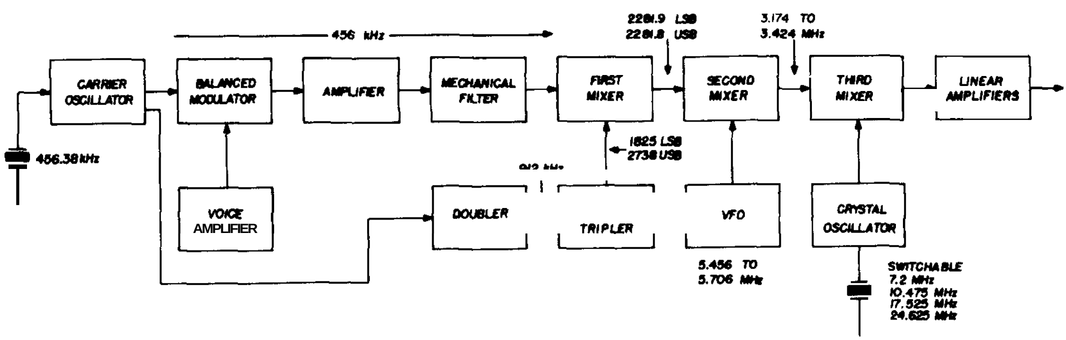
An interesting example of triple translation in a ham ssb transmitter is in the Sideband Engineers SB-34 transceiver. Fig. 4 is a block diagram of it. An interesting thing about this one is the use of the carrier oscillator to also furnish the rf signal at the first mixer. By careful choice of the carrier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

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

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

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

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



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

mixers that translate ssb

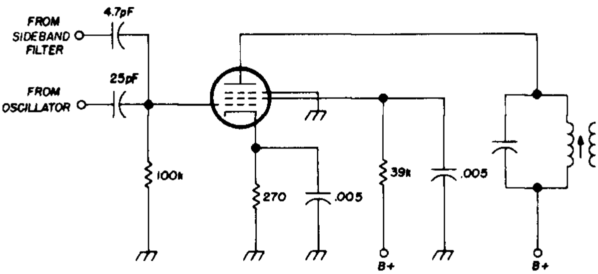
In most single-sideband ham transmitters, the mixer circuits are ordinary tube or transistor mixers. In one transmitter I know of, a semiconductor diode mixer is used for translating the carrier frequency to an intermediate frequency. Typical tube and transistor transmitter mixers are shown in **fig.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

These are not the only configurations used. by any means, but they are typical. Tube mixers are usually pentodes in modern ssb transmitters; seldom do you find a triode used for this purpose. The two signals are merely coupled to the grid, mixed inside the tube, and fed along to the next stage.

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

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

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



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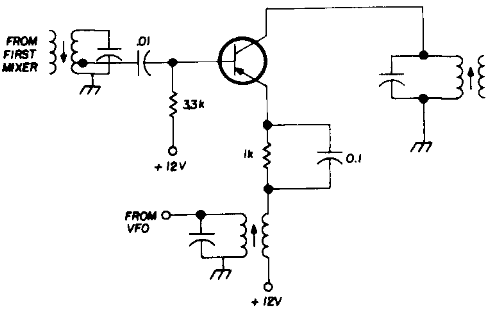

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create a problem. When two signals are beat together in a nonlinear mixer, the output consists of the two original frequencies, their sum, and their differences. As an example, suppose the VFO in the transmitter of **fig. 4** is set at 5500 kHz, and the lower sideband has been chosen. The frequencies applied to the mixer (the transistor in fig. *5)* are 2281.9 and 5500.0 kHz. The output consists of four frequencies: 5500.0 kHz, a sideband of 2281.0 kHz, a sideband of 7781.9 kHz (sum), and a sideband of 3218.1 kHz (difference). Picking out the right one is the job of the tuned circuits following the mixer. In this example, a broadband tuned circuit centered around 3.3 MHz can do the job. Only the sideband of 3218.1 kHz gets through. The tuned circuit thus eliminates the new carrier that was generated as part of the translating process, as well as the original sideband and the new image sideband.

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

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

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

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

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

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

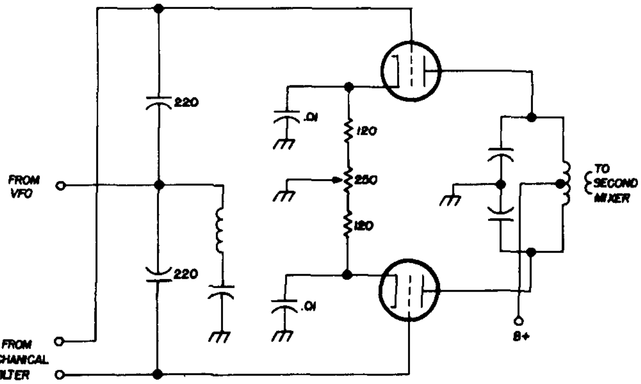


fig. 6. Balanced mixer used in some ssb transmitters to eliminate the carrier that is generated by frequency translation.

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

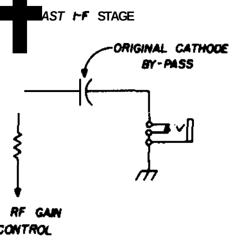
Next month I'll delve into another little-understood facet of the modern ham ssb transmitter: voice-operated transmission, better known as VOX. I'll explain various methods of accomplishing this type of hands-off operation. Also, we'll take a quick look at MOX—the manual version, usually called **PTT**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

ham radio

the i-f cathode jack

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

As long as nothing is plugged into the jack, it is a short circuit and the receiver works exactly as before the modification. When a phone plug is inserted, the i-f stage becomes a cathode follower, and provides a low-impedance i-f output for driving a Q-5'er, fm adapter, monitor scope, etc. An ac 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 measurements requiring dB comparisons or rf signal levels. Be sure to turn the agc off.



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

Fred Brown, W6HPH

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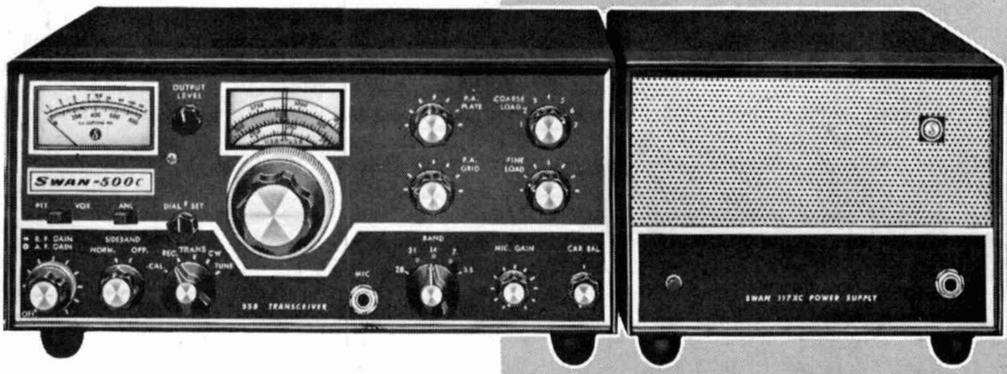

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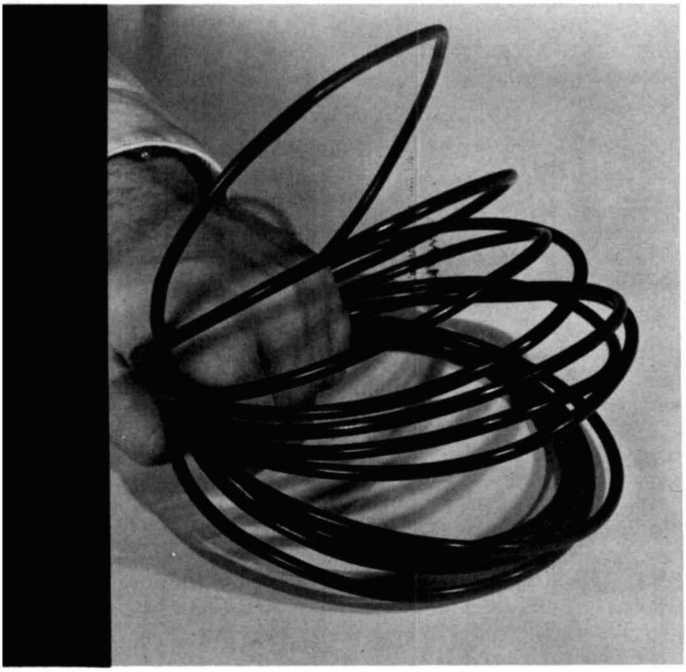


what you should know about standard RG-type coax cable

C arlesC am'llo, 3 9I, Amphenol Cable Division, Bunker Ramo Corporation

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

behind the problem

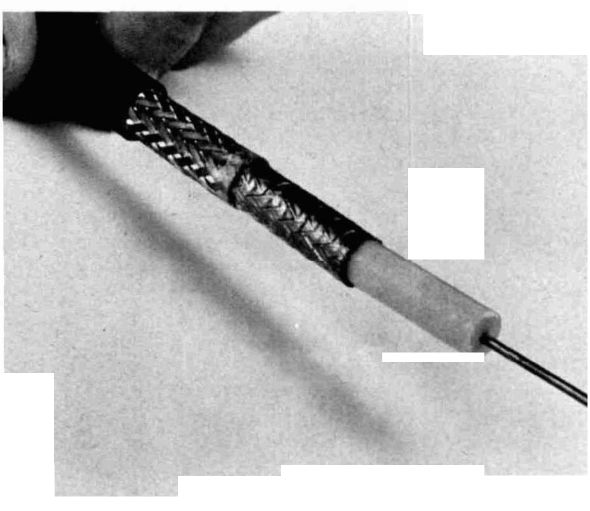
To all too many, standard coaxial cable is a foolproof commodity which can normally be bought "blind." Few realize that today a feedline made by one manufacturer can exhibit completely different characteristics from that made by another—even though both cables carry the same RG designation.

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

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

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

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



determine characteristics, spot problems

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

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

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

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

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

watch for capacity

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

To cope with this problem, coaxial cables are rated in terms of dielectric constants. As the constant approaches 1.00, the more nearly the capacity (and subsequent attenuation) approaches the low figure of open-wire lines. Knowledge of this figure allows you to analyze frequency-handling capabilities of the coax in question.

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

seldom exceeds 8.02 per foot.

rf power attenuation

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

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

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

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

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

solve migration problem now

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

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

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

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

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

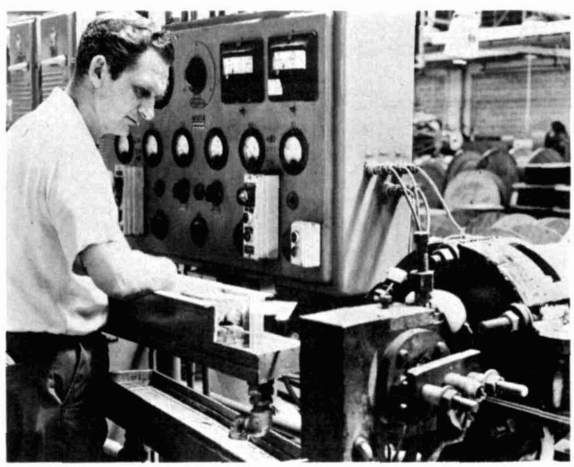


table 2. Sampling of popular **foamed polyethylene** dielectric coaxial cables.

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

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

determine power requirement

Note the maximum power ratings for standard communications cables listed in **table 1**. It is important to know both the power-handling capability of the cable in question as well as the specific frequency intended for use. The ratings in **table 1** assume **a** perfect match between transmitter and line, and line and antenna. If vswr is high, power-handling capability diminishes and losses run high as dielectric heating occurs.

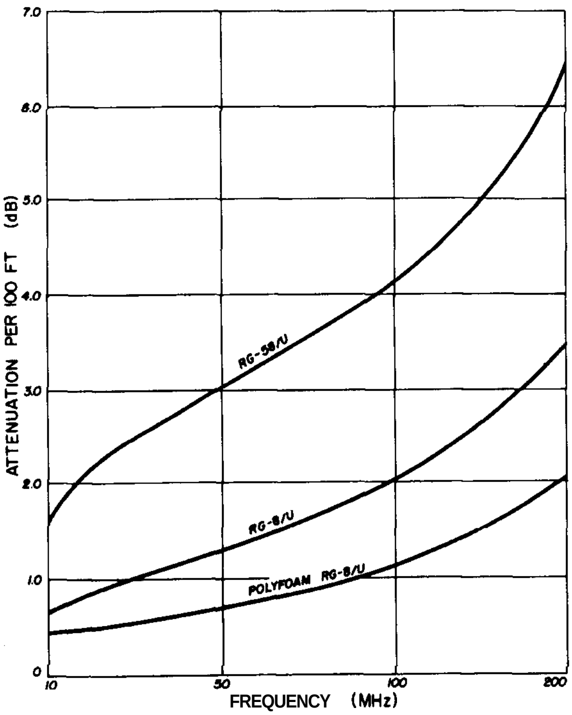


fig. 1. Attenuation curves for three popular types of coaxial cable.

check conductors

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

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

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

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

cable selection checklist

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

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

table 3. Coaxial cable joint jacket characteristics.

MIL-C-17 designation	jacket type	temperature limits
type I	black polyvinyl chloride	−40° C to +80° C
type IIa	black polyvinyl chloride, non-contaminating	under ¼", −55° C to +80° C over ¼", −40° C to +80° C

3. **Power:** What is theoretical corona voltage? Actual voltage?
4. **Jacket composition:** Is it of non-contamination material, according to MIL specifications? What are the changes in attenuation at 432 and 1296 MHz after aging tests?
5. **Jacket tightness:** Does it fit tight enough to show braid marks clearly, or is it loose, possibly indicating poor extrusion and instability?
6. **Connectors:** Has the right coaxial connector been selected for use with the standard RG cable chosen? Is it designed specifically for this purpose, or does it require an adapter?

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

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

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lightweight headphones

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

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

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

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

Don Miller, W9WNV, has been pictured wearing the Monoset unit, which has the transducer in the junction under the chin. Although these have been offered in 125 and 2000 ohms, Newark now shows them as the "new 799" in 15, 500 and 1000 ohms too. For those of you with 500-ohm equipment such as Collins, this appears to be the best solution.

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

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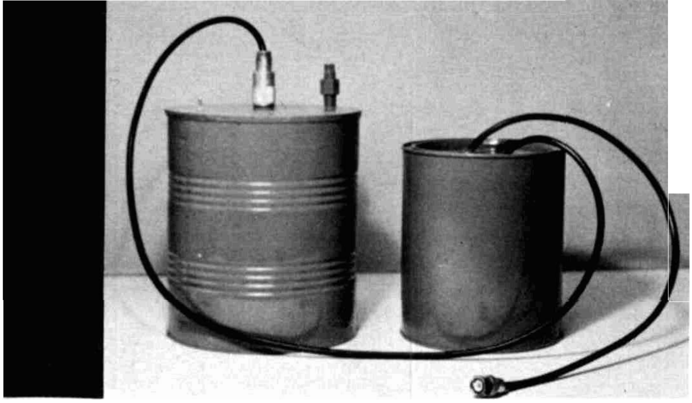


experimental dummy loads

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

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

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

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

In addition, at vhf the reactive effects can become rather formidable. Commercial practice is to construct the load shield in the form of tapered line sections to match the load impedance at all points as shown in fig. 1. Notice that the tubular resistor element is long compared to diameter. Also, the resistor-to-connector lead is in the form of a cone that tapers from the outside diameter of the resistor element to the diameter of the connector center conductor.

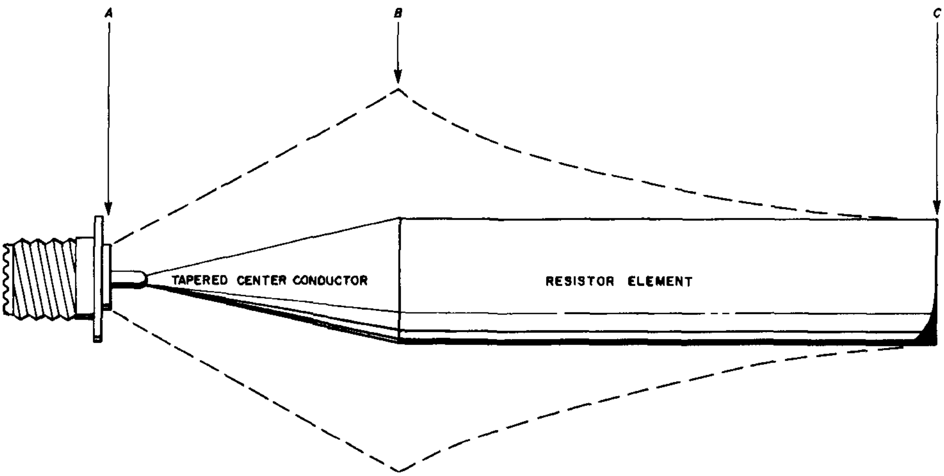
The coaxial structure between point A and B is a tapered line section which is designed so that its characteristic impedance is the same as the impedance of the resistive element at all points. The resistive element starts at point B, and from that point to point C the impedance diminishes uniformly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

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

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

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

fig. 1. Gaomatry used in commercial loads to maintain the load imprdanca at all points.



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

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

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

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

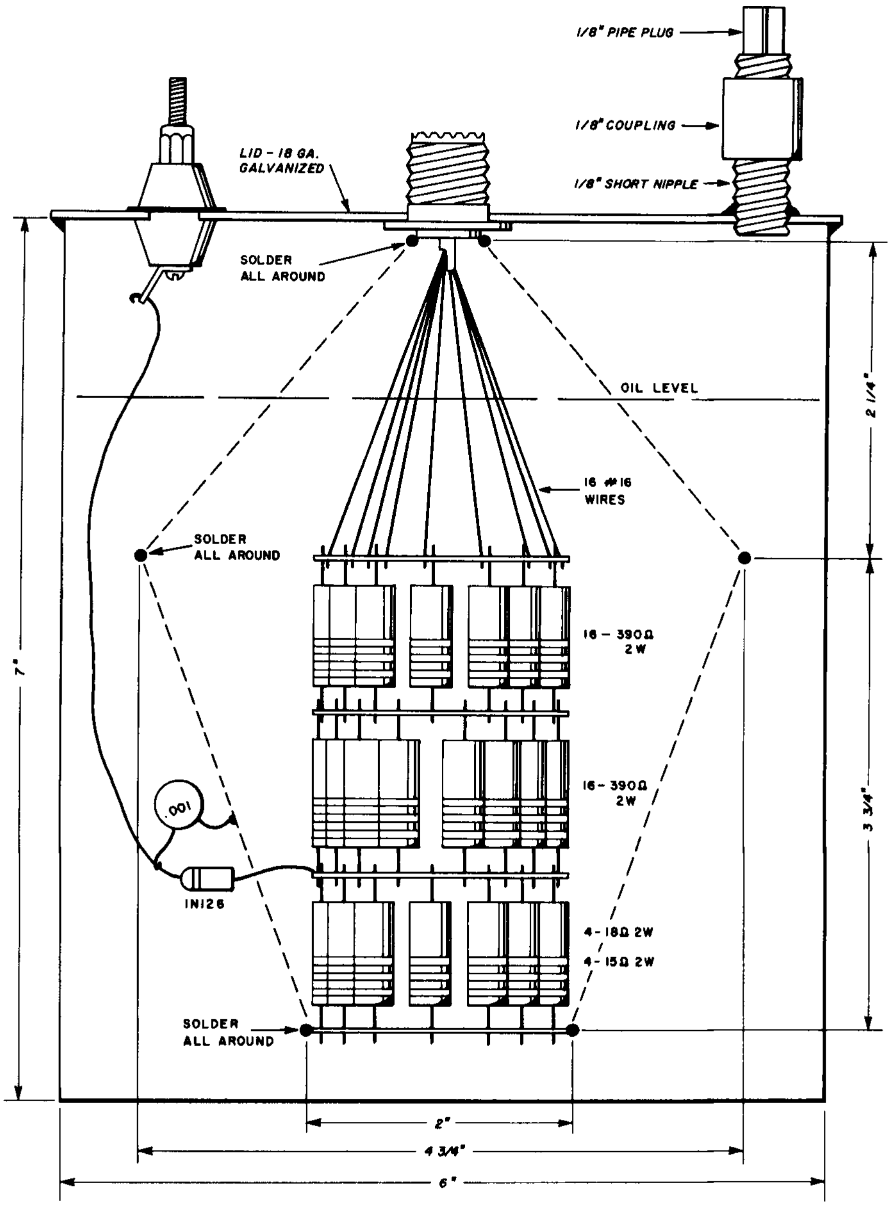


fig. 2. The air-cooled dummy load built by W8YFB. The performance of this load at various frequencies is plotted in fig. 3 and 4.

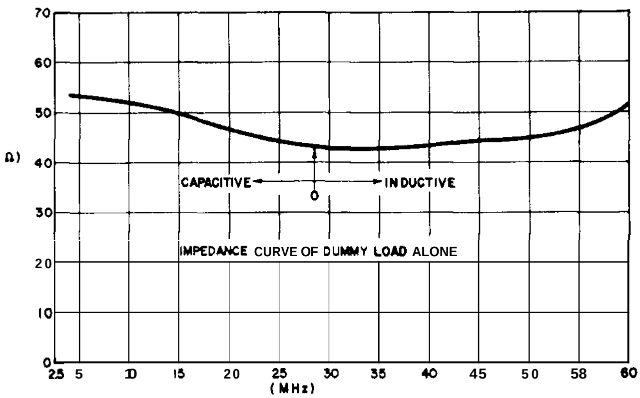
the load leveler

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

For example, suppose you were to feed a vhf transmitter into an SWR bridge, then out to a long length of coax which could be open or short circuited at the far end. If the cable were long enough to represent a 10-dB loss and you fed 100 watts to the line, only 10 watts would actually arrive at the far end. The remainder would be dissipated in line loss. Since the 10-dB loss works both ways, the line would appear beautifully matched, because the reflected signal would be 1/100 of the forward power indicated on the SWR bridge—and with an infinite mismatch at that! This, incidentally, is the reason why a vhf antenna should be measured at the antenna rather than the transmitter end of the line.

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

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



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

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

Similar hot spots will occur at half-wave intervals all along the line coiled up in the can, and inside the can, heat dissipation is impeded. It helps heat dissipation if the inside layers of coiled coax go to the dummy load instead of the transmitter. This is particularly true if the inside layers are coiled on a form. I used a I-112-inch wood dowel as a form.

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

choosing the resistors

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

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

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

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

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

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

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

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

$$I = E/R = 15.4/40 = 0.385 \text{ A}$$

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

$$I = E/R = 15.4/50 = 0.31 \text{ A}$$

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

$$0.385 + 0.31 = 0.695 \text{ A}$$

8. Determine total power capability for load:

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

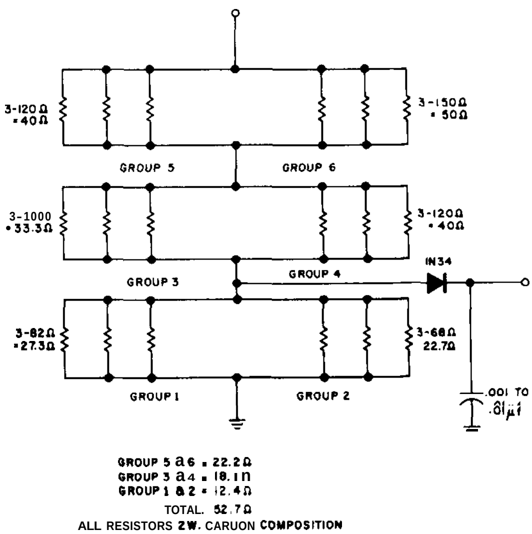


fig. 5. Method of setting up resistance values to obtain a satisfactory dummy load.

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

Some of you may question the value I obtained in step 5. Since we have eighteen 2-watt resistors, you may think that the load should be capable of dissipating 36 watts. However, remember that the current that flows through banks 1 and 2 also has to flow through banks 3 and 4 and 5 and 6. If you raise the current through banks 1 and 2 to the point where they are dissipating their maximum power, the same current exceeds the ratings of the higher resistance banks.

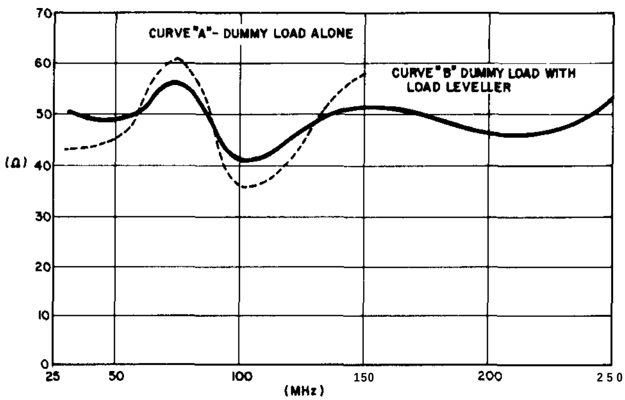


fig. 4. Impedance of the dummy load between 30 and 250 MHz. The effect of the load leveler at the upper frequencies is quite evident.

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

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

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

construction

Make four rings of thin brass shimstock, dairy tin (light gauge steel which is tinned on both sides) or light galvanized iron. See **fig. 8** for details. Space the resistors around a circle so there is an 118-inch or more between

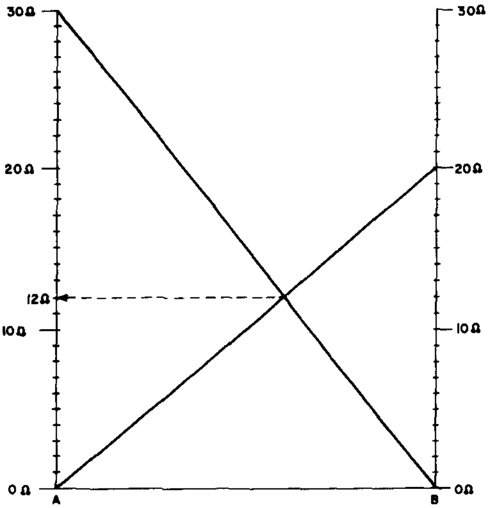


fig. 6. Nomograph for determining the equivalent parallel resistance of two parallel resistors. Example: what is the equivalent resistance when a 20- end 30-ohm resistor are connected in parallel? Plot a line from B on the right-hand scale to 30 on the left-hand scale and a line from A on the left to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

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

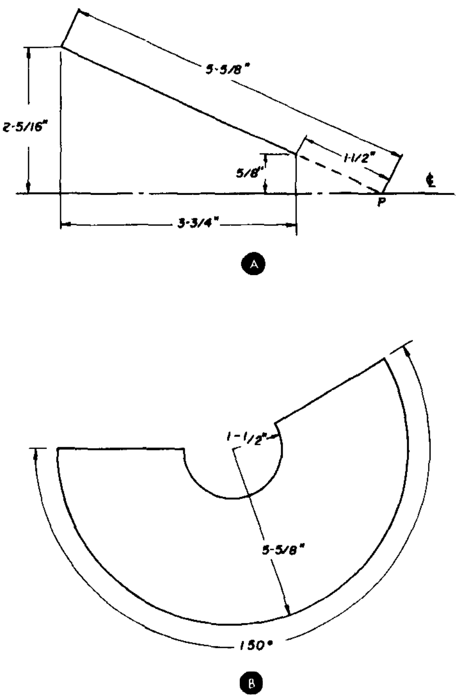
If your resistor banks are made up of unequal groups of resistance values, alternate the resistors around the circle. The conical group of wires connecting the ring nearest the connector are installed next. A dozen pieces of number 22 or *20* wire make an adequate cone. Get the apex of this cone of wires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

the coaxial shield

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

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

fig. 7. Laying out the outer shield.



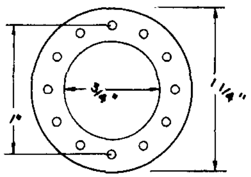
For oil dielectric, multiply the desired load impedance by $\sqrt{K}$, where K is the dielectric constant of the oil, to obtain the proper ratio of diameters using the conventional coaxial-line impedance formula. The dielectric constant of transformer oil is about 2.2 and the square root of 2.2 is 1.48.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

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

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

Regardless of the diameter of the resistor banks, you only have to multiply the resistor-bank diameter by 3.67 to obtain the correct screen diameter for a 52.7-ohm impedance in oil coolant. In this case, screen diameter will be about 4-5/8 inches.

fig. 8. Resistor mounting ring construction.



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

1. Find the circumference of the finished cone from circumference = π diameter = 3.14 × 4-5/8 = 14-1/2 inches

2. Determine the circumference of the outer circle you have drawn: 3.14 × 11-1/4 = 35 inches

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

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

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

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

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

calibration

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

Calibration can be done at 60 Hz by feeding the load with a variable transformer or a group of series-connected filament transformers. First, calculate a number of calibration points from $P = E^2/R$. With 12.6V applied to a 50-ohm load, for example, there is 3.2 watts into the load. After calculating a number of points, step through them with the variable transformer or other supply and record the meter readings. These points can be plotted on graph paper so the intermediate points can be read.

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

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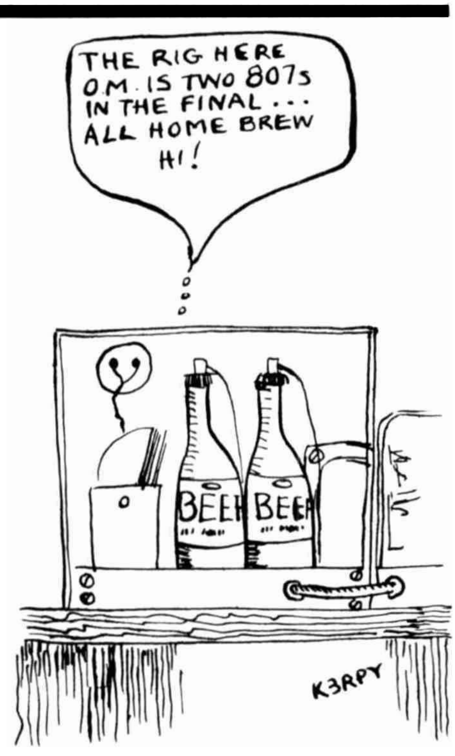


solid-state screen clamp

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

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

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

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

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

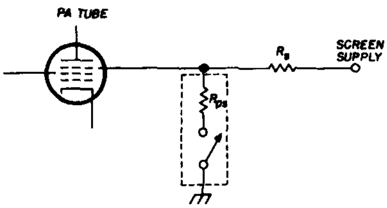
tube vs transistors

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

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

fig. 1. Clamp-tube equivalent.



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

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

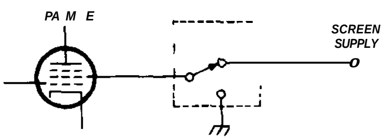
transistor circuit operation

The diagram of the solid-state screen clamp I use in my transmitter is shown in fig. 3. The basic principles of operation are quite simple. If excitation is present on the final amplifier grid, rf is coupled through the 470-pF capacitor to the 1N60 diode. The capacitor is charged during the positive half cycle of the excitation signal and puts a negative voltage on the base of the 2N3440, cutting it off. The collector voltage rises to a value determined by the 91k and 220k voltage divider.

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

fig. 2. Transistor clamp equivalent.



design considerations

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

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

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

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

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

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

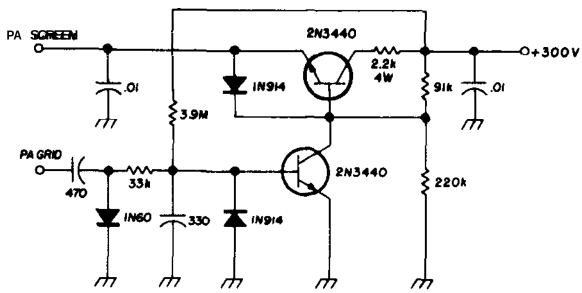
The design information in the preceding paragraphs is offered for those of you whose requirements differ greatly from the ones shown in fig. 3. If you have or are building a transmitter which has a low-voltage supply of about 300 volts and requires about 200 volts in the screen, the circuit may be used without changes.

operation

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

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

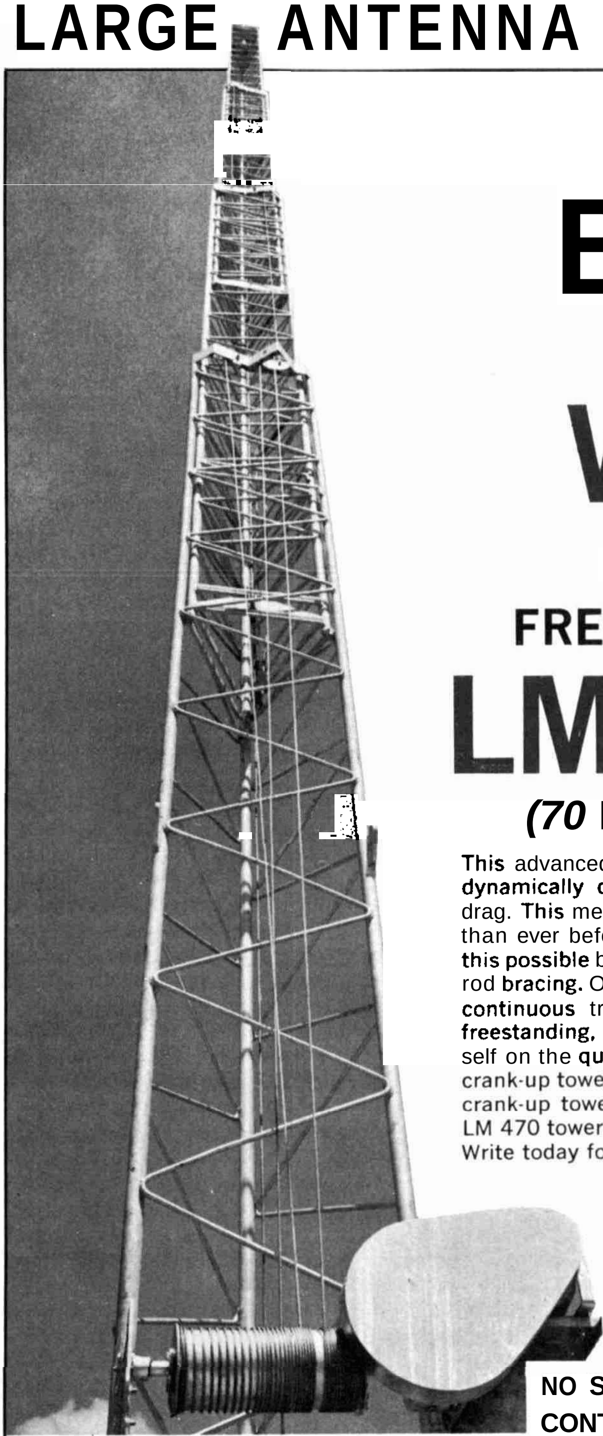
fig. 3. Schematic of the solid-state screen-clamp circuit.

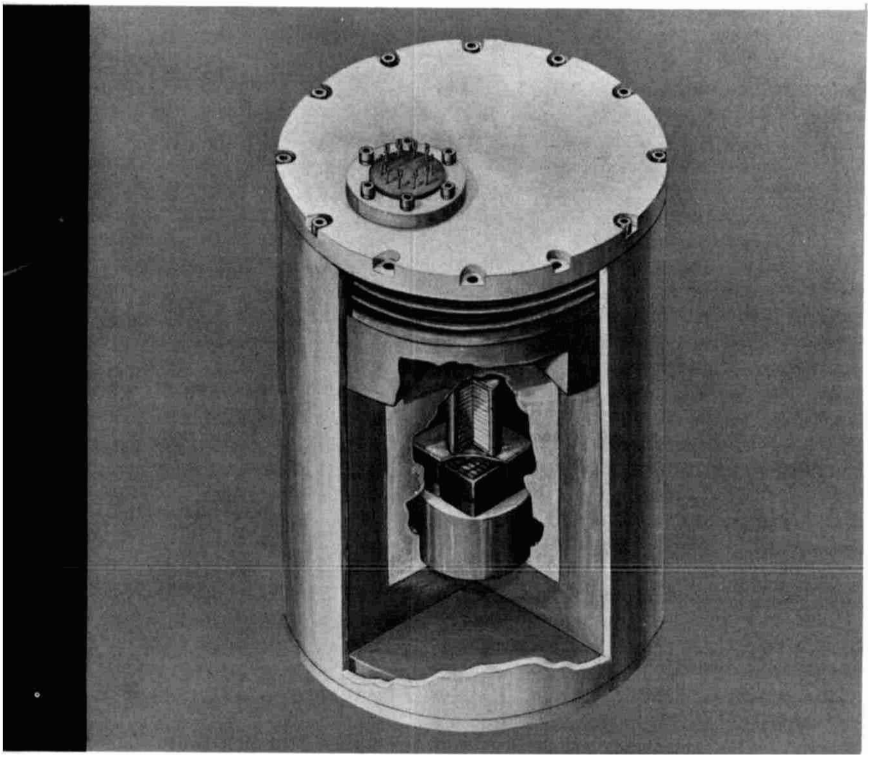


reducing the size of the 3.9M resistor. If the screen voltage does not rise with excitation, check for a low negative voltage at the base of the first transistor. If the voltage is present, it indicates the peak rectifier and filter are operating and that one of the transistors is probably defective.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

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

ham radio





a discussion of thermoelectric power supplies

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

Jesse Bryant, K1AJE, : Winnhaven Drive, Hudson, New H[redacted]hire 03051

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

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

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

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

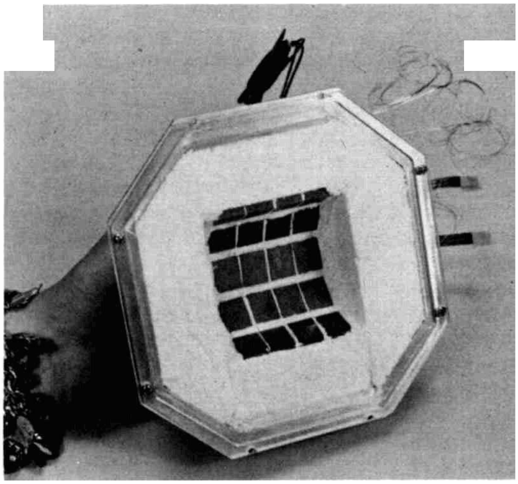
Waitl Before you stop reading, remember that although 72 watts is not much, it's enough to charge a ballast battery (preferably nickel-cadmium for long-life and low maintenance) during non-transmit periods. A 10-ampere-hour, nickel-cadium, 12-volt battery could supply a 120-watt steady-state load for one hour—but our loads are only intermittent. With properly designed solid-state transmitting and receiving equipment, the peak-power duty cycle is very low. The 12-watt thermoelectric generator would continue to supply power day and night for charging the ballast batteries—they would permit a relatively high-power drain during periods of peak power requirements.

theory

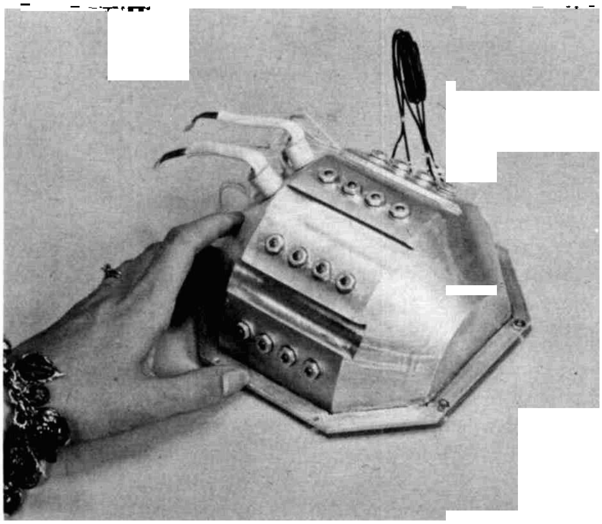
Direct conversion of heat into electricity has been with us for over 145 years. In 1822, Johann Seebeck reported he had observed that a magnetic needle was deflected when placed near a closed loop of two dissimilar metals when a temperature difference was maintained between the two junctions. A very basic thermoelectric circuit is shown in fig. 1. One end is hot and the other end is immersed in an ice bath; by using dissimilar metal wire in a closed loop, a small voltage is developed. With the exception of its usefulness in measuring temperature, Seebeck's discovery went rather unnoticed. In fact, you can buy a complete set of temperature vs emf tables for various dissimilar metal thermocouples from the U. S. Government Printing Office.

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

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



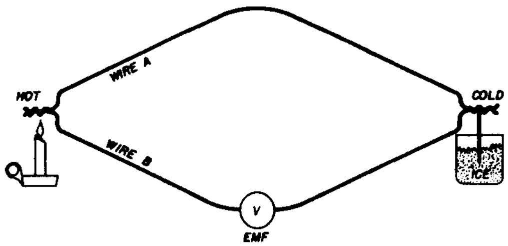
Commercial 1-watt thermoelectric generator. Total **weight** is 4.8 pounds; volume. 0.07 cubic **feet.**



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

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

fig. 1. The basic thermoelectric circuit.



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

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

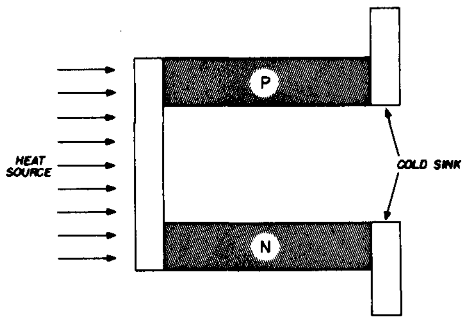
Some of the metals now used are silicon-germanium, bismuth-telluride and lead-telluride. They are usually selected according to operating temperature, ease of fabrication and cost per unit. As the state of the art improves in metallurgy, the temperature limit on the hot end is reaching 2000° F and above.

radioisotopic thermoelectric generators

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

With the present state of the art, the efficiency of thermoelectric generators is approaching 10%. If a generator output of 10 watts (thermal) is required at the end of 5 years (half-life of fuel), assuming 10% efficiency, the initial fuel load for the generator must be 200 thermal watts. At the beginning of life, 100 watts of thermal heat must be dissipated. This is wasteful but necessary if the rated power of 10 electrical watts is to be maintained at the end of 5 years. This is called power flattening or dampening.

fig. 2. Typical thermoelectric module.



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

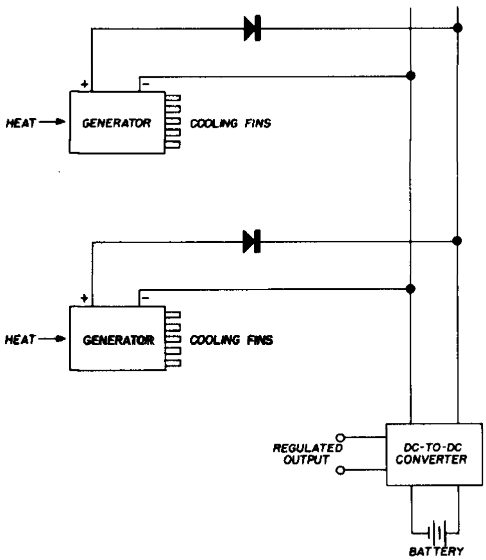
The high cost of efficient generators and the tremendous cost of the long-lived fuel, Plutonium-238, which is controlled by the Atomic Energy Commission, leads to caution when considering the use of isotopic generators. However, they can be used to charge batteries (made to last 5 years or more) which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

The battery pack will supply tremendous power for a short time and is rechargeable with a thermoelectric generator. At the end of the sensed battery life, the batteries may be disconnected remotely; low power would still be furnished by the thermoelectric generator until the end of several half-lives of the radioisotope fuel. For the Moonray project the thermoelectric generator has another benefit: when the cold lunar night sets in, the heat from the radioisotope fuel could keep the batteries from freezing.

For less sophisticated applications there is another breed of generators called terrestial generators. These are for remote applications and use cheaper fuel and abundant shielding. They can be placed anywhere. Automatic remote repeaters and microwave relay stations could operate using batteries that are recharged by the generator. A typical low-power generator is shown in the cutaway drawing.

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

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



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

In most thermoelectric generators, the voltage is very low—3 to 5 volts output. This is almost useless unless it's converted to a higher output voltage. Dc-to-dc converters are used to step up the voltage to a useable value such as 6, 12, 24 or 28 volts using highly efficient toroid-core transformers and solid-state circuitry.

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

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

other uses for thermoelectrics

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

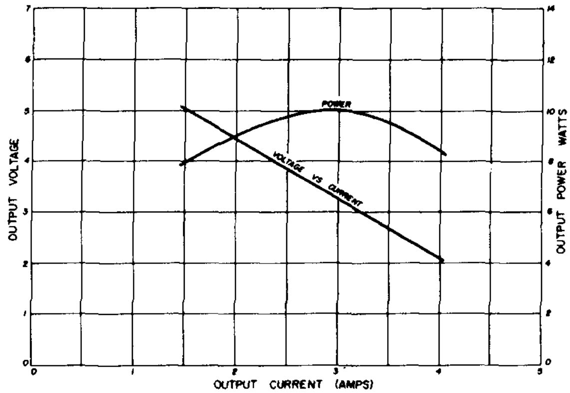
table 1. Typical half lives and power density of several radioactive elements that show promise in the thermoelectric field.

isotope	half life years	power density thermal watts/in³
Thulium 170	0.3	100
Thulium 171	1.9	10
Strontium 90	28.0	6
Plutonium 238	89.0	25

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

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

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



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

references

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

ham radio

"You wouldn't think it was so funny if your husband was building his first ham rig!"

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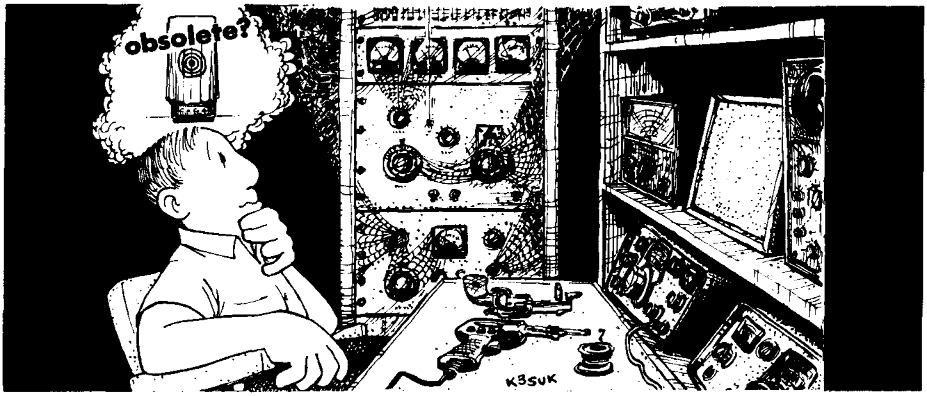


don't throw away that iron!

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

Bill Blankenship, WA4GNW/W1RDR 6709 Green Mead Court, Lou'svlle, Kentucky 40207

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

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

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

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

power supplies

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

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

amplifier modifications

I installed a small relay adjacent to the bias supply and connected the cut-off bias voltage to one set of relay contacts. A second slider tap on the bias supply resistor was positioned so the final draws the correct no-signal idling plate current when plate voltage is applied; this bias is connected to the normally-open relay contact. The armature is connected to a feedthrough capacitor which feeds the bias to the shielded grid circuit. The relay coil is connected to an "external relay" socket on the exciter. When receiving, the final is biased to cut-off, and when the exciter's PTT-VOX relay is closed, ssb operating bias is applied to the amplifier.

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

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

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

operation

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

In regulating the amount of drive to the amplifier, the exciter should be tuned for normal maximum output. If this produces excessive drive to the amplifier, the exciter drive control can be reduced slightly. However, make sure that the drive to the exciter's own final is not reduced below the manufacturer's specifications for minimum drive. Otherwise, the alc circuit and carrier suppression will be upset and signal quality seriously affected. If minimum permissible output from the exciter still results in excessive drive, use an attenuator.

construction

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

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

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

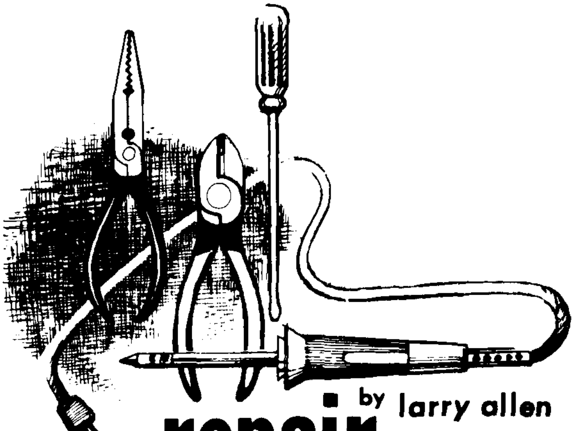
cooling

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

The top of the chassis forms the floor of the plate-circuit compartment with twenty-four '14-inch holes drilled around the periphery of each tube socket. Mounted on the rear edge of the chassis are two 4-inch centrifugal blowers rated at 250 cfm each with zero back pressure. Since the total area of the drilled holes is considerably less than the combined area of the two blower nozzles, there's a certain amount of back pressure, but this is inconsequential. Blower noise is at a very tolerable level and do a no-nonsense cooling job.

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

ham radio



the repair bench

by larry allen

aligning vhf transmitters

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

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

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

For oscillator alignment, you should plug in the highest-frequency crystal you have. This permits you to set the plate coil for best stability, since any instability of oscillator-tripler operation shows up first at the higher frequencies.

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

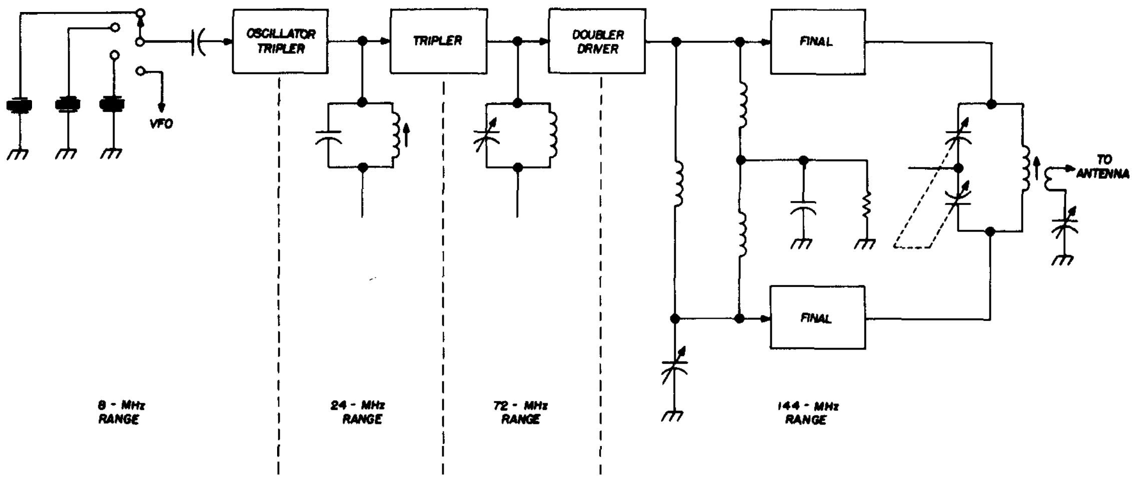
For the later stages, such as the driver and the final, it is better to monitor rf output. In vhf power stages, the plate-current dip dosen't usually occur exactly at the most efficient point. Therefore, for all output-stage adjustments, an rf wattmeter or other rf output meter is the most effective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

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

The order in which you make the final-stage and output-coupling adjustments can be a factor in vhf transmitters. Start with coupling as loose as possible (least rf output). If there is a plate-current meter, use it to dip the power-amplifier tank as a preliminary step; later, however, you'll retune the tank for maximum rf output. Drive in vhf transmitters is seldom adjustable; you just tune the driver stage for strongest drive to the final.

Start by increasing antenna coupling just enough to get some output reading on the rf wattmeter or whatever you're using as an output indicator. Ignore final plate current at this point. Now tune the plate tank for maximum rf output. Then peak the antenna trimmer, if there is one. In small steps, increase rf output to the rated wattage for the transmitter, repeating the sequence just outlined: increase coupling, tune plate, tune antenna. Don't allow plate current to exceed whatever is recommended for the tube; also, don't exceed the FCC-ordained plate power input. Don't tune the plate tank for a dip on the meter—tune for output, always. If plate current gets too high, back off the coupling so there's less rf output; then retune the plate and antenna controls for peak output.

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



a-m vhf transmitter

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

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

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

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

Output coupling systems for each band are separate and different. Tune the 6-meter band first, which uses the pi-network output system. The standard alignment procedure applies, except that there is no coupling adjustment. Tune the plate trimmer (C1) for maximum rf into the dummy load. Then adjust the antenna-tuning trimmer (C2). Peak them both, going back and forth a couple of times to make sure interaction doesn't affect the result. You'll have to refine them both a little when you connect an antenna.

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

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

single-sideband vhf

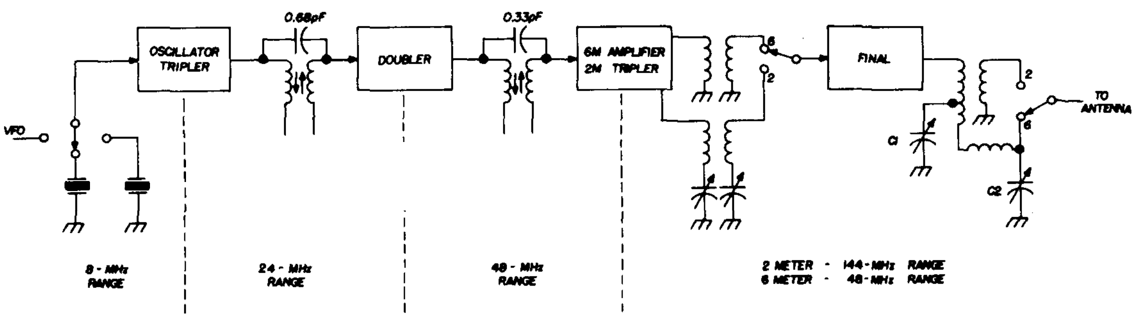
There are some differences between vhf transmitters for a-m only and those that operate in the ssb mode. The variations aren't major, but they are sufficient to alter the general alignment procedure. The block-schematic of **fig.** 3 shows the layout of one double-conversion vhf ssb transmitter—from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

This transmitter uses frequency synthesis to tune the 6-meter band in sectors. Four 35-MHz crystals handle this. The tuning oscillator lets you use crystals from 5.5 to 6.5 MHz to select spot frequencies within the 1-MHz spread of each sector; with the VFO, which is part of the transmitter, you can tune any frequency within the sector. The carrier oscillator is the 9-MHz type that is common in modern transceivers. It is this synthesis method of developing the vhf output frequency that dictates a special sequence for transmitter alignment steps. The procedure isn't as straightforward as in simpler a-m units with frequency multipliers.

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

The first step is to align the 9-MHz section. Disable the tuning oscillator or VFO. Connect a vtvm with rf demodulator to the output of the first mixer (the transistor collector, in this instance). Key the transmitter on, with its mode switch set for CW operation. Peak T10 for maximum. Move the vtvm probe to the 9-MHz input of the mixer (emitter, here). Adjust the crystal trimming capacitor (C23) for exactly 65 mV on the vtvm. Move the vtvm probe back to the mixer collector again, and again peak T10. The trimmer capacitor puts the crystal signal at exactly the right spot on the response slope of the sideband filter, and you have peaked T10 to

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



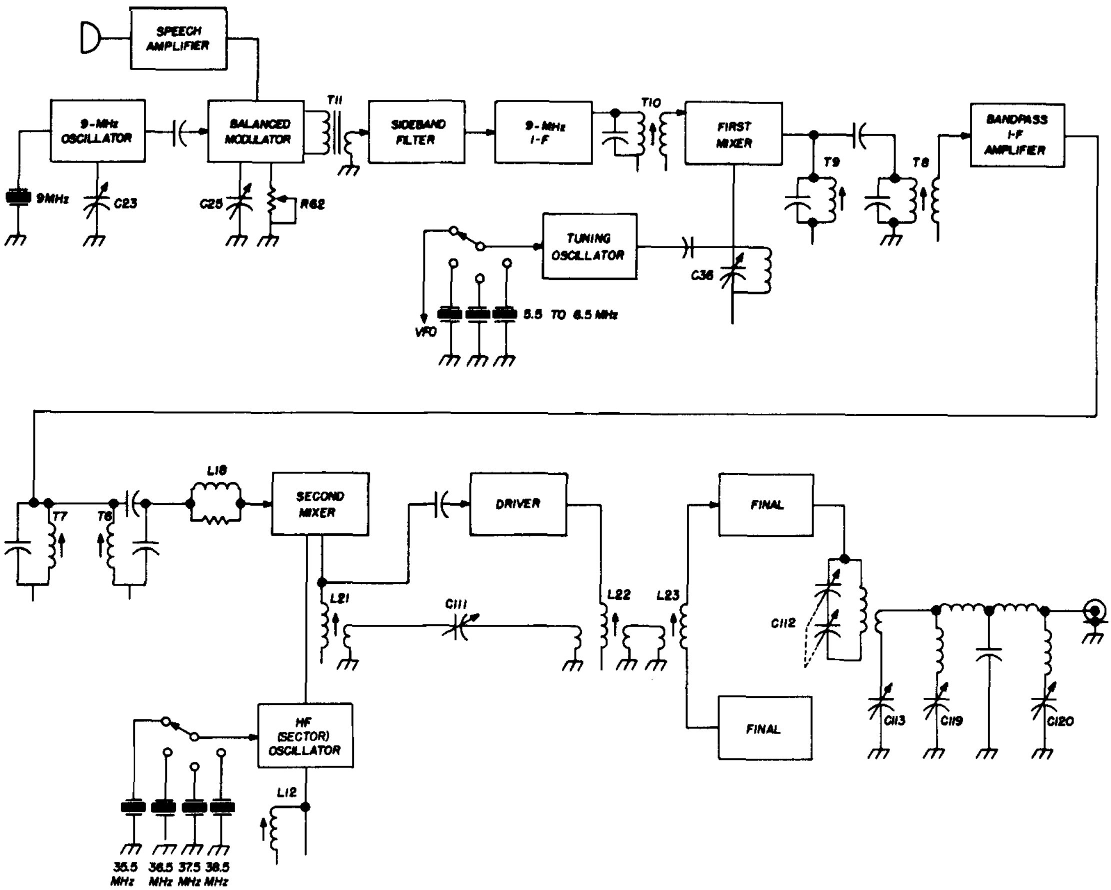


fig. 3. Adjustments and their sequence are complicated in this vhf single-sideband transmitter. Spacial equipment is recommended for adjusting some of the tuned band-pass circuits.

pass the resulting 9-MHz signal efficiently.

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

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

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

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

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

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

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

antenna adjustments at vhf

Any vhf transmitter should be connected to its working antenna for finishing adjustments in the final amp and output network. Antenna matching is critically important in vhf operation, and the adjustments in the final stages can seriously affect this matching.

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

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

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

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

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

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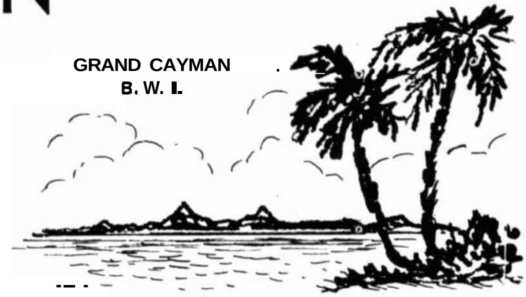

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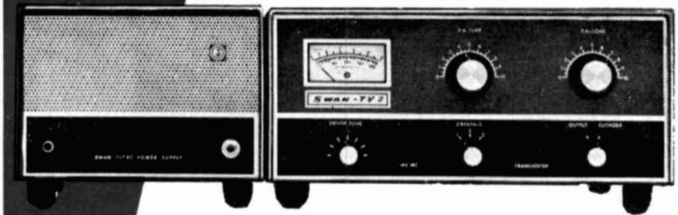

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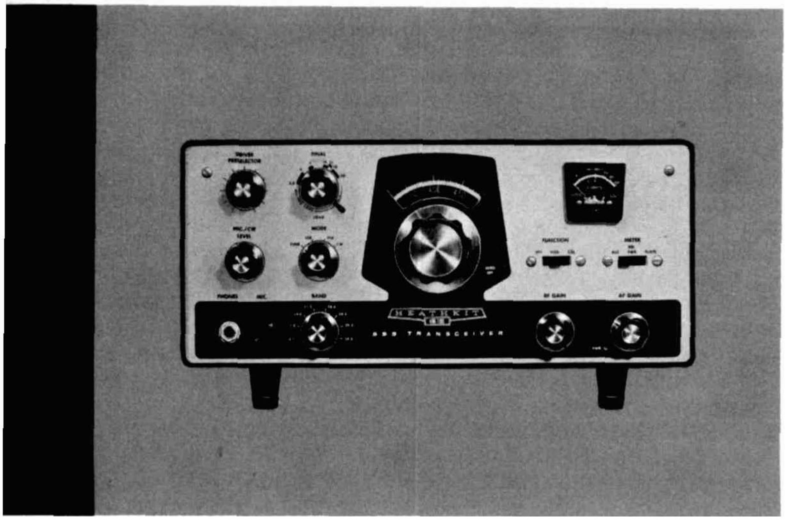


the new heath HW-100 five-band ssb transceiver

Skip mey, V1N L& Fay on, n '3

One of the more intriguing new pieces of equipment to be placed on the market in recent months is the Heath HW-100—a five-band ssb transceiver selling in kit form for only $240.001 This certainly bears looking into, and a more careful review uncovers a very interesting piece of gear.

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

When the kit arrived here at ham radio, we looked it over and were all quite impressed to find that the folks at Heath hadn't cut any corners quality-wise to give you a dependable piece of equipment. I'd certainly hate to try and put a similar group of parts on my bench for the same price. Among the branded components were many well-known names including Westinghouse, E. F. Johnson and Potter & Brumfield.

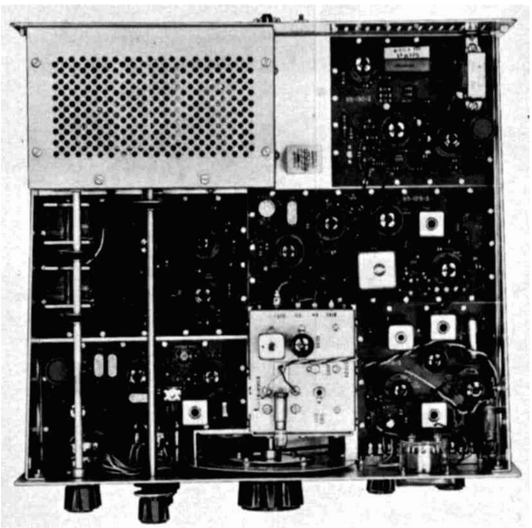
assembly

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

Most of the wiring was handled by nine circuit boards; a large wiring harness provided most of the interconnections between the boards, controls and other points. The only two major portions of the transceiver which are hand wired are the VFO and the final amplifier. Both of these were quite straightforward and went together very nicely. The manufacturer was very careful to provide ample warning regarding any operations requiring special care such as the use of heat sinks on semiconductor leads, etc.

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

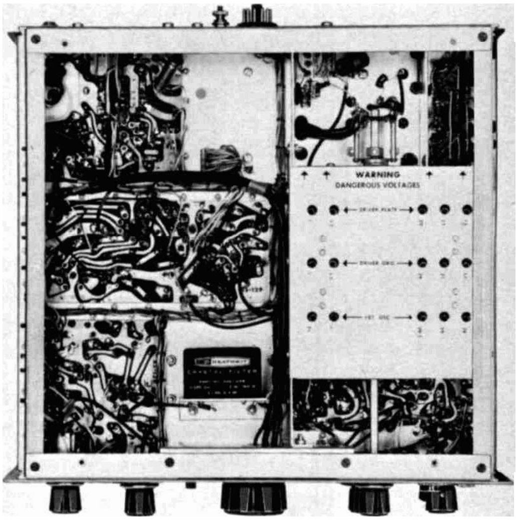
One point of interest which didn't affect our kit, but did trouble a neighbor, was the initial tune-up of the transmitter. The instructions assume that T1 at the output of the balanced modulator will be sufficiently close to alignment as shipped from the factory so some drive will be available before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



receiver section

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

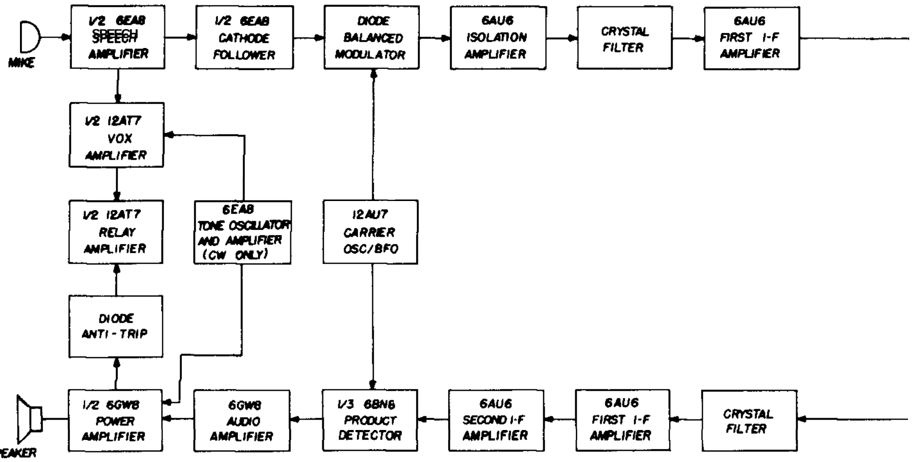
The incoming signal is first passed through a single stage rf amplifier/preselector and on to a crystal-controlled first mixer. The first intermediate frequency is tunable from 8.395 MHz to 8.895 MHz. The second mixer converts this signal to 3.395 mHz and uses the VFO/VFO-amplifier from the transmitter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



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

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

The operation of the receiver is identical fig. 1. Block diagram of the HW-100.



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

transmitting section

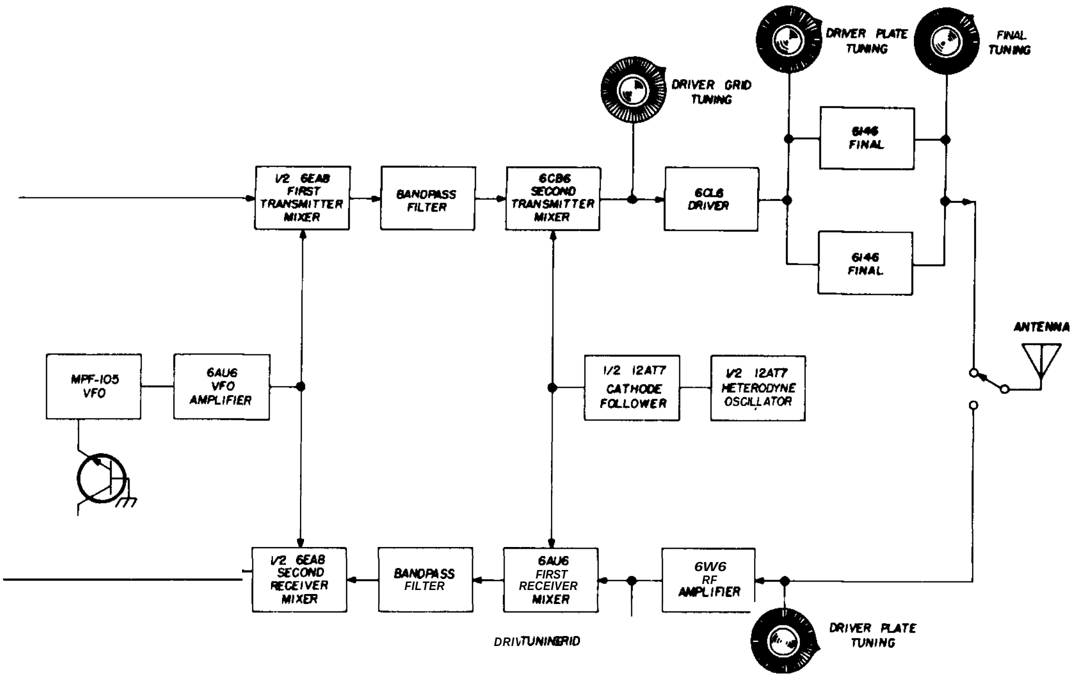
The transmitting section of the HW-100 is a typical example of a modern filter-type multiband ssb transmitter in the 150-watt class. In ssb operation, the audio signal, after passing through a speech amplifier and cathode follower, is mixed with the output from the carrier-oscillator/BFO in a four-diode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

The same four-pole crystal filter is then used to both attenuate the undesired sideband and further suppress the carrier. For CW or tune-up operation, the frequency of the carrier oscillator is shifted to place it within the pass band of the filter. This filter does an adequate job of sideband suppression. Although in running tests of my rig I have been able to copy the undesired sideband, it has been too weak to make any actual power measurements of it. I feel sure that the manufacturer's specifications of 45-dB suppression are being met and that under normal band conditions this sideband would never be detected.

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

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

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



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

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

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

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

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

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

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

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

harmonic drive mechanism

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

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

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

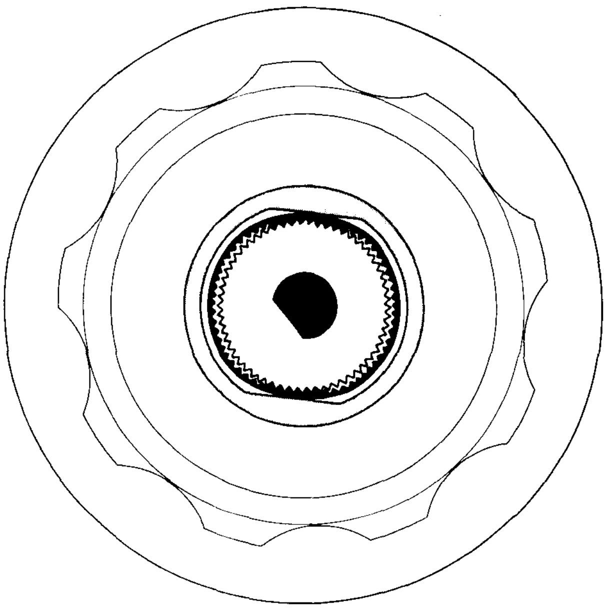
When this dial is first assembled, it may seem unsatisfactory. The "stiction" or breakaway force required each time you start to turn it may be annoying. **Don't worry!** With proper lubrication as outlined in the directions, and a bit of use, this dial rapidly becomes a real beauty. I have been well pleased with mine.

operation

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

In summary, I can say that the HW-100

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



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

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propagation predictions for september

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

Victor Frank, WB6KAP, 12450 Skyline Boulevard, Woodside, Californ'a 94062

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

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

predicting the muf

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

With this in mind, I scaled a number of vertical incidence ionograms taken at Point Arguello, California, (35.5" North latitude) during September 1967 to determine the MUF for the east-west path with a control point at 35.5" North latitude. The transformation from vertical to oblique incidence propagation is not without its problems, and the MUF's derived this way are invariably low. This is due to a number of causes. However, the MUF error for east-west F2-layer propagation in temperate latitudes is usually less than 8%.

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

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

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

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

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

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

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

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

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

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

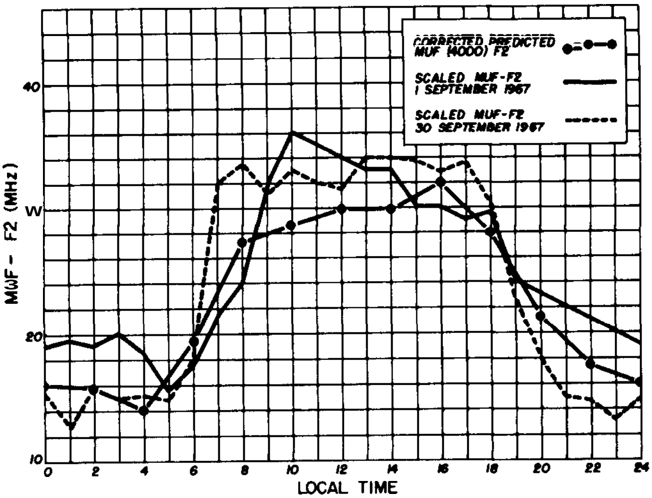


fig. 1. Maximum usable frequencies (F2 layer) scaled from vertical incidence ionograms taken at Point Arguello, California (latitude 35.5° N) on September 30, 1967 as compared with corrected predictions (see text).

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

Within 2000 miles of the magnetic equator (between 8° and 14° South latitude over South America), transequatorial forward scatter (TE) may extend the operational MUF to almost twice that obtained by ordinary refraction during the evening hours near the equinoxes. During disturbed conditions (usually early in the disturbance), TE may be worked at higher latitudes than normal. It is not known whether this extension is due to sporadic-E or a distortion of the normal F2-layer.

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

If the predicted variation of MUF with latitude holds (MUF's increase proportionately), then you could expect the daytime MUF's to be 13% higher at 30° N, 30% higher at 25° N, and 43% higher at 20° N than those at 35.5" N. In addition, if the MUF's are 5% higher this year than last, then you could expect 50 MHz to open for a control-point latitude of 30° N at least once during the month, to 25° N at least five days during the month, and to 20° N at least ten days during the month.

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

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

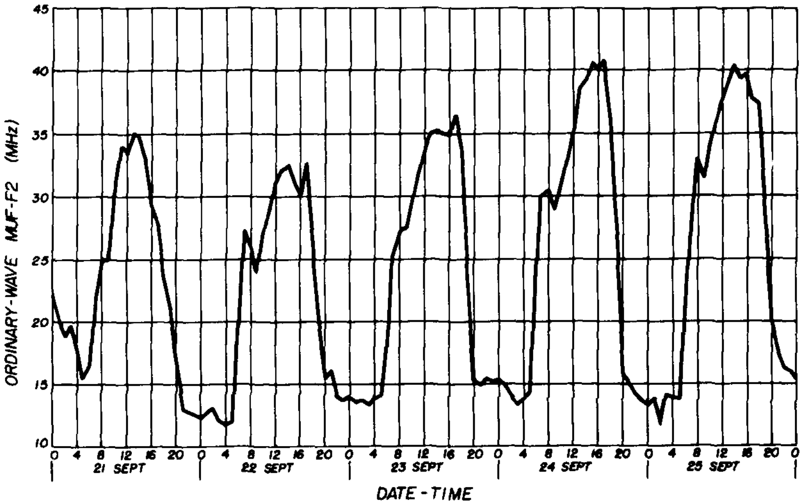


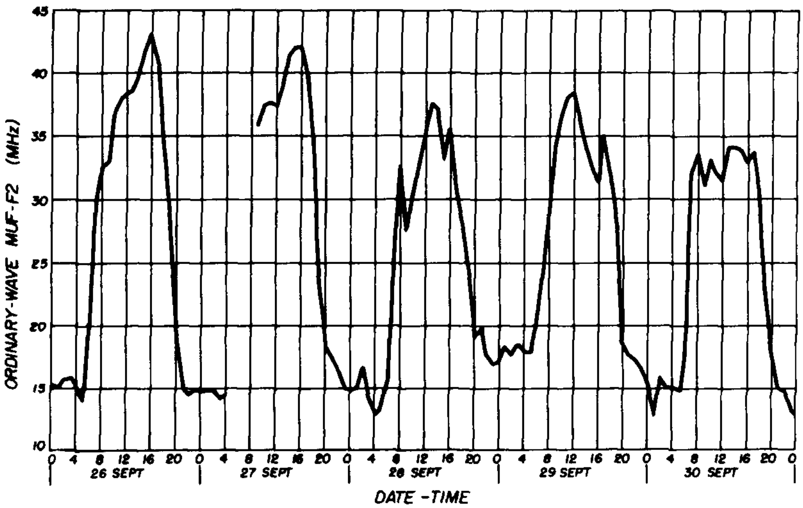
fig. 2. Maximum usable frequencies (F2 layer) scaled from vertical incidence ionograms taken at Point Arguello, California between September 1 and September 30, 1967.

maximum usable frequency

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

The time and latitude are those of your control point, 1250 miles away from your station in the direction of propagation. If only one control point is considered, all that is guaranteed is that a frequency below the MUF will propagate to a distance of 2500 miles during half the days of the month. A signal at this frequency may, and probably will, propagate much further.

The time chart can be treated as a contour



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

maximum range

Time charts of maximum range determined by absorption and atmospheric noise levels are shown in **fig. 4** to **fig. 8.** These charts are based on an output power of 100 watts CW or 800 watts ssb with combined receiving and transmitting antenna gains compared to an isotropic radiator of –12 dB for 80 meters, 0 dB for 40 meters and 12 dB for 20 meters.

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

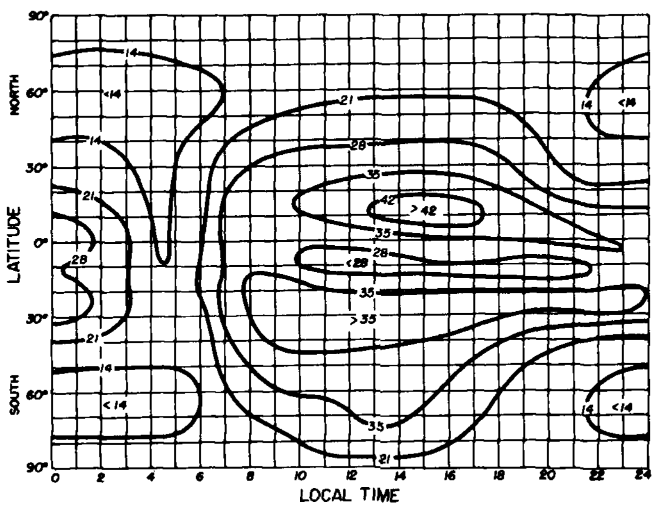


fig. 3. Time chart of predicted median MUF is for September 1968 for the American continent.

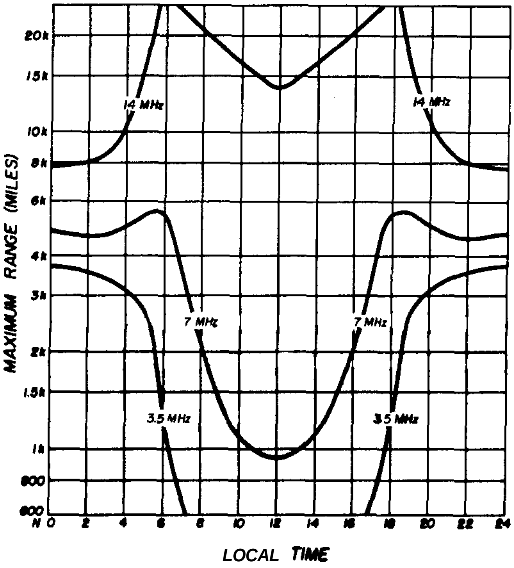


fig. 4. Maximum range due to absorption and noise vs local time from 38° N latitude to the **north.**

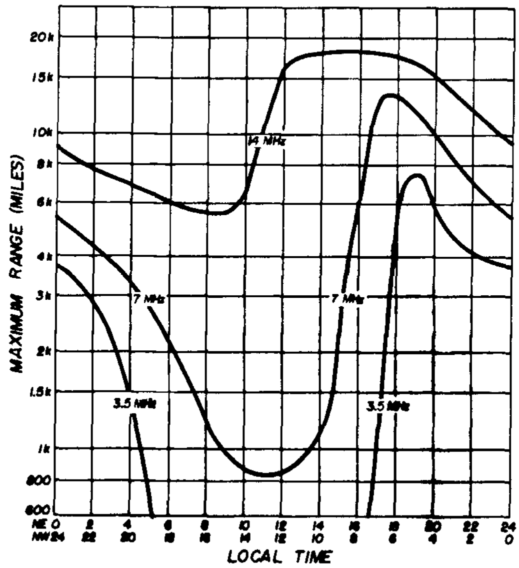


fig. 5. Maximum range due to absorption and noise **vs** local time from 38° N latitude to the northeast (top time scale) and to the northwest (bottom time scale).

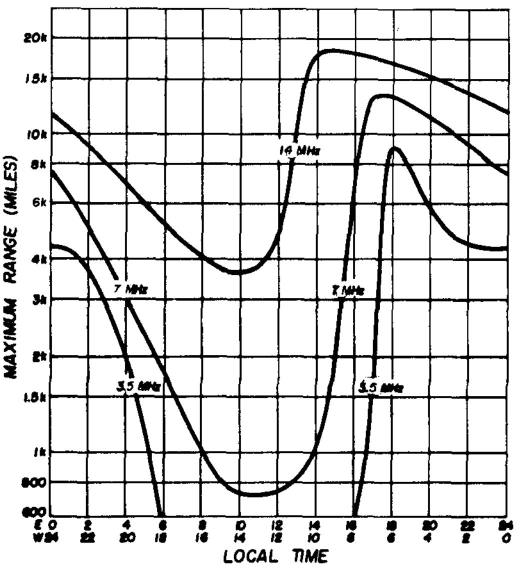


fig. 6. Maximum range due to absorption end noise vs local time from 38° N latitude to the east (top time scale) and to the **west** (lower time scale).

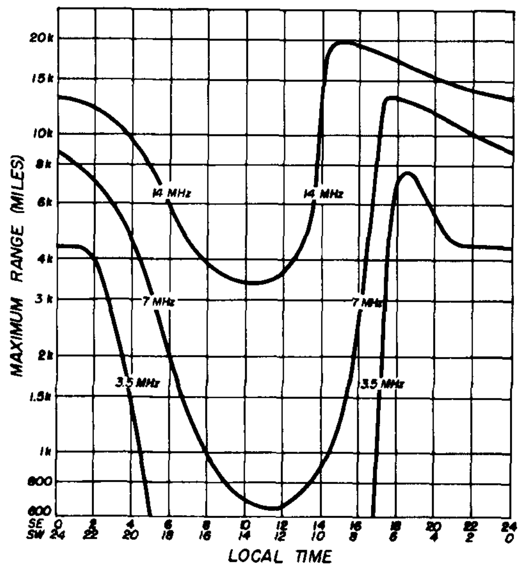


fig. 7. Maximum range due to absorption and noise vs local time from 38° N latitude to the **southeast** (top time scale) and to the southwest (bottom time scale).

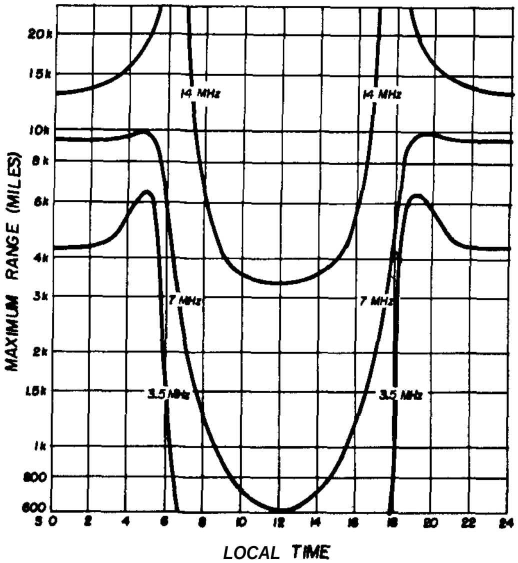


fig. 8. Maximum range due to absorption and noise vs local time from 38° latitude to the south.

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

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

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

propagation summary for september

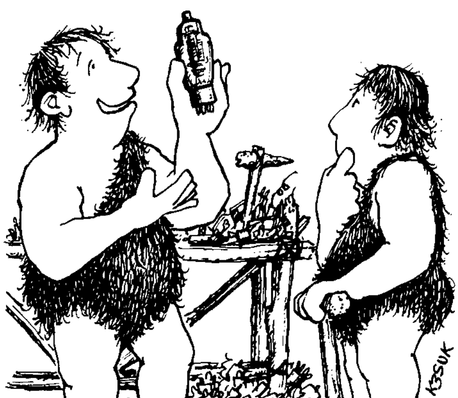
80 and 40 meters. On these two bands, decreasing noise levels and increasing hours of darkness will improve DX conditions in the Northern Hemisphere.

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

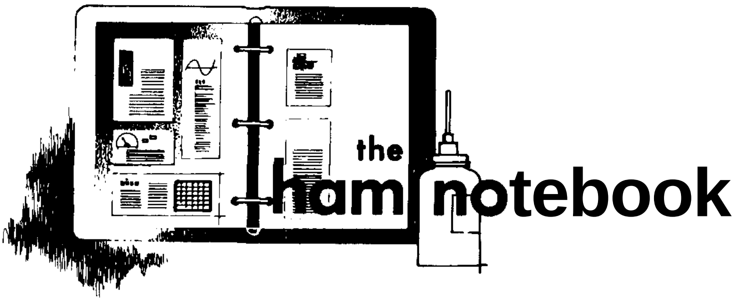
15 meters. Fifteen will open and close earlier as the month progresses. Polar paths should be workable from most of the United States during hours of daylight (at both ends of the path) most of the days of the month.

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

ham radio



"After 806 failures—success!"



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 untieing 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

using industrial cartridge fuses

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

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

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

D. E. Hausman, VE3BUUE

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 gauage (.062") or larger, the assembly can be used as a third hand when soldering.

Tony Felese, W2KID

makeshift test equipment

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

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

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

Keith Olson, W7FS

vhf antenna switching without relays

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

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

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

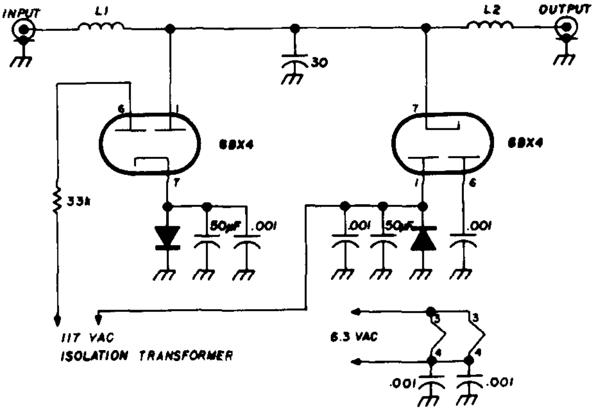
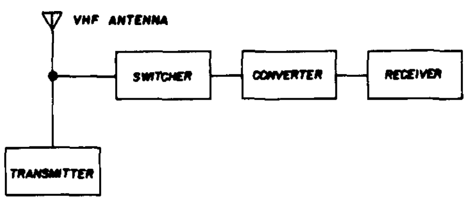


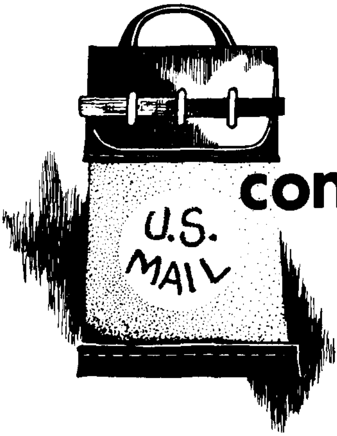
fig. 2. Block diagram showing how the vhf switcher is installed.



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

Some trial-and-error may be required with coils L1 and L2 to obtain optimum performance and accurate low-level triggering. I found that for 220 MHz, both coils can be a 314-turn of number 18 wire, 3/4" diameter. At two meters, 1-3/4 turns of the same provided best results. For six meters, 6-1/2 turns of number 18 enameled wire on a 3/4" diameter form did the trick.

Bob Brown, K2ZSQ/W9HBF



comments

Dear HR:

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

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

Bill Santiff, WA2QKT

Dear HR:

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

Ted Cohen, W9VZL/4

Dear HR:

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

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

Lawrence W. Banks, W1DYJ
Massachusetts Institute of Technology

Dear HR:

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

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

Your magazine is great, keep it up.

Bob Jones, KH6AD

From pages 52 and 54 of "The Telegraph Instructor," by C. M. Dodge, Valparaiso, Indiana, copyright 7908:

Ham—A telegraph operator who is not proficient.

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

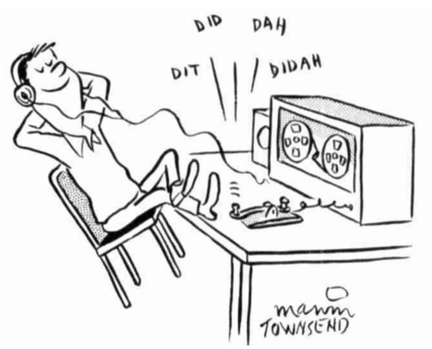
In the same issue of QST, *Paul Codley, Ex-2EZ, wrote that in 1907 he had asked an old-time wireless operator with whom he was working why a particular operator was called a ham. The old timer replied, "He's got a ham for a 'fist'!"*

Editor

Dear HR:

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

Bill Dickson, K4BQ
1914 call—BK



new products

dictionary of electronic terms

The field of electronics continues to expand at a rate that has never been equalled in the arts and sciences in the long history of man's pursuit for knowledge. Of necessity, the language of electronics has kept pace to cover semiconductors and integrated circuits, space and nuclear technology, computers and automation, and audio and video recording. The new edition of the **Dictionary of Electronic Terms** has been enlarged and revised to include new terms and to clarify old ones with accepted and up-to-date usage.

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

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

applications note catalog

Motorola has just published a new Applications Note Catalog with the title and a brief summary of more than 130 technical write-ups that describe semiconductor design. A selector guide in the front of the 11-page catalog lists the notes by application categories for easy reference. Also included in the index is a partial list of selection and cross-reference guides, as well as full-line catalogs, available from Motorola. For your copy of this handy guide, the Motorola Application Note Catalog, write on your company letterhead to Department TIC, Motorola Semiconductor Products, Inc., Box 20924, Phoenix, Arizona 85036.

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New 40- and 60-volt npn transistors recently introduced by Motorola can be paired with 30-ampere 2N4398-99 pnp devices to furnish the highest silicon power transistor complementary symmetry capability ever possible. Since the pnp-npn complementary symmetry design approach can now be used in high-current applications, they are ideal for cost-cutting circuit-simplifying applications in audio and servo amplifiers. The pair also furnishes a higher degree of frequency stability in both ac- and dc-driven loads without the addition of expensive impedance-matching driver transformers.

Both the new 2N5301-2 npn series and the 2N398-99 pnp devices can handle 200 watts of power to 30 volts V_{CE}. They can be used as plug-in replacements of germanium types in difficult amplifier, voltage-regulator and modulator designs. Also, because of the low thermal resistance of these devices, simpler and less-costly heat sinking can be used. Other advantages of these new transistors include excellent gain linearity over a 5- to 20-ampere range, 50-A peak current and TO-3 packaging: f_T of these devices is 4 MHz. For more information, write on your letterhead to Motorola Semiconductor Products, Inc., P.O. Box 955, Phoenix, Arizona 85001.

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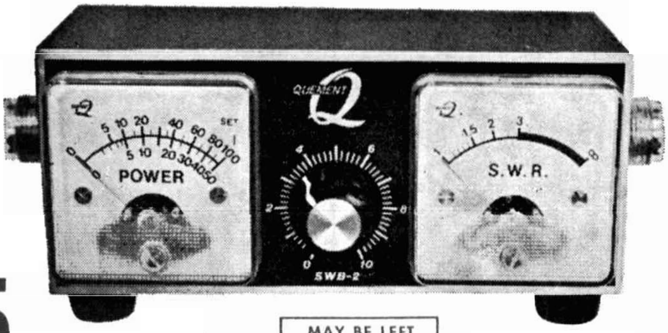

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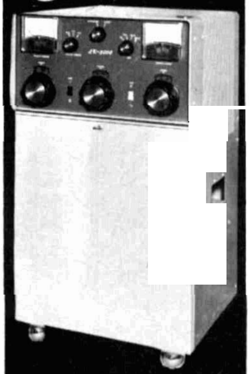

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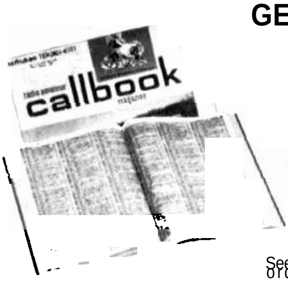

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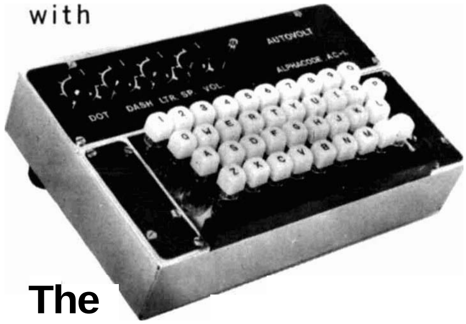

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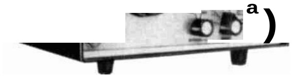

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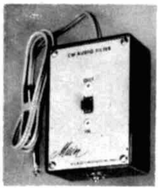

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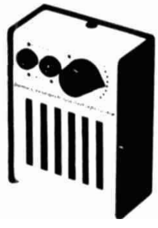

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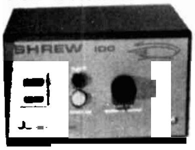

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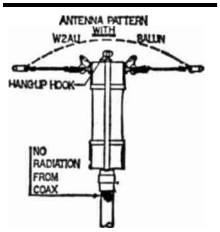

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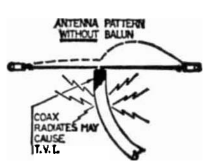

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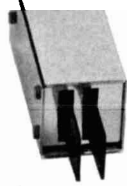

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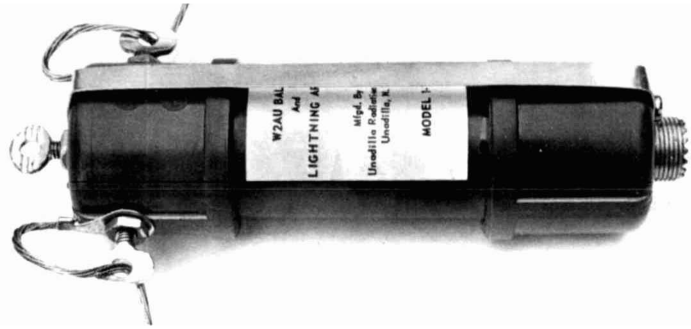

WHEN YOU WERE A KID

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

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

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

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

In my preceding ads I have repeatedly stressed the need for low VSWR. High reflected power causes more anguish and frustration than anything else. By balancing your antenna system, you may lower your VSWR, but more important is the fact that with the balun, your radiation efficiency is often higher. Normally, the grounded chassis side of the coax connects to one side of the dipole, and this side has very little induced electro-magnetic radiation from it (the dipole isn't a particularly good transformer). Thus a balun, and especially a well-designed product like the 2AU, accepts the unbalanced input from the coax and permits equal distribution, and consequently equal electromagnetic radiation, from both halves of your antenna.

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

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

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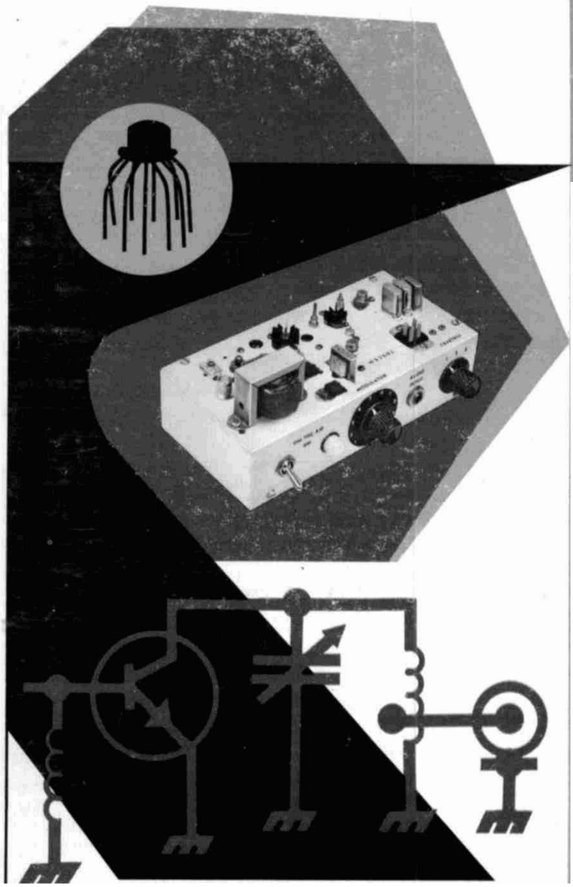
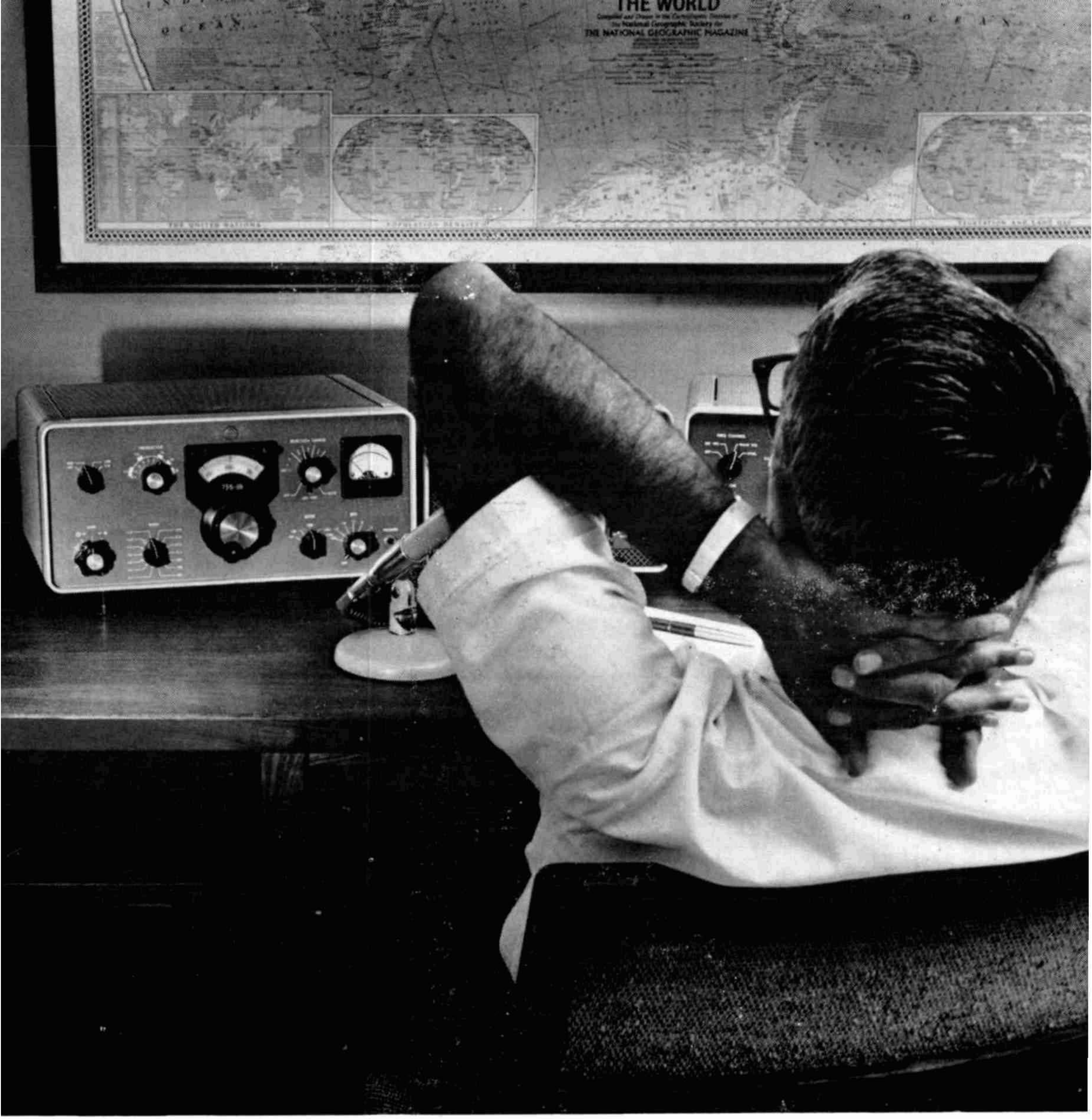


this month

solid-state transmitter for 6 meters



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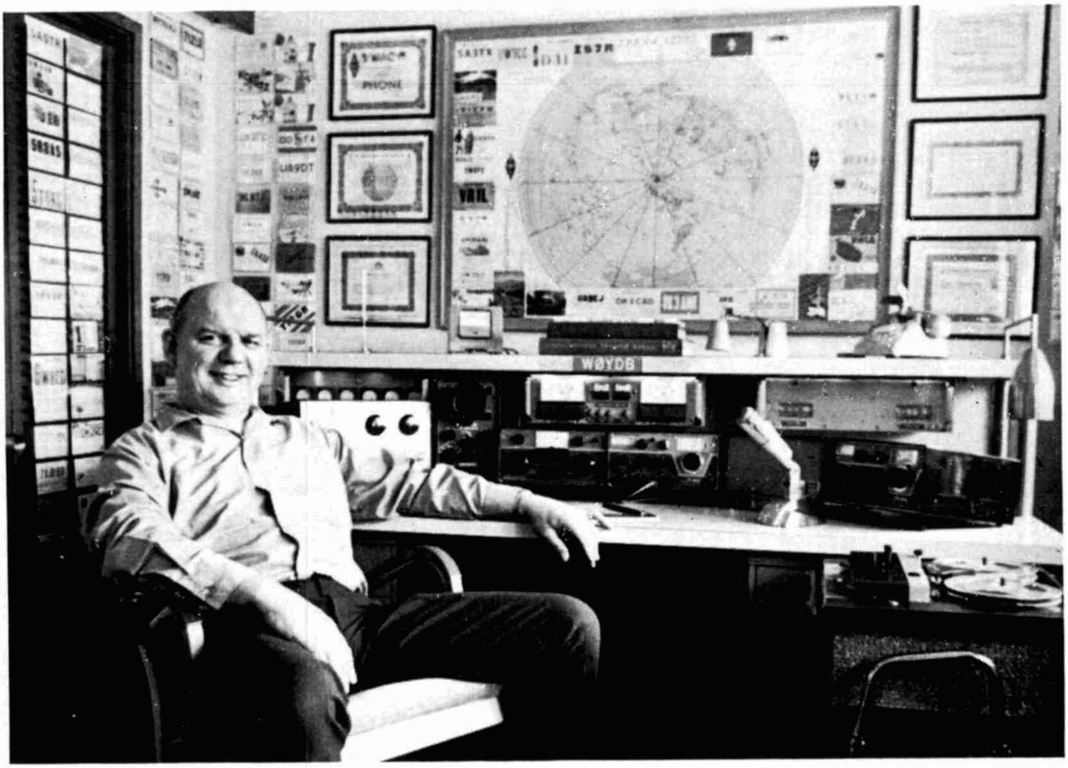
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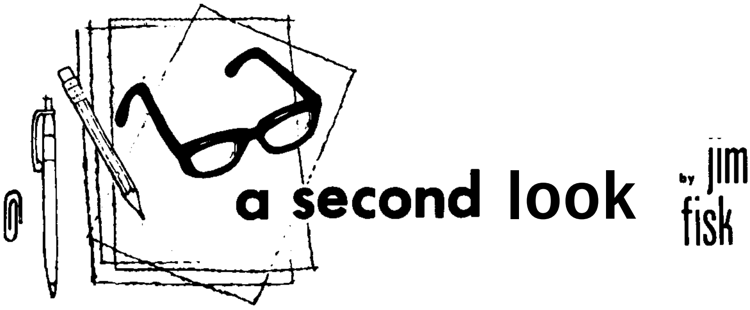

If you've been watching WB6KAP's monthly propagation reports, you're probably aware of rapidly Improving DX conditions. During the summer, propagation was extraordinary, with 20 meters open almost any time of the day or night. Conditions on 15 were great in September and the current 10-meter openings offer a lot of DX catches with a flick of the dial. If you're interested in making DXCC, you should be able to do it in a couple of weekends—if you listen.

It's most amusing to scan up and down the 20-meter band looking for rare ones, noting all the W/K's calling "CQ-DX" and picking up FB8WW in Crozet, TJ1AJ in the Cameroons and 9H1AG in Malta in between! It's more frustrating than amusing when you hear a strong W/K station calling "CQ DX" on top of JT1KAA, KR8EA or UJ8AI. The point is, if you want to work the rare ones, you've got to listen—listen, listen and listen some more.

You can work some pretty good DX by calling CQ if you have a powerful signal, live in a rare state or are well known, but you'll improve your country total a 1000 times faster by listening more. I know you're not all interested in working a new one, but even if you're interested in chewing the fat, you're not going to do it by calling a dozen times and signing your call once. You'll wear out your mike, your key and your final, but you won't put very many entries in your log. More power isn't the answer either—good operating is the only thing that will do the job.

I've noticed a lot of activity on the CW portions of the bands lately, so a lot of amateurs must be working on their code speed for the Extra class license. Also, during some of the DX contests and state QSO parties it was evident that many of the stateside CW operators had been working on their code—speeds were up and operating practices were better.

If you haven't thought about a higher class license yet, now is the time to do it. Next month the new sub-bands go into effect—at the present time about 50,000 amateurs will be able to use them. The new Advanced class is available to everybody, and I've been surprised that so few amateurs have even tried for it. By latest count, the number of advanced licenses has only increased about 5%. On the other hand, the number of Extra-class licenses has gone up more than 50% over a year ago. Perhaps the small increase in Advanced licenses is due to fellows going directly to the Extra class.

In any event, the number of higher class licenses is paltry when compared to the total number of amateurs in the United States. It looks to me like about 120,000 amateurs should be able to qualify for the Advanced class with a simple multiple-choice test. Another 30,000, the Conditional licensees, can get into the Advanced class with the addition of a code test. Since it's only 13 words per minute, that shouldn't be any problem.

Jim Fisk, W1DTY
Editor

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Filament Current A	14 5	14 5	21 5 23 0	28/33	51	94/104

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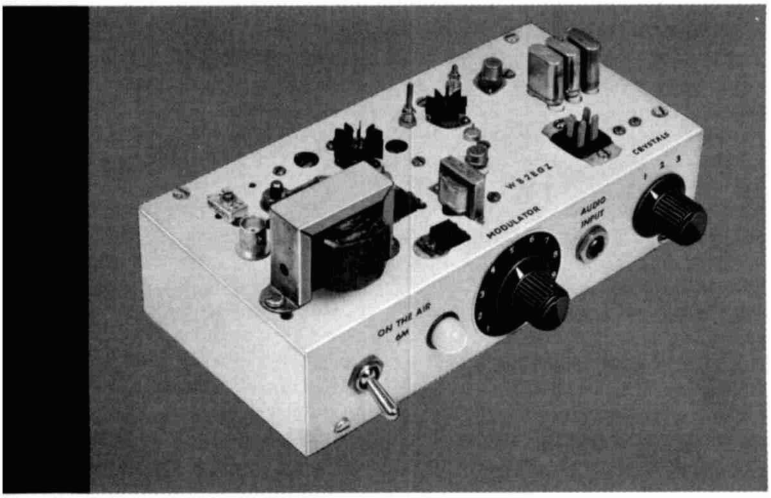


higher power from transistors on six meters

Although the cost of solid-state rf power is still not competitive with vacuum tubes, transistors offer some interesting advantages

Don Nelson, WB2EGZ, 9 Green Ridge Road, Ashland New Jersey

It's time we took the next step in power with rf transistors! To my knowledge, Hank Cross, W1OOP, holds the QRO honors with his 4-watt, 6-meter rig.[1] Other amateurs have probably made transmitters as large as Hank's, but they have failed to share their efforts through publication.

Ed Tilton, W1HDQ, published a notable 6-meter a-m design2 which appears in the new ARRL Handbook. This circuit, because of its relative simplicity, is preferred for beginners. Don't misunderstand. Ed's design is out of the toy-and-gimmick class, although I confess I had to see the unit personally and try a few designs of my own to appreciate the Tilton genius for finding the least expensive and most flexible design.

There have been a number of 2-transistor transmitters in the amateur radio magazines, but time has relegated these to gadget Valhalla. My own work on two-meters" prompted me to try higher power—this time on the six-meter band. Whether you choose this approach or something simpler, here are a few useful hints to help you along the road to success.

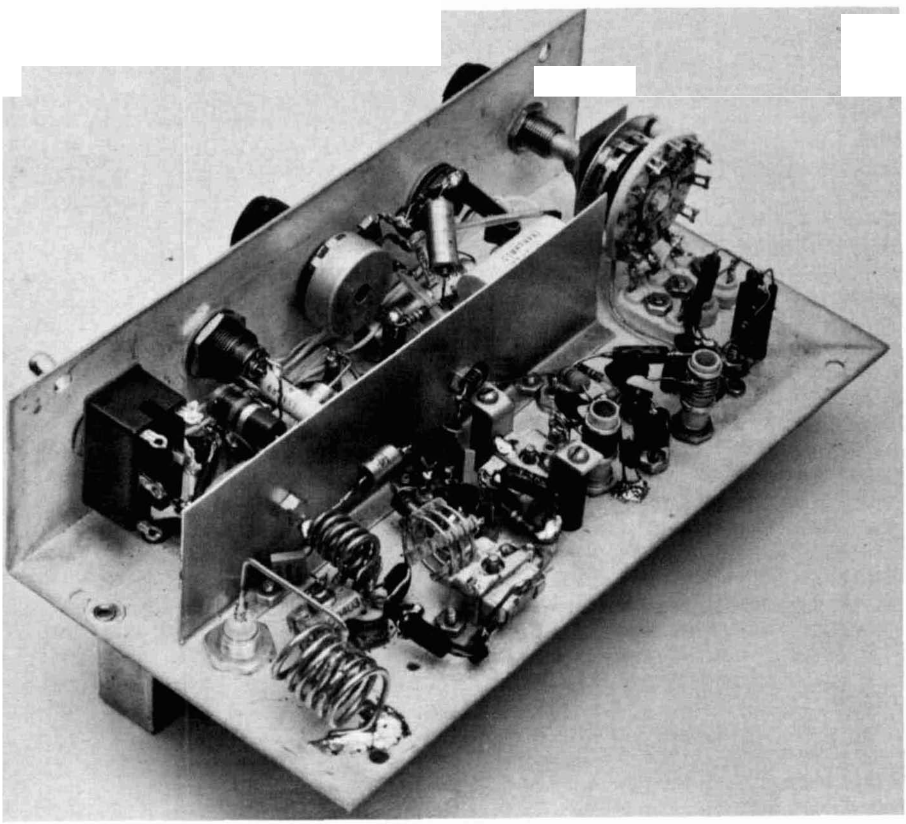
the circuit

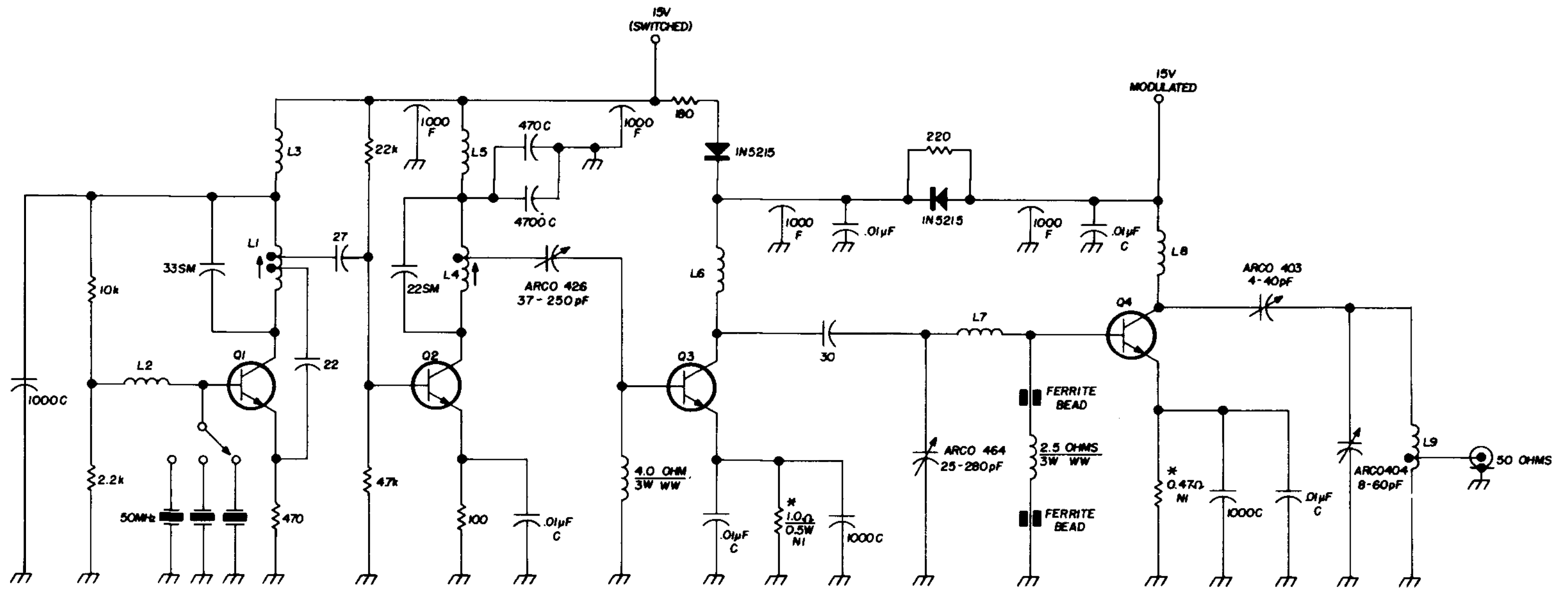
On a dollar-per-watt basis, transistors are still not competitive with tubes. In order to compensate, low-cost parts were used where possible without sacrificing performance. The oscillator stage exemplifies this philosophy. An RCA 40081 is used in a 50-MHz overtone crystal oscillator circuit. Low cost International Crystal EX crystals are suitable here. You could possibly start at 8 MHz with only one additional stage, but I felt that the project was sufficiently difficult for a band already troubled with TVI without adding to the problems with high-order frequency multipliers. Any of three crystals may be selected by a switch on the front panel. It is practical to switch frequency over a 300-kHz band without retuning.

Following the oscillator is a class-A buffer using an RCA 40405 or 40519. The circuit is designed to reduce loading on the oscillator while providing moderate gain to the next stage. I used similar circuitry in my two-meter transmitter and it has proven to be a reliable design. This stage may run warm, so I used a clip-on heat sink. Several larger transistors (TO-39) were tried; although they ran cooler, they didn't have the gain of the RCA 40519.

The next amplifier, an RCA 40290, differs from the previous stage in that it is modulated (a 2N3553 may also be used in this circuit). Diode switching of the modulated collector voltage is very effective in supporting upward modulation of the final. Although low-Q chokes are frequently used as base returns in class-C circuits to suppress motorboating, I used wirewound resistors for both this stage and the final instead of special low-Q chokes which may be hard to find.

The final is a 2N2876 mounted under and heat-sunk to the chassis. Other types which





L1	5 turns number 22 on Miller 4500-2 slug-tuned form (red core); tapped at ½ and 1 turn from power supply end.
L2, L3	6.8 uH (Miller 74F686AP)
L4	8 turns number 22 on Miller 4500-4 slug-tuned form (white core); tapped 2 turns from ground end.
L5	4.7 uH (Miller 74F476AP)
L6	6.2 uH (Miller 4610)
L7	3 turns number 20, ½-inch diameter (B&W 3002)
L8	5 turns number 18, spacewound, ¼" diameter
L9	7 turns number 16, spacewound, ½" diameter, tapped 2½ turns from ground
Q1	2N3118 or RCA 40081
Q2	RCA 40519 or 40405 mounted on Wakefield NF201 heat sink
Q3	2N3553 or RCA 40290 mounted on Wakefield NF207 heat sink
Q4	2N2876 or RCA 40291 for 6.5 W output

* The 0.47-ohm non-inductive resistor in the emitter of Q4 consists of two 1-ohm, ½-watt resistors in parallel (see fig. 5).

Capacitors: C = ceramic; F = ferrite-loaded feedthorugh; SM = silver mica; resistors: WW = wirewound; NI = non-inductive

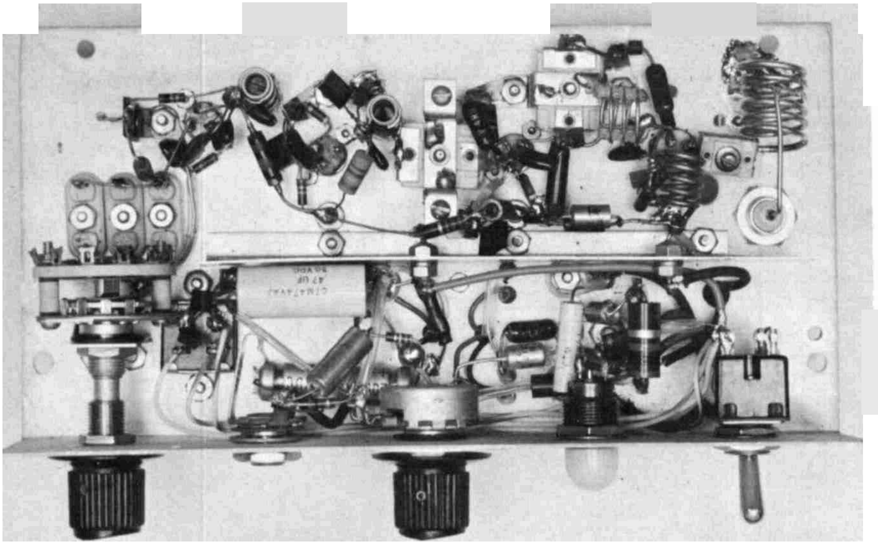
fig. 1. Schematic diagram of the solid-state six-meter transmitter. The inductors in the base leads of Q3 and Q4 are wirewound resistors.

may be used in this stage will be discussed later. A 112-ohm resistor in the emitter limits current peaks. While this is degenerative and costs a little power output, it will prevent burning out the 2N2876 by excessive rf drive or certain regenerative phases which occur during tune-up. Unlike the base return, this

capacitor and the 0.05-μF capacitor between the 2N5295 collectors for the most pleasing results.

construction

Construction begins with a Bud CU592 Converta box. This is the fastest approach to



resistor should be non-inductive. The double by-passing shown in fig. 1 will overcome most of the loss created by the emitter resistor. Note that one bypass capacitor is a high value (.01 μF). This is needed to help prevent motorboating in the final.

Other ways to prevent low-frequency oscillation motorboating are by using low-Q chokes in the base return, ferrite beads on emitter and/or base leads and double bypassing of the 15-volt supply points. You shouldn't encounter motorboating with these precautions, but some transistors are more prone to oscillation than others.

Modulating this rig is not too difficult—beyond the need for rf driver modulation. A CA3020 integrated circuit is transformer coupled to the push-pull 2N5295's (fig. 2) for a powerful but economical amplifier. Some changes in tone may be desirable depending on your microphone. Experiment with the value of the microphone input

a single-chassis design. Since the box is cadmium-plated steel, the chassis is not suitable for 50-MHz work. A wrap-around plate of the same dimensions was cut and formed from .040" brass as shown in fig. 7 and silver plated. If you can't silver plate, copper is preferable to brass.

With all the parts mounted on the same plate, assembly of the transmitter is simpler than the photograph indicates. However, the order of construction must be followed. First, mount sockets and components on the top and front of chassis allowing for correct positioning of the crystal switch next to the crystal sockets (fig. 3). It may be practical to mount and wire the pilot lamp after the modulator and switch wiring is completed. **Don't** install the shield (fig. 4) until the modulator is wired.

The second step is to wire the modulator and power switching at the front of the unit. I found the wiring of the integrated circuit socket to be tedious. It's helpful to use Teflon sleeving over each IC pin connection after soldering to reduce the possibility of solder bridging the leads. You can apply power to the modulator for checkout before going on to the rf wiring. With an 8-ohm speaker as a load and a crystal microphone at the input, the amplifier is a very effective PA system. This is the way I checked out my unit.

Finally, the rf section is wired at the rear

1. With all transistors in their sockets and the power off, tune the oscillator, buffer and final tank circuits with a grid-dip oscillator.

2. Remove the 40290, apply power and tune the oscillator and buffer tanks for maximum output. This is probably best seen on the S-meter of your receiver which should be on during tune up.

3. With power removed, install the 40290

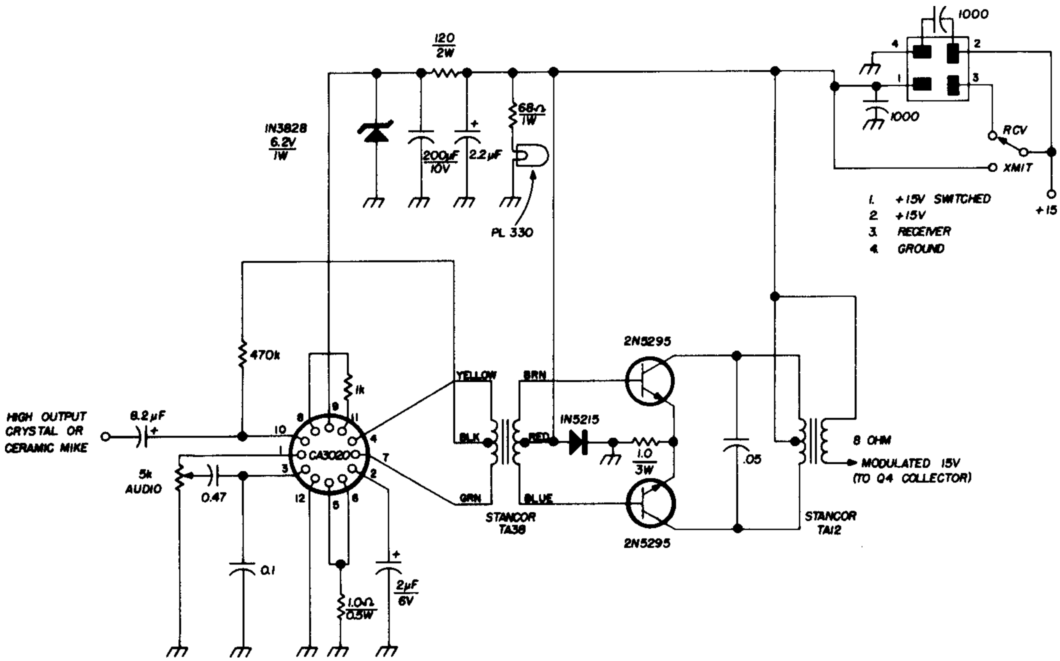


fig. 2. Five-watt modulator for the six-mater transmitter.

of the chassis. Keep the leads on the bypass capacitors short. If double bypassing is used, route the second capacitor to a different ground position than you used on the first. Meticulous wiring of the rf section will pay dividends in output power. The layout of the 2N2876 emitter circuit which is obscured in the photographs is sketched in **fig. 5.**

tuneup

The tuning procedure is less difficult at 50 MHz than at 144 MHz, but certain costly pitfalls are still present. Suggested steps for alignment are as follows:

and connect the output of the transmitter to a suitable 50-ohm load. Assuming the circuit is stable when power is applied, the interstage coupling networks and final tank may be tuned for maximum output.

An SWR bridge is useful for checking output to the load. I should mention that several combinations of tuning-capacitor settings may give good output. Use the optimum combination. A good 2N2876 should deliver 6 to 7 watts. Slight retuning of the oscillator and buffer may be necessary for best results.

If you detect spurious oscillations or motorboating in the receiver at any time during tune up, power down! Check all rf transistors for excessive heat and don't apply power until overheated units have cooled. Retuning will usually correct the instability. Overheating of the transistors may occur during tune up. **Check** for **excessive heat frequently;** power down for cooling.

4. Optimum modulation characteristics will only require a minor tuning adjustment. For this procedure it's preferable to use a

fig. 3. Crystal-switch assembly; switch bracket is shown in fig. 4.

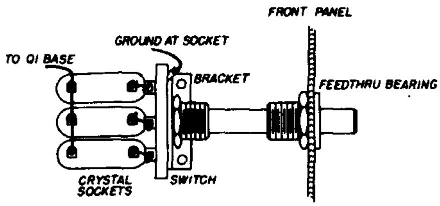
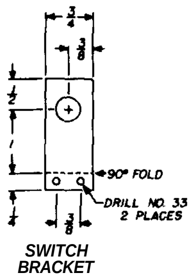


fig. 4. The crystal-switch bracket and rf shield are made from 0.040 aluminum.



low-level sine wave driving the modulator. With a monitor scope or the setup shown in **fig.** *6.* In this set up, the vertical input of the scope is connected across the output of the final i-f stage in the receiver. The rf envelope of the transmitter will be displayed when the receiver is tuned to the frequency of the transmitter.

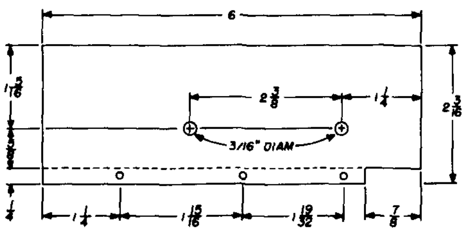
Increase modulation to 30% and retune for minimum distortion. Advance the modulation 50%, then to 80%, adjusting for minimum distortion each time. Don't try to reach 100% modulation! Remove the sine-wave generator and check the modulation level with the microphone. Note this setting on the front panel for future reference.

the fruits of our labor

While the primary focus of interest was to build an all solid-state 6-meter transmitter, some experiments in the use of different final transistors provided interesting and useful material. Three transistor types, the 2N2876, the 2N3375 and the 2N3632 were available and suitable for the application. The recorded output power readings for each type are shown in **table 1.**

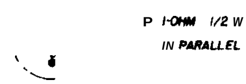
There are conceivably many errors in the equipment, setup and tuning of the circuitry because my shack is not a quality laboratory. Furthermore, rather large differences between similar transistors made by different manufacturers can be seen, so don't expect to duplicate these results precisely. Certain conclusions may still be made, however.

1. The insertion loss of the Drake filter was measured at 2 dB. This means that 19% of the 50-MHz power will not be delivered to the antenna. A simple calculation of chart data will show that filter output is less than 81% of its input. Additional loss is assumed



RF SHIELD

fig. 5. Wiring the rf power emitter bypass capacitors are grounded directly to the transmitter case.



to be in harmonics which are suppressed by the filter. I would recommend using a filter with this transmitter because of the magnitude of these harmonics. The second harmonic—which falls in the fm band—is the prime offender.

table 1. Comparison of several rf power transistors.

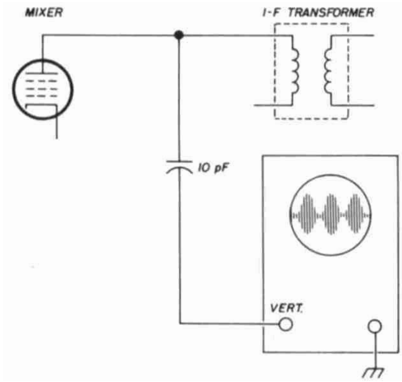
transistor	power out (with filter) (watts)	power out (unfiltered) (watts)	price
2N2876	4.8	6.5	$14.50
2N3375	7.5	10.0	$14.52
2N3632	9.5	12.5	$20.00

2. The 2N3375 is the most efficient, but not significantly different than the 2N2876. On a dollar-per-watt basis, the 2N3375 is also preferable. Why then, did I choose a 2N2876 for the transmitter? The primary reason is that the transistor is a lower frequency device and inherently more resistant to burnout through improper use. Probably the most common trouble is the occasional second-breakdown destruction of a transistor during tune up. While the circuitry used here should not be prone to second breakdown, I still prefer the 2N2876 for trouble-free operation at 50 MHz.

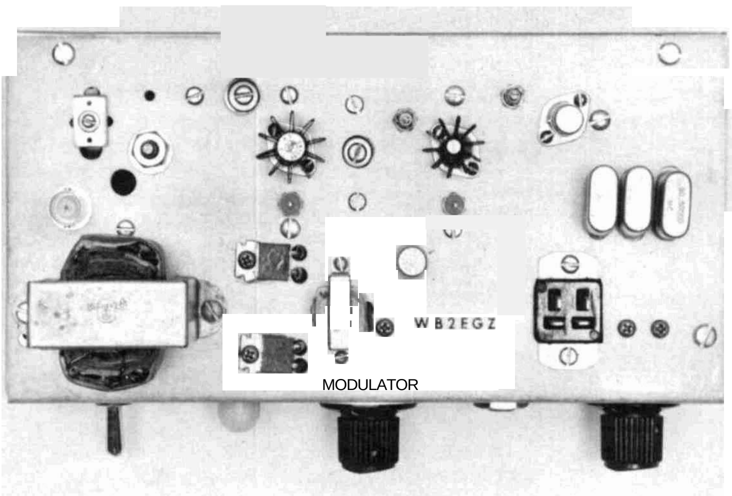
3. The 2N3632, which contains two matched 2N3375's in the same package, doesn't give twice the output of the latter type. This is a well-known fact in engineering circles but perhaps not well known to amateurs. My personal experience with the 2N3632 was less than gratifying. The first did not survive tune up, while the second was gassed by a lengthy *QSO.* A better heat sinking arrangement would be helpful for persistant proponents of that device.

Some experiments were performed using different coupling techniques with efficiency and purity of output as objectives. Results showed a preference for tapped tank circuits over L's or π's for rejecting unwanted harmonics. Matching to the tank is a trifle more difficult, but once achieved, power transfer is comparable. Before you attempt to achieve

fig. 6. Method used to connect the transmitter signal to an oscilloscope.



the ultimate in matching, be sure that you have 6 to 7 watts output from the circuit as shown. The greatest losses are more likely to be caused by lead dress and poor bypassing than by incorrect matching of tank circuits.



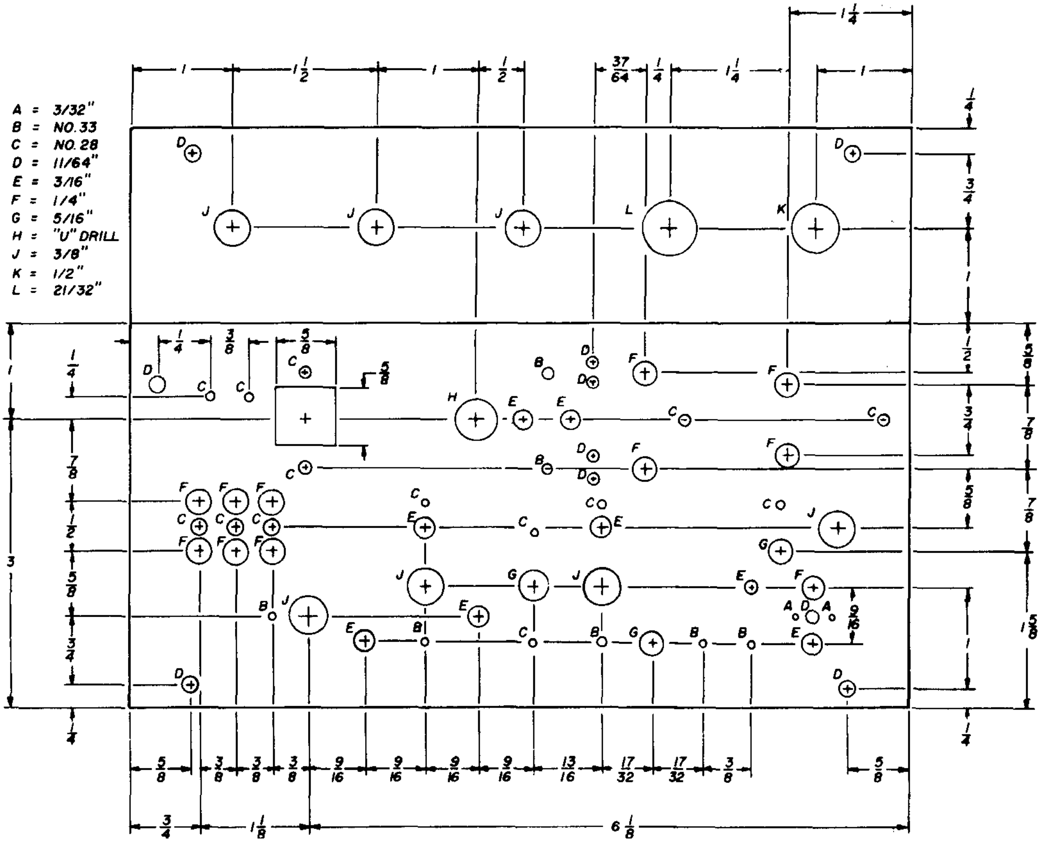


fig. 7. Chassis layout for the six-meter a-m transmitter. Make from 0.040 brass and silver plate. This chassis is designed for a Bud CU592 Converta box.

conclusion

Here is a fine little transmitter with clean modulation and power in the class of many commercial rigs. The cost of the semiconductor complement is approximately $33.00—this is probably more than half the total cost. The 40290 runs warm on prolonged transmissions, but not warm enough to cause concern. All other semiconductors are cool to the touch when heat sunk as shown. I've performed no miraculous DX, as yet, but that's because I'm only using a halo antenna.

When using a 15-volt power supply, the transmitter idles at 950 mA. Modulation peaks are about 1.5 A. If the power supply is unregulated, the transmitter should be tuned at the lowest voltage that will be encountered. Power output will vary with supply-voltage changes, but the detuning effect which is troublesome with rf power transistors at higher frequencies is undetected on 6 meters.

This is not a project for the beginner, but it's not really difficult if the plan is carefully followed. When you tell others what you're running, you'll have a real feeling of accomplishment.

references

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3. D. W. Nelson, WB2EGZ, "A Solid-State A-M Transmitter For Two-Meter Operation," RCA Ham Tips, Fall, 1966.

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ham radio

a simple
panoramic reception
method

Complete panoramic adapters are expensive, but for occasional tests that require panoramic displays, here is an approach that uses test equipment from around the shop

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

Panoramic reception, which lets you see as well as hear received signals, has many uses for both operating and testing equipment. Various commercial adapters have appeared on the amateur equipment market but if you already have the required test equipment—a sweep oscillator and an oscilloscope—a panoramic unit can easily be built for any receiver with a simple adapter. If you don't want to tie up the test equipment permanently for panoramic reception, you can arrange the adapter so that the test equipment can be disconnected and used elsewhere when desired.

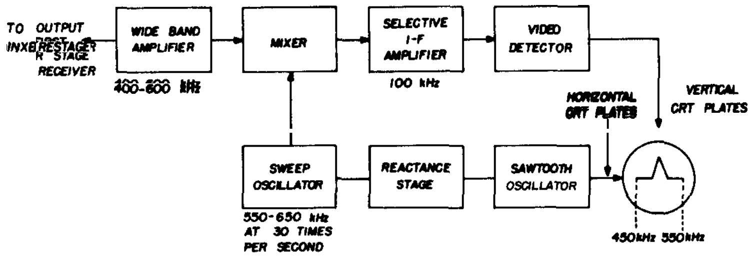
basic panoramic reception

The block functions of a conventional panoramic adapter connected to a communications receiver are shown in **fig. 1**. Basically, the panoramic adapter is an electronically-tuned receiver with its output displayed on a cathode-ray tube on an amplitude-vs-frequency scale. The sawtooth oscillator is used to drive the horizontal sweep on the CRT at the same time it varies the frequency of the oscillator. The vertical deflection circuit is driven by the rectified output of the i-f stage.

The same effect can be obtained if you have a separate receiver which is manually tuned through the i-f of the main receiver. The analogy to a manually-tuned receiver is worth-while because with this in mind it is easy to appreciate both the values and limitations of panoramic reception. As with any manually-tuned receiver, separation of closely-spaced signals requires good selectivity. However, the better the selectivity, the slower the receiver must be tuned—otherwise signals will be missed. Therefore, a panoramic adapter must be a compromise between selectivity for signal resolution and frequency scan speed (sweep rate).

Expensive units have elaborate controls for varying these parameters while less expensive units use compromise settings. Also, the band of frequencies displayed affects the circuit doesn't have high Q. Most sweep oscillators also provide a fixed, non-sweeped, marker output so you can calibrate the horizontal frequency scale on the oscilloscope.

fig. 1. Block diagram of a conventional panoramic adapter. Frequencies are for processing a single 500-kHz signal; sweep width is 100 kHz and sweep rate, 30 Hz. The 500-kHz pip is displayed on the CRT each time the deflection plates and sweep oscillator are swept through their range.



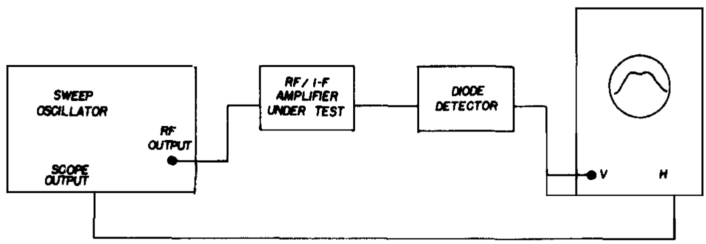
scan rate. Extreme selectivity and a slow scan rate (1 second, for instance) may be desirable if you are scanning a small frequency range —say ± 3 kHz—either side of the i-f. If you are scanning a larger frequency band (500 kHz, for instance) resolution must be sacrificed for using a scan rate of up to 60 times a second.

sweep-oscillator circuit

The conventional way of using a sweep oscillator to check the selectivity response of an i-f amplifier is shown in **fig.** *2.* The similarity to panoramic reception is apparent. Many commercial sweep oscillators cover a fairly wide range—from a few MHz through vhf. The frequencies that they usually sweep are a maximum of about ±5% of the center frequency. The sweep range is set on all but the most expensive units at 60 Hz to correspond to the ac line frequency.

The resolution on the oscilloscope display is usually only adequate when the sweep

fig. 2. Method of using a sweep oscillator for checking or i-f amplifier response curves.



A sweep oscillator and an oscilloscope can be combined to function as a panoramic adapter as shown in **fig.** 3. The CRT display circuits, the sawtooth oscillator, reactance stage and sweep oscillator are all contained in the oscilloscope and sweep oscillator units. The adapter supplies the buffer (or preamplifier) mixer and i-f and detector stages. It should be noted that the adapter is equivalent to the front end and i-f stages of a conventional receiver. In fact, a conventional receiver may be used to build the adapter. The output from the first mixer in the receiver is connected to the antenna input of the auxiliary receiver; the detector output of the auxiliary receiver is coupled directly to the oscilloscope vertical input. The local-oscillator signal to the mixer in the auxiliary receiver is supplied by the sweep oscillator.

If you use care when selecting the injection frequency from the sweep oscillator, you'll find that a wide variety of auxiliary receivers are usable. In fact, when the first major output of the basic receiver falls within the a-m broadcast band, as many double-conversion receivers do, an inexpensive transistor BC receiver will work nicely as the adapter.

A more generally useful but relatively simple adapter is shown in **fig. 4.** Basically, this signal. The 2N2672 i-f amplifier is a conventional high-gain circuit with a neutralizing loop to improve stability. The second 1N541 serves as a video detector. Because of the time constants used in its filtering circuit, it is coupled directly to the oscilloscope vertical input.

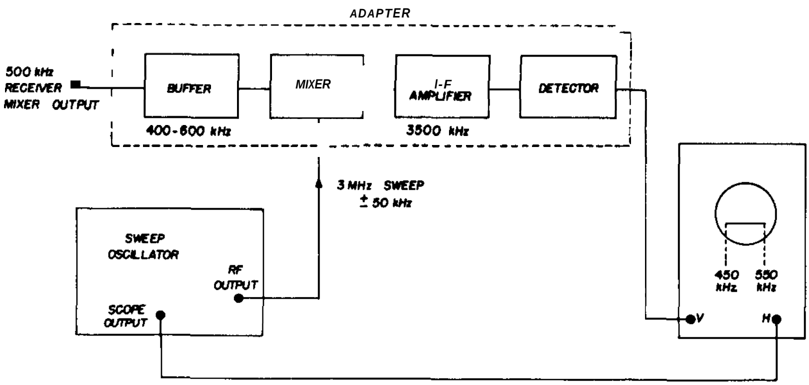


fig. 3. By using an adapter, a sweep oscillator and oscilloscope can be used to provide panoramic displays.

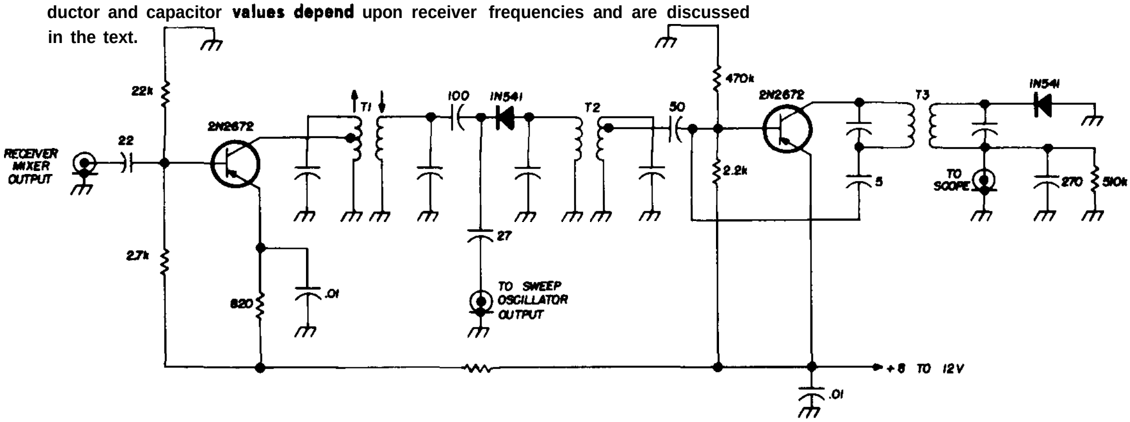
unit encompasses all the stage functions shown in **fig.** 3 and can be adapted over a wide frequency range. The 2N2672 input stage is coupled to the first mixer stage in the receiver. This is the signal that will be panoramically scanned. The input is untuned and presents an insignificant load-detuning effect on the basic receiver as long as a short length of low-capacitance cable is used to couple it to the basic receiver.

The first 1N541 diode mixes the output from the first mixer with the sweep-oscillator

The adapter can be built to cover almost any high frequency range (to about 40 MHz), depending on the mixer output frequency and the sweep oscillator range. If your frequency range is the same as shown in **fig.** 3, T1 **(fig. 4)** is a 500-kHz i-f transformer, and T2 and T3 are 3500-kHz transformers. When aligning the adapter, carefully peak T1, T2 and T3. T2 and T3 are simply peaked for maximum at the desired i-f frequency. T1 must, however, be tuned a bit more carefully.

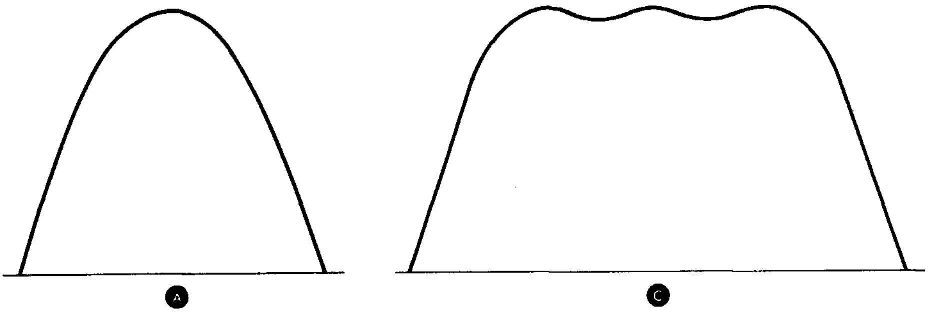
The selectivity curve of the first mixer is

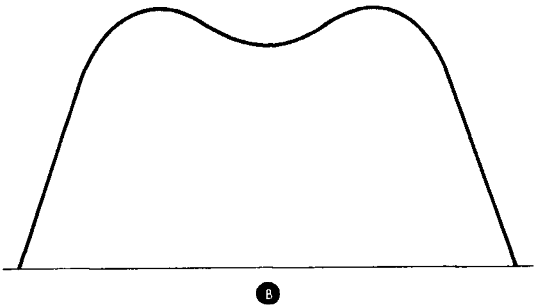
fig. 4. Simple solid-state panoramic adapter unit. The first 1N541 is the mixer; inductor and capacitor values depend upon receiver frequencies and are discussed in the text.



approximated in **fig. 5A.** The response is fairly broad since the selectivity is a result of the first stages of the i-f strip. The panoramic adapter must be connected at this point—not later in the i-f strip. Otherwise it will be impossible to scan anything but a frequency range restricted by i-f selectivity. If T1 is

fig. 5. Tuning of the buffer-stage collector circuit must be stagger tuned as shown in A so the receiver selectivity curve shown in B doer not degrade response at the ends of the passband. The resultant selectivity curve is shown in C.





simply peaked to the same response as **fig. 5A,** the oscilloscope display will favor signals near the center of the i-f pass band.

To avoid this, one half of T1 is peaked at the low end of the scanned frequency range (400 or 450 kHz, for example) and the other half is peaked at the high end of the frequency range (550 or 600 kHz, for example). The response form of T1 will look like **fig.** 5B; the over-all response will be somewhat like **fig. 5C.**

This provides equal amplification for all signals within the scanned frequency range. Alignment is easily accomplished by coupling a signal generator to the mixer input of the adapter and peaking each half of T1 for maximum amplitude with the generator set alternatively to the low and high limits of the scanned frequency range.

summary

Once you understand the basic theory of panoramic displays, it should be simple to devise an adapter unit for any individual situation. As I mentioned before, in some cases a simple transistor BC receiver can serve as the adapter. In other cases, an adapter similar to the unit shown in **fig. 4** can be used.

In any case, you should appreciate both the advantages and disadvantages of this method of panoramic reception to obtain useful results. The relatively fast sweep rate of most sweep generators precludes the use of an adapter with very sharp i-f selectivity. Therefore, the adapter is less expensive to build, but displayed signals will appear to be much broader than they really are.

Some sweep generators provide variable sweep rates. If you have a sweep generator like this, the i-f selectivity of the adapter unit can be increased by using better i-f transformers or crystal filters. With narrow selectivity, detailed oscilloscope displays are possible—even to checking transmitters for intermodulation-distortion products. However, even with broad displays you can scan a broad portion of a band for general DX activity or compare relative amplitudes of various signals.

Although you may not want to look at received signals continuously, the panoramic adapter described here provides a uniquely simple way of using readily available test equipment.

ham radio

basic electronic units

Many amateurs learned basic electronics by rote so they were never exposed to the basic units– here is a building block that will further your understanding of radio communications

Jim Ashe, W2DXH, Peterborough, New Hampshire 03458

Have you ever wondered what an ohm really is? Or a volt, or an ampere? It's interesting to know that these names, and many others in the electronics field, are the names of workers who did much of the research which makes modern electronics possible. But that doesn't tell us very much about the units. Is it necessary to know what an ohm 'really' is? No, and some people feel that such knowledge can be cataloged under the heading of "useless facts."

I can't agree with that. This information is the useful basic knowledge which (if not taken in too large doses) lets you keep up with modern ideas. It helps provide new answers to problems that will stump your less inquisitive fellow workers. Basic knowledge is not useless facts.

The appropriate basic knowledge will give you a feel for the meaning and application of electronic units. Once you know most units are built up from a very few basic ones, and see how this happens, you can begin to understand something about how the units should behave on paper and in real live circuits. Plugging numbers into formulas is like walking blindfold in a strange room; you tend to feel uneasy.

dimensional analysis

This awesome title applies to a common-sense idea. When one or more basic units are combined into a more complex unit, they don't just disappear. An essence remains, usually not very far below the surface. The part that stays is called "dimension." For example, the dimensions of speed are distance and time. Even if there were a special automotive speed unit called **'oldfield,'** whenever anybody used the term, they would mean miles traveled by hours for the trip, or miles per hour. Think about **hertz** and other examples in **table 1.**

It's unfortunate that electronic units are commonly introduced to beginners without any mention of their inside composition. For instance, capacitors are frequently described as things that hold electricity. But who mentions how much electricity they hold? Or how the circuit sees it? I think you will find electronics much simpler once you've been introduced to the basic unit of electricity: the **coulomb.** When you have a little feel for coulombs in capacitors you can appreciate, for instance, that 10 microamps for 1 second will fill a 1-μF capacitor to 10 volts; the idea of volt-seconds will make inductive circuits more accessible to common-sense thinking.

some common units

A basic unit is a kind of building block, something like a brick or an atom. That is, upon close examination, you can find inside detail, maybe a lot of it, but that detail doesn't mean much for practical purposes. You can build excellent walls without being an expert on the theory and design of cinder blocks, and in the same way there is not much reason for getting into philosophical detail about seconds, coulombs, and volts.

If you become interested in this subject, you'll discover some of the basic units I chose for this article could be replaced by other basic units. The result would be different definitions, which for practical purposes would work out as well as the ones I've used here.

Let's begin with the **second.** Time is fundamental. Everything we do, and so far as I know, everything that happens in the universe, occurs in such a way that time is a necessary part of the action. We measure time by looking at some device that ticks it off in equal-sized units, based upon the second. If seconds are too short we can take them in batches from minutes to millenias, or if too large, in decimal fragments from milliseconds to femtoseconds. You haven't heard that term? A femtosecond is one-billionth of a microsecond, used by nuclear physics workers to time rapid events inside atoms. But that's another subject. Very careful measurements have brought out some slightly different opinions as to how long a second is, but for our purposes, they are all the same; available in liberal supply from WWV.

The next basic unit is the **coulomb.** Electric current is moving charge, or coulombs in motion. There is a natural unit of charge, the electron, but this unit is too tiny for practical purposes. From a human viewpoint, the coulomb is a more meaningful unit.

A coulomb is a perfectly definite number of electrons: 6.23 x 10^{18} (in scientific notation). I've emphasized the quantity aspect in **fig. 4,** which may not be as unlikely as you think. There is some interesting work being done with solutions of free electrons, reported in **Scientific American Magazine.** To a physicist, one coulomb of electricity is that positive charge which, placed 1 meter from an equal positive charge, repels it with a force of 1 newton. That's about equal to the weight of 3.6 ounces in your hand.

Electrochemistry offers another definition: passing 1 coulomb of electricity through a silver chemical solution will plate out about 1.1181 milligrams of silver. Each electron

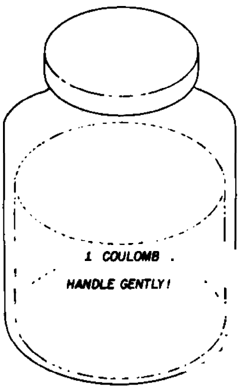
table 1. Electrical quantities and how they are built.

unit name	unit quantity	breakdown
hertz	frequency	cycles / seconds
second	time	basic
coulomb	charge	basic
volt	pressure	basic
ampere	flow	coulombs / seconds
ohm	resistance	volts / amperes
joule	heat/work	volts x coulombs
watt	power	volts x coulombs / seconds
farad	capacitance	amp-seconds / volts
henry	inductance	volt-seconds / amps

moves one atom of silver onto the cathode, and since there is a perfectly definite number of electrons, and each silver atom has the same weight, the silver buildup indicates the charge in coulombs passed through the bath.

Pushing coulombs through wires takes pressure, and this is measured in **volts.** Voltage is a kind of electrical pressure, responsible for the flow we call current, or we may find it more convenient to think, caused by the current. Voltage and current are related by Ohm's law, which we will come to shortly. Voltage, like the push of a spring, is not necessarily accompanied by motion. Slightly peculiar fig. **2** emphasizes the pressure aspect of voltage. We don't need a really good definition of voltage because our meters are calibrated to read it directly. This is the last of the basic electronic units. Now we can start putting them together.

fig. 1. If we could store a coulomb in a bottle, it might look something like this.



the ampere

We come first to the **ampere,** one coulomb of charge per second. A meter reading of amperes can be interpreted as coulombs per second, or we can say amp-seconds equals coulombs. This is dimensional analysis again, useful in work with timing circuits, batteries, transistor circuit design and other circuit matters.

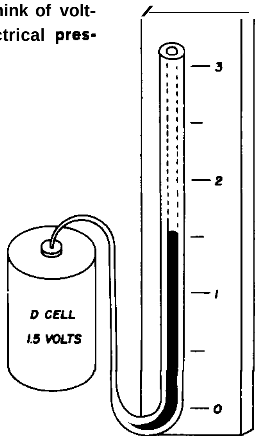
Electrical resistance is something we always find in wire, components, transistors, etc., which turns some of our electrical energy into heat. Electrical pressure in volts is required to push coulombs across resistance. This necessity is summed up in Ohm's law: $R = E/I$. Dimensional analysis tells us ohms equals volts per amp, which we can put to work immediately. See fig. **4.**

We want to choose a resistor to give us 10 volts from 2 milliamperes. Sure, we can substitute these values into Ohm's law and come out with a figure. But try it this way: ten volts per two milliamps is five volts per milliamp, or 5,000 ohms. An ohm is a volt per amp, but this is rather large for electronics, so I restated it as 1000 ohms is a volt per milliamp. One hundred thousand ohms is a volt per 10 microamps or a rnilliamp per hundred volts, etc. Remembering dimensional analysis we change our view to get the best hold on the things we're working with. If your screwdriver points the wrong way when you pick it up, you change your grip before you try to turn screws with it.

work and power

The **joule** is the electronic unit of work, or heat. When we push a coulomb across a volt of potential, one joule of heat is released. One ampere through one ohm, or across one volt, which is the same thing, will release one joule per second as shown in fig. **5.** The joule sounds like a basic unit, but we can make it up as a volt-coulomb, volt-amp-second, $amps^2$-ohm, etc. A joule will heat 4.186 grams of water l-degree centigrade, which is more useful knowledge than you may think: this relates quantity of heat and temperature change. The joule is a perfectly definite

fig. 2. Let's think of voltage as an electrical pressure.



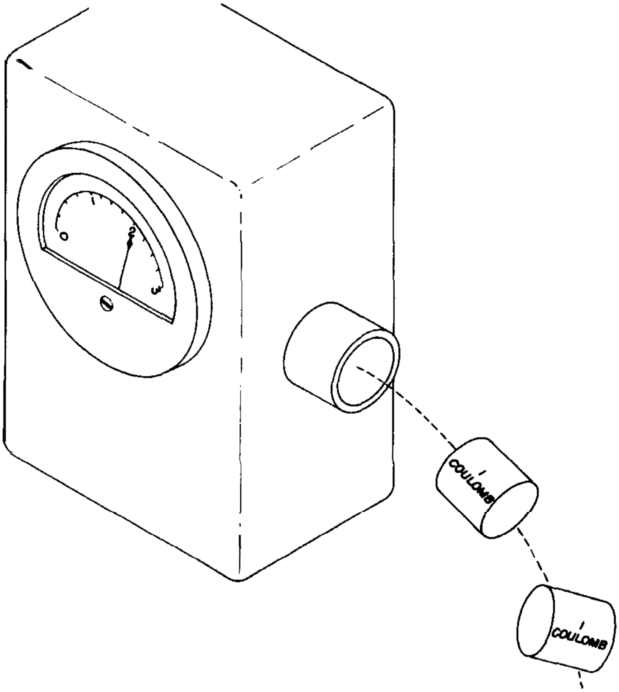
amount of energy, a standard unit of heat or of work. Stored energy, as in capacitors, is calculated in joules.

The watt is the unit of energy flow. In electronics it may refer to power input or output, or merely to heat dissipated in a working component. A watt is one joule per second of heat or work, continuously produced. A one-watt resistor can dissipate up to one joule per second, although good design practice is to limit this to maybe 70% of maximum.

We can put these ideas to work by finding out how to measure transmitter power output without making any rf tests. We merely turn the output power into heat, and find out how much heat is produced in, say, five minutes. Heat is much easier to measure than a combination of rf voltage, current and phase angle.

This method is called calorimetry, which is "measuring heat." You can improvise the required equipment from things that should be around most any ham shack, but since you are

fig. 3. Electric current is a flow of charge which is measured in coulombs. Here are 2 coulombs per second, or 2 amperes.



improvising, some care is required. Begin by putting your dummy load into a thermally insulated box. Say, three inches of Zonolite all around, at top and at bottom. This prevents heat dissipating in the load from leaving the load, which could be destructive if carried too far. We aren't going to overdo, and if we are careful, no harm will result. I've pictured the setup in **fig. 6**.

fig. 4. Electric current produces voltage across resistance.

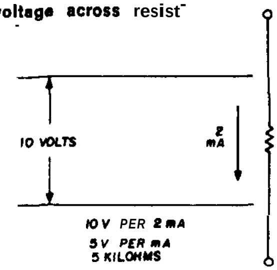
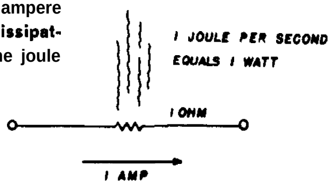


fig. 5. Here is an ampere through an ohm dissipating one watt, or one joule per second.



Next, we calibrate the setup by feeding some dc power into it and measuring the resulting temperature increase. We start from room temperature. Knowing dc voltage and current, we know power in watts; watt-seconds gives the total amount of heat in joules poured in during the calibration run. For instance, in five minutes at 100 watts we have put 30,000 joules into the load, which has become 10 degrees warmer. This is 3000 joules per degree. After the load has cooled to room temperature (overnight) we can do a transmitter test run, and let us suppose the load temperature went up 15 degrees in five minutes. This must have been 150 watts. To be doubly sure, we repeat the test later with the same amount of dc power, and we should see the same temperature change.

I've only touched on calorimetry here. There's a lot to it, which you can find in any basic physics book. A good calorimetric test setup will completely eliminate all uncertainties about rf voltage, current and power.

inductance and capacitance

Before defining the units of capacitance and inductance, I want to mention their surprising complex-yet-simple behavior. If you do not immediately see what this is all about, you have a lot of company. Capacitors and inductors, unlike resistors, are strongly frequency dependent. In addition, there is a curious difference in their properties which you will shortly discover is a kind of similarity. Engineers call it duality, and I've emphasized this relation in **table 2.** Reference to handbooks is always good practice (if you understand what you are doing), and in the case of problems dealing with inductance and capacitance, it is particularly appropriate if only to refresh your memory.

The unit of capacitance is the farad, and in electronics we usually see this as the micro- or one-millionth-farad. The farad is too huge for most applications, although there are some one-farad capacitors around now.

If you feed a coulomb into a 1-farad capacitor, measurement will show one volt across its terminals. A typical 1-farad capacitor, rated at 3 volts, will store 3 coulombs. If you fill the capacitor to any definite voltage, its stored energy in joules (watt-seconds) will be $1/2CE^2$ joules, where C is capacitance in farads and E is voltage. Thinking in amp-seconds is appropriate for timing circuits and estimating time to charge a photoflash capacitor; and thinking in joules is suitable for choosing a capacitor in a photoflash lamp project, once the lamp specs are known.

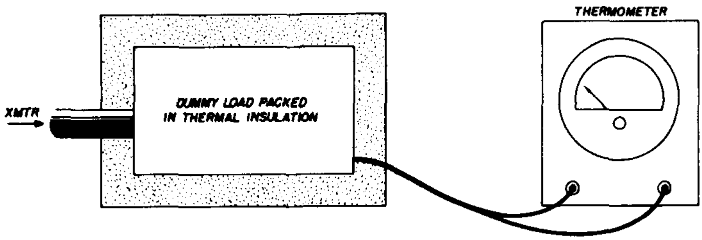


fig. 6. You can measure rf power without an rf voltmeter if you calibrate a thermally-isolated dummy load in terms of temperature rise per joule fed into it. Since the heat has no way to escape, be careful.

Inductors store energy. How do you charge inductors? By feeding volt-seconds to them. Look at **table 2** again. If you apply 1 volt-second (1 volt across terminals for 1 second) to a l-henry inductance, measurement will show 1 amp through its terminals. Since the ampere must continue to flow, you cannot disconnect the inductor as you can the capacitor, but when you see the 'dual' relation in their properties you will understand this is not really a difference. Stored energy in the inductor is $1/2LI^2$ joules, L henries and I amperes. Inductors can be used for timing circuits, although capacitors are most commonly seen in this application. All electrical components are more or less spoiled by unwanted resistance, and capacitors are easily made more "pure" than are inductors.

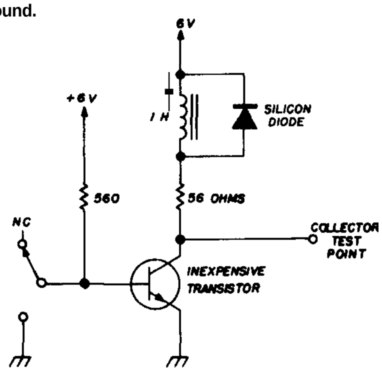
Let's think about inductive circuits that

table 2. The odd mirror-like characteristics of capacitors and inductors.

parameter	description	unit	storage	AC reactance
Inductance L	**Volts across terminals for current in; energy stored in a magnetic field**	**henries equals volt-seconds per ampere**	**$J = 1/2 LI^2$ short terminals to store energy**	$X = 2\pi fL$
Capacitance C	**Current into terminals for volts across; energy stored in an electric field**	**farad equals amp-seconds per volt**	**$J = 1/2 CE^2$ open terminals to store energy**	$X = \frac{1}{2\pi fC}$

blow transistors. Most of us have met, in the book or on the bench, a circuit something like fig. *8.* The relay has an inductance of 1 henry,

fig. 7. Here is a simple inductive-load transistor switching circuit. For simplicity, the transistor is turned off by shorting its base to ground.



which we measured by placing a capacitor across it and finding its audio resonant frequency.

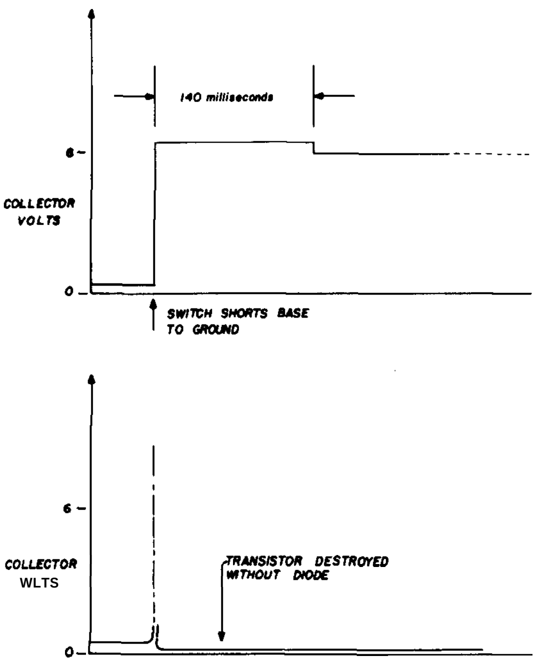
The transistor merely turns coil current on and off: **on** when the transistor gets lots of base current and goes into saturation, **off** when we short the transistor base terminal current to ground. A resistor in the transistor collector circuit controls collector current, since I've assumed the coil has inappreciable resistance. You might want to work this out later, assuming the resistor value is the coil resistance.

Now, we know the turn-on current, which was established by the resistor. But circuit-limited current means we don't have a volt-seconds figure to work with, as we might in a pulse circuit problem. So we restate the inductance in henries as volt-seconds per amp, and we know the circuit will somehow have to dissipate one volt-second for each amp the transistor turns off. The current is 100 mA, so we have to discharge 1/10 volt-second.

Since the coil can produce large voltages if there is no easy way out for its stored energy, we provide a silicon diode. As the magnetic field collapses upon turnoff, the voltage across the inductance builds up to 0.7 volts or so, the forward voltage of the diode. Transistor collector voltage goes to 6.7 volts. Since there is 1/10 volt-seconds to be dissipated, the current flows around through the diode for about 140 milliseconds. Stored energy gone, the voltage across the winding drops to zero, and we can start the cycle again. Estimated collector voltage through the cycle in **fig. *9*** should help with this experiment. I hope you'll start with dime surplus transistors and a transistor socket.

Well, maybe that was a bit of digression, but it was a good one. If you work at these ideas in simple breadboard circuits, you will find things becoming far more real. The usual presentation of electrical units is very stilted and hidebound, and practice in seeing these units at work will aid in circuit design and

fig. 8. This is whet happens to the transistor's collector voltage when the collector current goes off.



servicing problems.

There are some other electronic units that could use a good going over. I'm working on that now. They are the hertz, the mho, the steradian, and the dBm. Perhaps they sound kind of hairy, but they aren't really. Look for "more electronic units" one of these months.

ham radio

vox and mox systems for ssb

There are two popular ways of controlling ssb transmitters–voxand mox; here's how these systems work

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

Among the several advantages offered by single sideband, one of the handiest is its easy adaptability to hands-off automatic operation. This kind of transmitter control is called **vox.** The letters are an acronym for **voice-operated transmission,** but they are usually just pronounced "vox."

At a vox-equipped station, when the operator speaks into the microphone, the transmitter is keyed on, and the receiver is muted. When he stops talking, the transmitter is taken off the air and the receiver operates again.

Even at a station with vox, sometimes it's desirable to turn the transmitter on and off by hand. A push-to-talk (ptt) circuit, operating all the changeover relays from a single switch or mike button, is the most convenient. In contrast to vox, this kind of operation is **mox,** which stands for **manually operated transmission.** Because vox circuitry prevents manual operation, a mox circuit is laid out so it can override the vox system.

importance of vox

Nothing new, vox was developed back in the days when nearly all hams used double-sideband a-m. The reasoning was sound: why run the carrier (and mute the receiver) continuously—when the operator would often pause for several seconds between sentences or thoughts? Someone decided: suppose every time the operator pauses (for more than a few seconds) the transmitter is taken off the air and the receiver turned back on. The ham he is talking to can talk back instantly. In fact, so can anyone else who happens to be on the channel. In other words, any station could **break** in—and that's the term applied to this sort of operation.

Shutting down the carrier of an a-m rig when there's no modulation conserves transmitter power, but not many a-m transmitter power supplies were designed to cope with the sudden up-and-down voltage and current demands. With ssb's suppressed-carrier way of operating, there is nothing being transmitted during speech pauses anyway. There is little problem in just shutting the transmitter off for those intervals and letting the receiver open momentarily to any incoming ssb signals. Vox/break-in operation speeds voice communications and has become a valuable part of sideband operation.

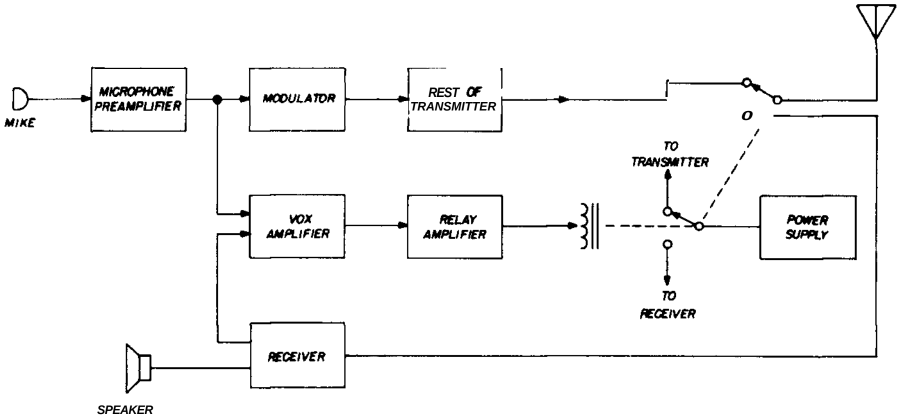
basic vox

In the block diagram of **fig. 1**, you'll see how the typical vox circuit operates. Incoming microphone audio is taken from the output of the mike preamp and fed to the vox amplifier stage, which drives a relay amplifier. When the operator speaks, current through the relay amplifier tube closes the relay. Its contacts switch the antenna and the power supply to the transmitter, putting the transmitter on the air. In some units, the power-supply relay merely removes B+ from certain receiver stages, or biases them off. The result is the same: the receiver is disabled while the transmitter is on the air.

When the operator stops talking, the relay drops back to its normal position. It switches the antenna and power back to the receiver.

At some ham stations, a separate transmitter and receiver are used; at others, a transceiver. In both arrangements, transmission and reception are side by side. Something must prevent sound from the receiver loudspeaker from activating the vox circuit. Refer again to **fig. 1**. A small amount of audio signal from the receiver output stage is fed to the vox amplifier, where it cancels any receiver sound picked up by the microphone. The circuit is called anti-trip or sometimes anti-vox.

fig. 1. Basic vox circuit used in ssb equipment. The anti-trip circuitry that keeps receiver audio from triggering a switchover is also shown.



simple vox

One vox system used in some commercial transceivers is shown in detail in fig. 2. Voice signals from the microphone preamp stage are fed through dc-blocking capacitor C3 and isolating resistor R5 to the grid of VI, the vox amplifier. (C4 bypasses any rf which might be present, to prevent false triggering.) VI normally runs saturated—maximum plate current and very low plate voltage. Positive half-cycles of incoming speech signals have no effect.

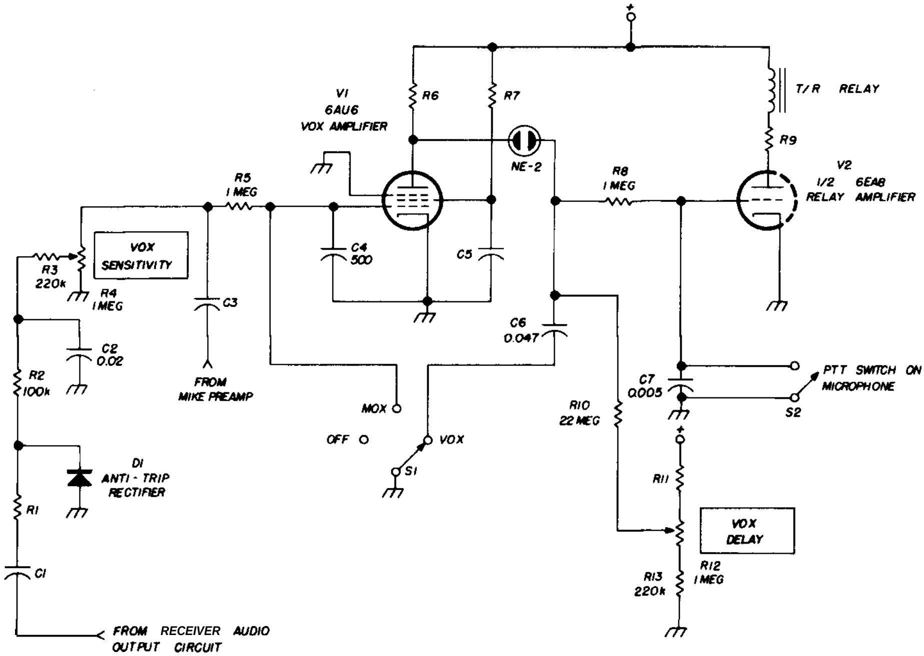
Negative half-cycles, however, drive down the too-positive grid bias of VI, reducing plate current and causing plate voltage to increase during each half-cycle. When the plate voltage gets high enough, the neon lamp fires, applying a dc voltage across R10, control R12, R13, and capacitor C6. C6 charges, and the long time constant of R10, R12, and R13 hold the voltage across C6 fairly steady. The junction between the NE-2 and R8 becomes highly positive with respect to ground, as long as the voice signals continue.

Relay amplifier V2 is normally at cutoff—

with grid-leak bias developed by the high-resistance path of R8, R10, R12, and R13. When the neon conducts, the positive voltage at NE2/R8 makes VI conduct, operating the T/R relay. The relay contacts put the transmitter on the air and mute the receiver.

the vox circuit, audio signals from the receiver output stage are coupled through dc-blocking capacitor C1 and isolating resistor R1 to D1, the anti-trip (or anti-vox) rectifier. D1 rectifies the audio, and the **positive** dc voltage is filtered by R2 and C2. It is then fed

fig. 2. This vox circuit uses the voltage across e neon lamp to charge up a capacitor; this voltage triggers a dc amplifier that operates the transmit/receive relay.



The time constant of C6, R10, R12, and R13 is purposely long to prevent the T/R relay from dropping out between syllables or words. Should the operator stop talking momentarily, VI returns to its normally saturated condition and the neon doesn't conduct. This would remove the positive potential, turn off V2, and let the T/R relay drop out—except for one thing. Capacitor C6 retains its charge for several seconds since it must discharge through such a high resistance. The grid of V2 is held positive, and the relay in V2's plate circuits is held in.

The setting of **vox delay** control R12 determines the time required for C6 to discharge and turn V2 off. It is adjusted so the transmitter won't go off the air with very brief pauses in speech.

To prevent receiver sound from tripping to R4, the **vox sensitivity** control, through isolating resistor R3.

From R4, the positive voltage—which varies according to the strength of speaker signals picked up by the microphone—is fed to the grid of VI. This grid is already biased positive enough to hold the tube in normal saturation, so the anti-trip voltage merely tends to keep it deeper in saturation; the setting of R4 determines how much deeper than normal.

Negative-going half-cycles coming through the microphone but originating in the loudspeaker are weak, and because of the extra saturation bias from R4 are inadequate to bring the vox amplifier tube out of saturation. The stronger such unwanted receiver signals are, the deeper the anti-trip dc voltage saturates the tube. Therefore, receiver sounds can't trip the relay.

Speaking into the microphone, however, produces voice signals that are not repeated in the anti-trip circuit. Only ordinary saturation bias is present on VI during these signals, so their negative-going half-cycles can operate the tube and hence the relay amplifier that follows.

For manual operation, a push-to-talk switch (S2) on the microphone is used. Function switch S1 is placed in the **mox** position, grounding the grid of VI and disabling the tube. Switch S1 also removes *C6* from the circuit, so it can't store noise pulses and accidentally trigger the relay tube. When the microphone ptt switch is closed, the grid of V2 is grounded, the tube conducts, and the T/R relay operates.

another vox circuit

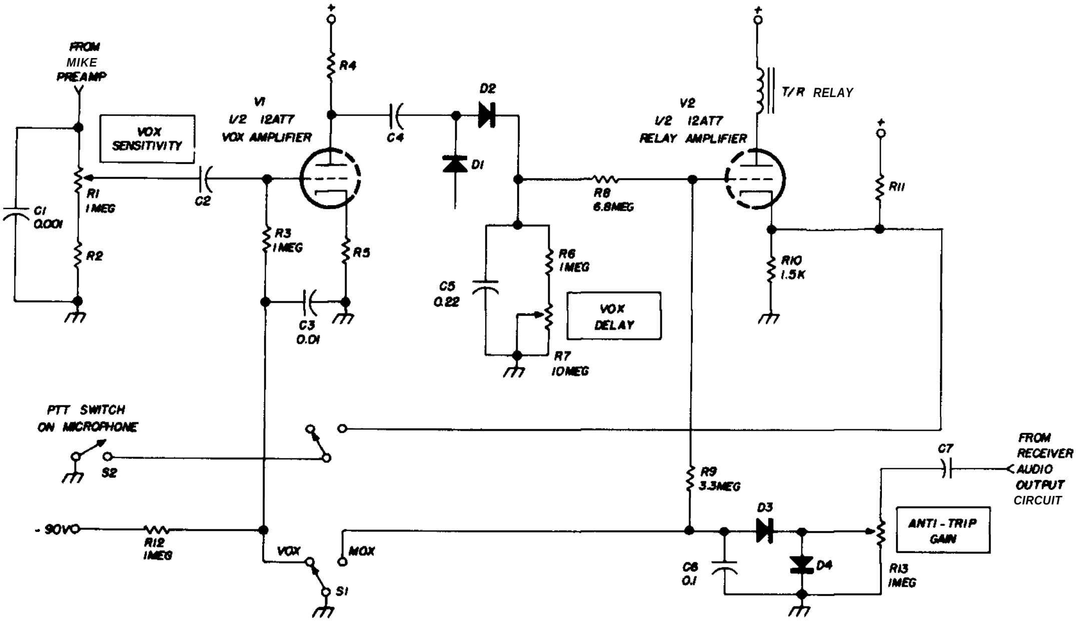
Fig. 3 shows a slightly different way of keying the transmitter by voice. Audio from the microphone preamp is fed to voltage divider R1-R2. (Capacitor C1 grounds out any rf.) Vox sensitivity control R1 taps off some of the voice signal and feeds it to the grid of vox amplifier VI. In the VI plate circuit, voltage doubler D1-D2 rectifies the amplified voice signals to produce a positive dc voltage that is fed to the grid of V2, the relay amplifier.

Relay amplifier V2 is normally held near cutoff by cathode bias applied across resistor R10 through R11 from a positive voltage supply. The positive dc voltage from diodes D1-D2, applied at the grid, turns on V2. It conducts and pulls in the T/R relay, putting the transmitter on the air and muting the receiver.

To hold the transmitter on between normal pauses in speech, a time-delay network follows the voltage doubler: capacitor C5 in parallel with series combination R6-R7. Since R7 (the **vox delay** control) is variable, the time constant can be set to prevent transmitter dropout during brief speech pauses.

In this vox stage, signals from the receiver audio output circuit are fed to R13, the **anti-trip gain** control. Some of the signal is tapped off R13 and rectified by voltage-doubler rectifiers D3 and D4. The resulting negative dc voltage is filtered by C6 and fed through isolating resistor R9 to the grid of relay amplifier V2. Receiver signals that exist in both the microphone **and** speaker circuits

fig. 3. This vox system uses an amplifier and a voltage doubler to develop a positive trigger voltage for the relay tube. Negative voltage from D3 and D4 is used for anti-trip.



produce opposing dc voltages at the grid of V2. The relay does not operate.

Note that a negative 90-volt dc supply is connected (through isolating resistor R12 and grid resistor R3) to the grid of VI. When function switch S1 is in the **vox** position, the bottom of R3 is grounded; the −90 volts does not affect VI. But when S1 is thrown to the **mox** position, the −90 volts biases VI beyond cutoff, disabling the vox stage. The same pole of S1 grounds the output of anti-trip rectifiers D3 and D4 for mox operation, disabling the anti-trip circuit.

The other pole of S1 connects the microphone ptt switch to the cathode of relay amplifier V2. When the ptt switch is pressed, it grounds V2's cathode, unbiasing the tube and turning it on. That operates the relay and switches the transceiver (or transmitter) to transmit.

operating hints

When using a vox system, you can improve results if you'll observe certain rules. Keep the mike away from the receiver loudspeaker, or its volume may be too loud for the anti-trip circuit to handle. The operator's voice must be louder than any speaker signals. It's also a good idea to get the habit of working pretty close to the mike with a vox system.

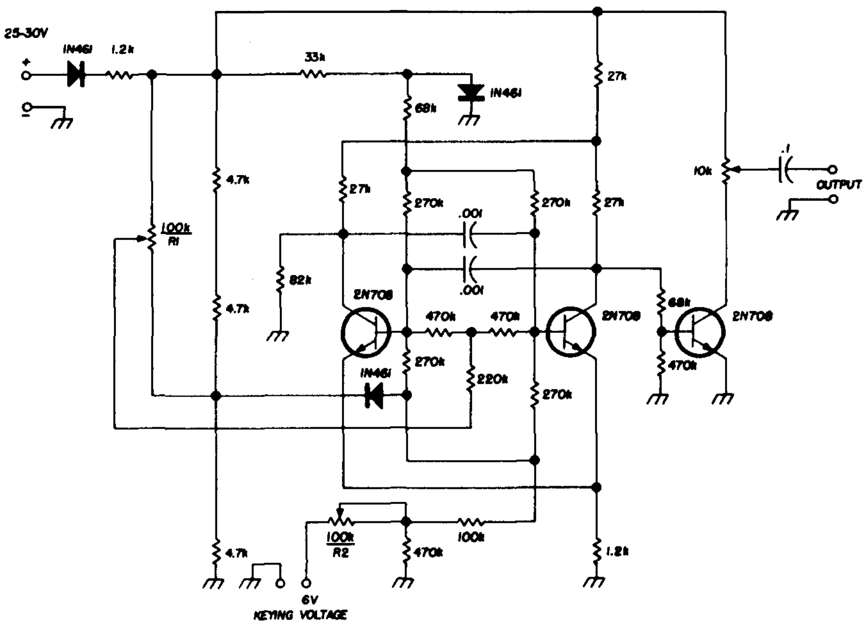
If the vox circuit doesn't work, or if the transmitter chatters on and off erratically, the vox sensitivity and/or vox delay controls may be set wrong. Exact settings depend on the equipment, but the vox sensitivity control should generally be set high. The setting of the delay control depends on how fast you talk and how much pause you want without the transmitter switching.

ham radio

solid-state afsk oscillators

Here are two types of transistorized afsk oscillators which may be built for less than $15. I think they are the best I have seen so far. No special layout is needed with these circuits. The parts I used were all standard: 10% resistors, ceramic disc capacitors, 1N461 diodes (although other types may be used), and 2N708 transistors. I tried 2N1302's,

fig. 1. Afsk oscillator for positive or negative keying voltage. R1 adjusts frequency and R2 adjusts amount of shift.



but they didn't have as good rise and fall times.

With the circuit shown in **fig. 1**, the keying voltage is positive or negative 2 to 10 volts. The supply voltage may be varied from 25 to 30 volts with only 8 Hz change in output frequency. Total drift from turn-on to one hour is about 5 Hz. The circuit shown in perforated printed circuit board and mounted in a 3 x 4 x 5-inch mini box.

To adjust the circuit shown in **fig. 1**, apply voltage and, with no keying voltage, adjust R1 for an output of 2125 or 2975 Hz. Then apply keying voltage and adjust R2 to the other frequency. It may be necessary to see-saw between the values of R1 and R2. To

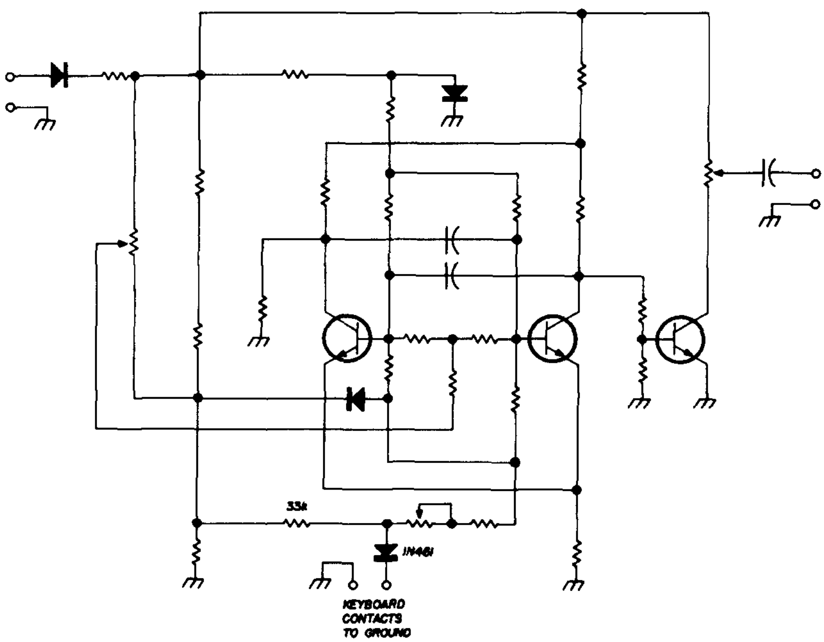
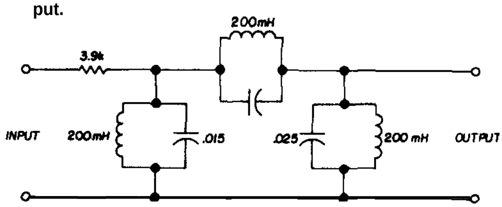


fig. 2. Afsk oscillator designed for keying to ground. Except for the parts that have values, this circuit is identical to the one in fig. 1.

fig. 2 doesn't require a keyed voltage—a key in the ground line shifts the frequency. Supply voltage variations from 25 to 30 volts result in no shift in output frequency, and warmup drift is about 5 Hz. The input current drain with a 30-volt supply is 4.5 mA with both circuits. Output impedance is 3000 ohms.

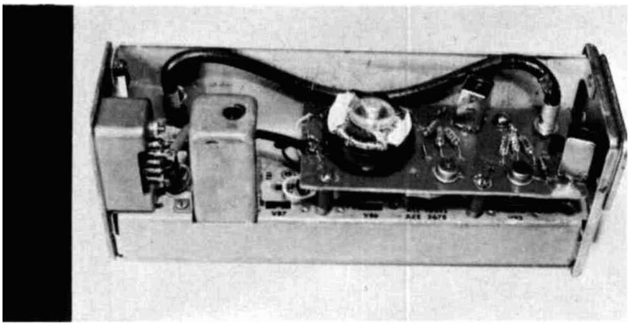
For easy adjustment, use ten-turn potentiometers for R1 and R2. The output is a square wave. If you want a sine wave output, use a band-pass filter as shown in **fig.** 3. The center of this filter is set at 2424 Hz, and is 4'12-dB down at 2125 and 2975 Hz. At the 20-dB points, the pass band is 1500 Hz wide with less than 1% distortion. The complete oscillator and filter were built on a piece of

fig. 3. Band-pass filter for use with the afsk oscillators to provide a sine-wave output.



adjust the circuit in **fig.** 2, ground the keyed input and adjust R1 for mark; remove the ground and adjust R2 for **space.** Either of these circuits may be used for narrow shift by adjusting R1 and R2.

Dale V. Dennis, WA4FGY



using integrated circuits in a narrow-band fm system

Frank Jones, W6AJF, 850 Donner Avenue, Sonoma, California

A number of nbfm VRC-19 receivers have been released to surplus around the country, particularly to operators in the MARS system. Since the local MARS group acquired a large number of these units, we decided to use them in conjunction with a vhf repeater system. However, the original receiver has one very serious disadvantage—the subminiature vacuum tubes it uses are rather short-lived and hard to find. I was selected to convert the units to solid state—here is the approach 1 used in the i-f system.

The photograph shows the end product after modifications to the 455-kHz i-f system of the R-394/U receiver which is part of a surplus VRC-19 or FRC-27 narrow-band vhf unit. The subminiature tubes used in this receiver are difficult to obtain and are usually not long-lived in continuous operation. This system was made by Motorola and is well engineered with good sensitivity for nbfm signal reception. The 455-kHz i-f system has seven subminiature tubes, including two limiter stages. Even though some of the stages are resistance coupled, the unit exhibits amplifications of a few million. Transistorizing this system isn't too easy since it works out of a rather high-loss selective filter with high impedance requirements.

the circuit

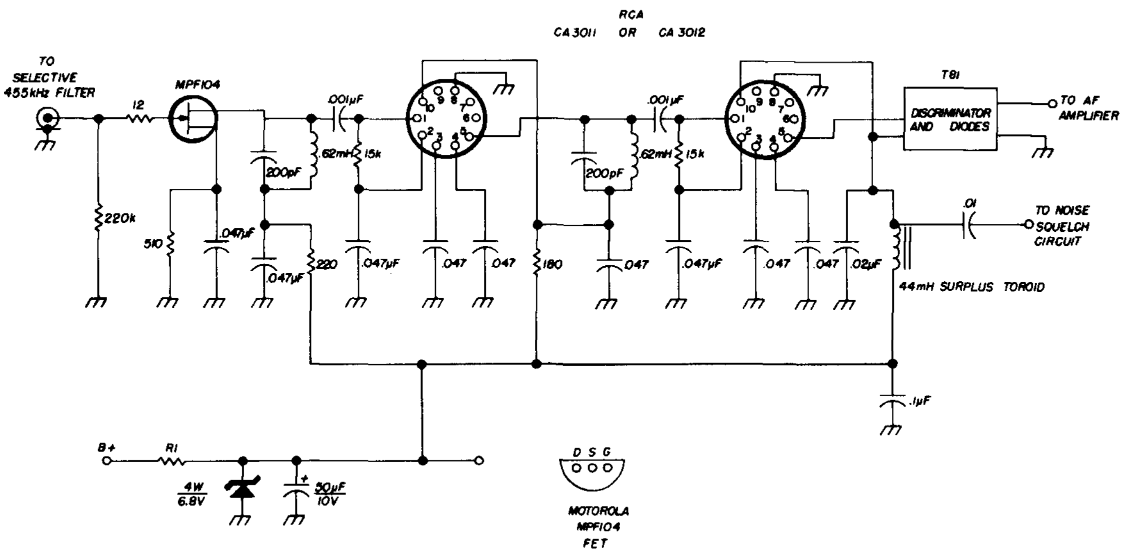
The high gain and limiting requirements can be met by using two integrated circuits in cascade. The RCA CA3011 integrated circuits contain a whole flock of transistors and resistors in one transistor case with ten leads out the bottom. These units are about $1.50 each so they're more economical than equivalent separate transistors. Some large external by-pass capacitors are needed, and the power supply must be limited to less than about 7 volts for safe operation. The current drain with a 6.8-volt supply is about 25 mA for the whole i-f unit.

The IC's not only amplify at nearly any i-f from 100 kHz to 10 MHz, they are excellent limiters for use with a nbfm discriminator. The tuned i-f coils, at 455 kHz in this case, are from small i-f transformers used in transistor broadcast receivers. I didn't use the low-impedance output windings. The discriminator transformer, diodes and small parts are the same ones used in the tube version. However, with a transistor squelch system, it was necessary to put a tuned circuit in the positive lead to the discriminator transformer primary; this tuned circuit is broadly resonant around 5 to 6 kHz in the

The integrated circuits only have a moderately high input resistance, so a field-effect transistor was used to keep loading on the selective circuit system to a minimum. Motorola MPF 104's, 103's or 102's are suitable for this purpose and are all in the one dollar price range. Incidentally, I tried five of these FET's in an i-f system, but the over-all gain was too low and the limiting effects didn't seem to be very good on strong signals.

The CA3011 data sheet indicates 60- to 70-dB gain per IC, but in these experiments I found the gain was more nearly 50 dB (be-

fig. 1. Solid-state 455-kHz i-f amplifier. The value of R1 depends upon the supply voltage: 150 ohms, 1 watt for 12 Vdc; 5000 ohm, 10 W for 145 Vdc.



"hiss" region above the audio cutoff of the audio amplifier.

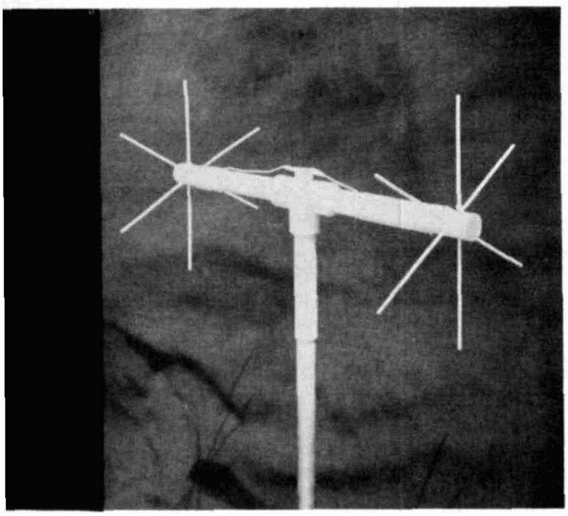
construction

The two IC's and some of the other parts were mounted on a piece of bakelite board about 4 x 1½ x 1/16 inches. This board was mounted on the old tube chassis with three standoff sleeves and screws after all the wiring had been completed. I didn't use any sockets; the ten-lead IC's were mounted on the boards by drilling small holes around the circumference of a small circle. The unit shown in the photograph was strictly experimental since many changes were made while trying different ideas.

fore much limiting takes place), so two IC stages were needed. With limiting on fm signals, the gain is reduced just as with an agc system on a-m signals. The 455-kHz i-f system shown in fig. 1 is not suitable for amplitude-modulated signals, but it's very good for frequency-modulated signals with deviation of about 5 kHz or less.

The original discriminator system and selective filter were designed for 15-kHz signal deviation. Some improvements could be made by replacing these parts with narrowband units, but the cost seemed to be too high to me, so these replacements weren't made.

ham radio



three-band ground plane

A low-cost
high-performance
omnidirectional antenna
for 20, 15 and 10
with low-angle radiation

Fred Brown, W6HPH, Pine Cove, Idyllwild, California

This antenna was inspired by a mobile contact; a solid 20-meter QSO with a Los Angeles station from G3NMR/mobile. This is rather spectacular performance for a mobile in motion down on the street level of London. I was even more impressed when I saw that G3NMR's antenna was only 6 feet long. It was a commercial three-band mobile antenna for 20, 15 and 10 known as the Mark Products HW-3.

Since this antenna worked so well mobile, I thought it would work even better as a home-station antenna, especially if. I made it bigger and put it up in the air. The active length of the HW-3 is about 5 feet; by extending this to a quarter wave on ten meters, I thought it should be possible to eliminate the ten-meter loading coil and improve efficiency on 20 and 15 at the same time.

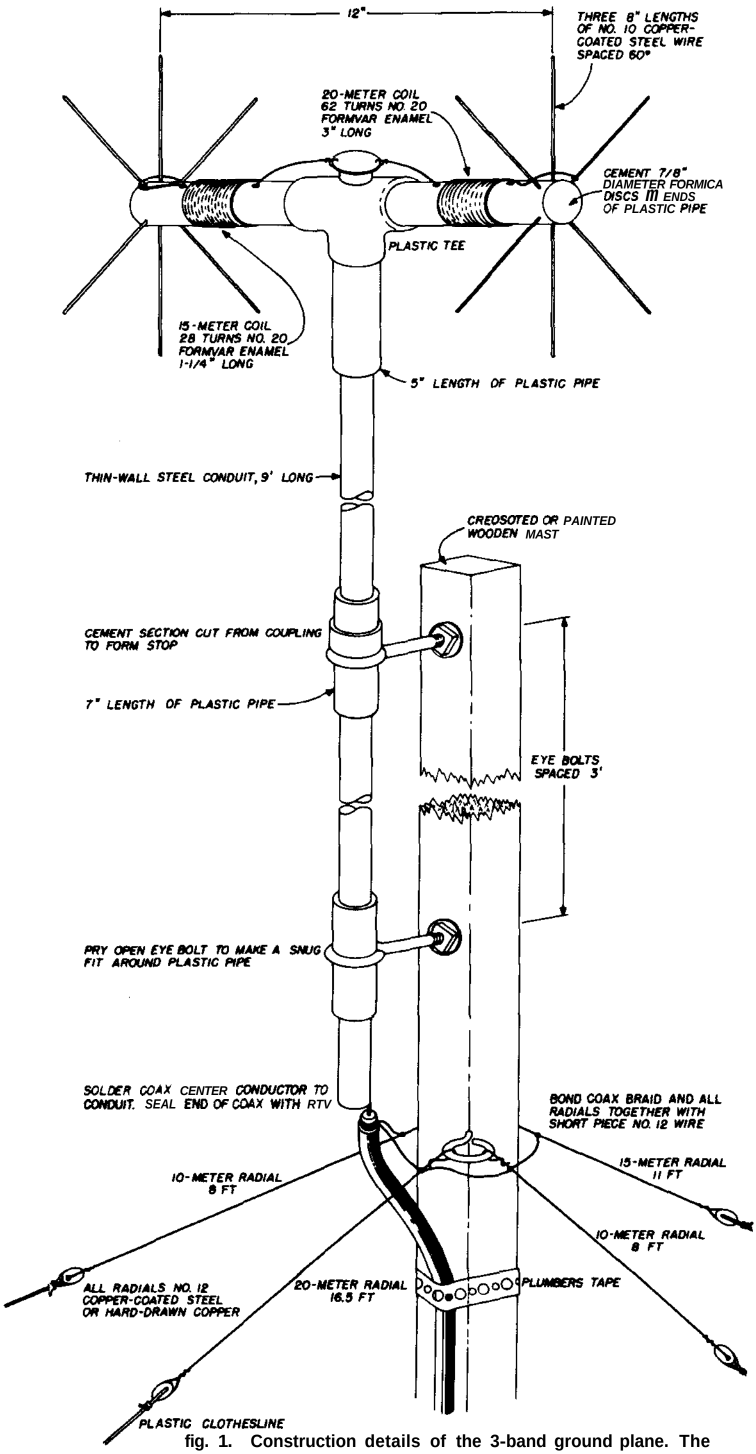


fig. 1. Construction details of the 3-band ground plane. The top tee is bored out so the end of the conduit sticks through; then a copper plate is soldered to the conduit for connecting the loading coils. Coil leads must be soldered to all three spikes.

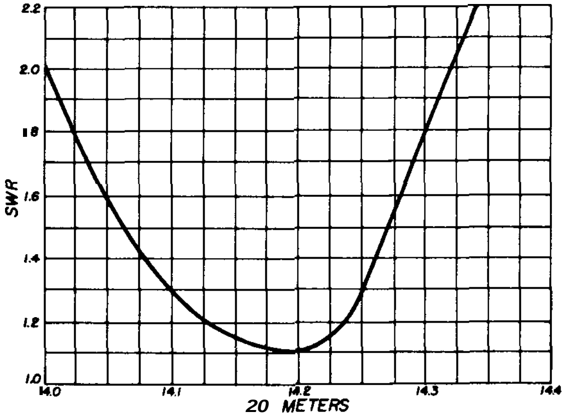
construction

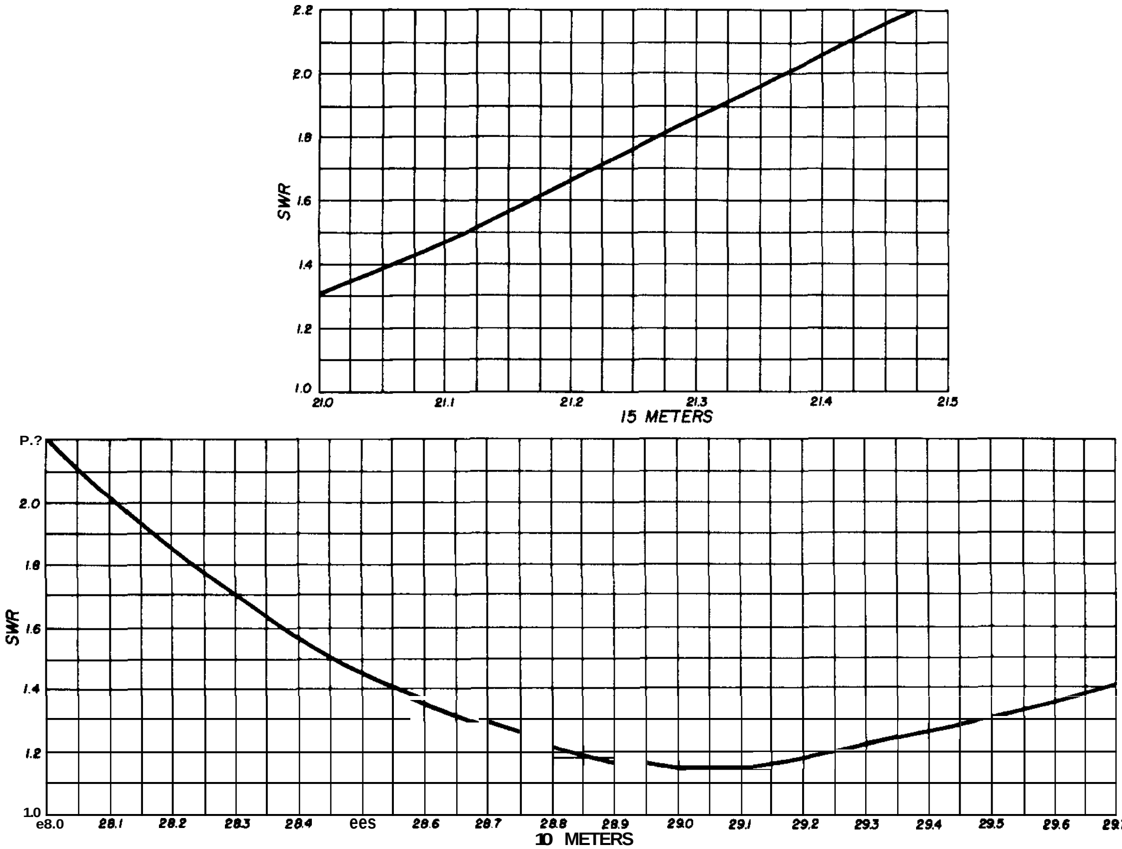
For the monopole, I used ordinary thin-wall steel conduit; it's available in ten-foot lengths from any building supply store. At first I thought the 20- and 15-meter loading coils would act as a slightly capacitive load so the resonant length on ten meters would be somewhat less than 8 feet. After cutting one piece of conduit to 8 feet—then successively shorter lengths—without obtaining any resonance on ten meters, I finally realized that the loading effect is really **inductive!** Therefore, the length should be **more** than a ten-meter quarter wave; a nine-foot length worked out very nicely.

Most of the construction details should be evident from the drawing shown in **fig. 1.** Half-inch plastic water pipe (which is actually 7/8" OD) is a perfect fit over the conduit. You'll need 3 feet of plastic pipe and one tee. The top of the tee is bored out so the conduit sticks through for connections to the loading coils.

The loading coils for 20 and 15 are terminated in capacity hats made from three pieces of number-10 wire forced through the ends of the plastic pipe. This permits loading coils with far less turns than would be the case if they were unterminated and helps reduce coil losses. It also improves the bandwidth since the L-to-C ratio is reduced. The coils would exhibit higher Q if they were larger in diameter, but in the interest of low wind resistance they were wound directly on the plastic **pipe.**

Losses in the twenty-meter coil can also be reduced by usingfewer turns and making the capacity hat twice as large (16-inch spikes instead of 8-inch). This will also improve the swr bandwidth on twenty. However, if you make the 15-meter capacity hat larger, it will fig. 2. Swr performance of the 3-band ground plane.





result in too small a coil which will foul up the ten-meter resonance. In the interest of appearance I made both capacity hats the same size.

I used number-21 Formvar insulated magnet wire for the coils because I happened to have this size on hand, but number 20 or number 18 would be better. The number of turns is quite critical and it's best to start with a few too many turns and remove a half turn at a time until the frequency of minimum swr is near the band center. The inductance can also be trimmed by changing the spacing between turns. There is practically no interaction between 20- and 15-meter resonance.

The resonant frequency will be lowered about 150 kHz when the coils are painted, so it's a good idea to be about that much too high before painting. I used several coats of white spray-can enamel. This holds the turns in place nicely and makes the antenna fairly weather-proof. However, the frequency of minimum swr on twenty meters is still lowered about 100 kHz when the antenna is wet. The conduit is zinc plated but sprayed with clear lacquer as a further rust preventive.

The best way to tune the antenna is with an swr meter. You can make a good counterpoise by spreading four 8- to 10-foot lengths of wire on the ground to make a big X. The monopole can then be suspended over the X with plastic clothesline. This will put the coils only 9 feet off the ground so they can easily be reached with a step ladder. Solder the center conductor of the coax to the base of the monopole and the outer conductor to the junction of the X. Alternatively, you can use a car body or other non-resonant object as the counter-poise.

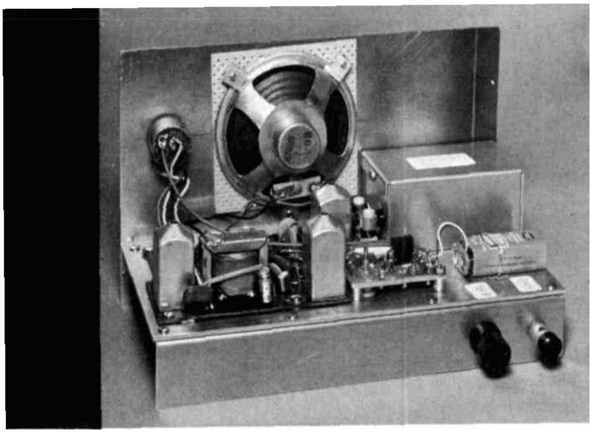
In its final form, the antenna has four radials which also serve as guy wires, forming a drooping ground plane. Two of the radials are resonant on ten, and these run in opposite directions. One is resonant on 15 and one on 20. I found it important to have at least one radial resonant on each band to prevent rf current from flowing down the outside of the coax. The relative values of rf current in the radials and on the outside of the coax can be checked with the rf current probe described in ham notebook on page 76. So long as the current in at least one of the radials is more than four times the current measured on the coax, radiation from the coax will be negligible.

results

The final swr-vs-frequency plots are shown in fig. 2. Bandwidth is more than adequate on 10 and 15 meters, but is just barely good enough for covering twenty with an swr below 2.0:1. These measurements were made through 45 feet of RG8/U, but essentially identical results were obtained with a short piece of coax and the antenna mounted over a counterpoise on the ground. The swr curve for 15 meters would indicate that perhaps the 15-meter loading coil has a half turn too many; but 15-meter resonance is so broad I haven't bothered to change it.

This antenna will not equal the performance of a good full-sized rotary beam, but it will give excellent low-radiation-angle omnidirectional coverage on 20, 15 and 10. Field-strength measurements and flattering signal reports received from DX stations around the world indicate the antenna does all that could possibly be expected of it. Total cost of materials, neglecting the mast and coax, was under $4.00. For this price, you'll find it hard to beat.

ham radio



low-frequency weather receiver

A weather receiver is a pretty useful addition to an anemometer and a wind-direction indicator if you want to know what's happening on the weather front

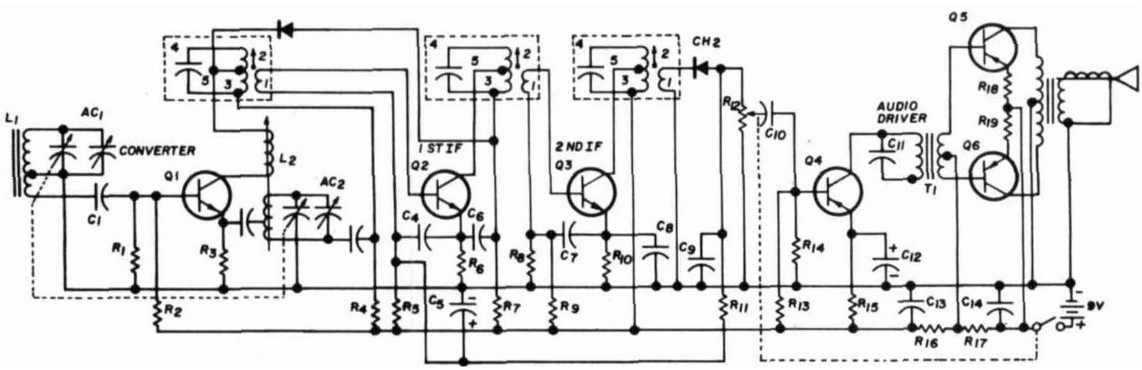
Hank Olson, W6GXN, P. O. Box 339, Menlo Pa[illegible]lifornia 94025

In addition to having an anemometer and a wind-direction indicator, it's helpful to hear what the aircraft weather broadcasts have to say. In metropolitan areas near large commercial airports continuous aircraft weather predictions are broadcast just below the BC band. For instance, in my area, the weather is put out from Oakland on 362 kHz. You can tune in 362 kHz any time of the day or night and listen for a few minutes while a prerecorded tape plays back the latest weather report.

These weather reports and forecasts are broadcast over the communications facilities of the Federal Aviation Agency. Although this weather information is prepared and selected with the pilot in mind, the broadcasts are popular with amateur weathermen and members of the general public who are planning outdoor activities. These aviation weather broadcasts are perhaps best described as **continuous** and **scheduled.**

Scheduled weather broadcasts are made twice each hour: at 15 and 45 minutes past the hour. At 15 minutes past the hour, current weather reports for eight to twelve locations within 150 miles of the broadcasting station are transmitted. At 45 minutes past the hour, similar reports are broadcast for important cities and airports within a 400-mile radius. Scheduled broadcasts include advance warnings of potentially hazardous weather, such as squall lines, thunderstorms, fog, icing and turbulence. If available, weather radar reports are also given.

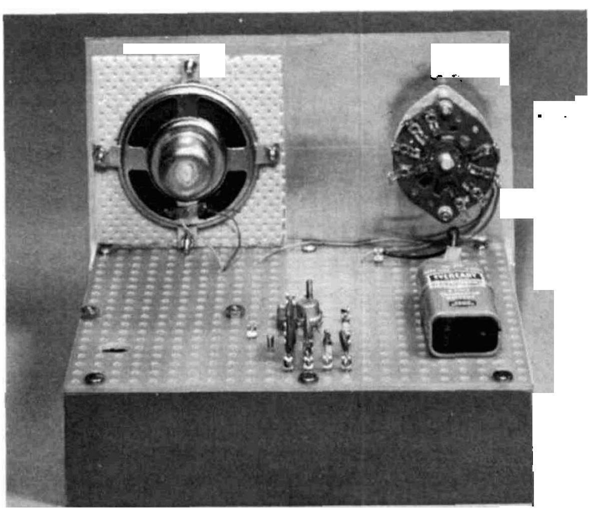
Continuous transcribed weather broadcasts include a radius of 250 miles from the broadcasting station. These broadcasts include a general forecast of conditions over a broad area, aviation weather including upper-level winds and any warnings, as well as pilot and radar reports. In addition, current weather reports for eight to fifteen locations within the area of the broadcasting station are transmitted. Material is updated as new reports come in and revised forecasts are issued. These items are repeated to provide a continuous weather service.



AC1	190-pF variable capacitor with trimmer	R1	6.8k, 1/4 watt
AC2	90-pF variable capacitor with trimmer ganged with AC1	R2	27k, 1/4 watt
C1	.02 μF	R3	1.5k, 1/4 watt
C2, C3	.01 μF	R4	470
		L1	435 μH loopstick antenna
		Q1	2N194A

fig. 1. Typical six-transistor broadcast receiver.

Top view of the modified broadcast raceiver showing the new local-oscillator stage, speaker and battery.



receiving the broadcasts

The BC453, BC1206 or any one of several other low- and medium-frequency surplus receivers can be used for this purpose. However, unless you just happen to have one in the rubble-heap, a simpler approach can be taken. An ordinary transistor radio (you know, the kind every teen-ager carries with him constantly) can be easily modified for reception a few hundred kilohertz below the BC band.

The most direct method of modification is simply to pad the two sections of the tuning capacitor so that some point on the dial corresponds to the frequency of your local weather station. This approach has been only partially successful in several BC transistor radibs that I have worked on.

The problem lies in the nature of the converter used in most BC sets. Fig. 1 shows a typical BC converter (an autodyne), where one transistor is operated as both a mixer and

fig. 2. Modified converter stage of a six-transistor broadcast receivar to cover 362-kHz FAA weather broadcasts.

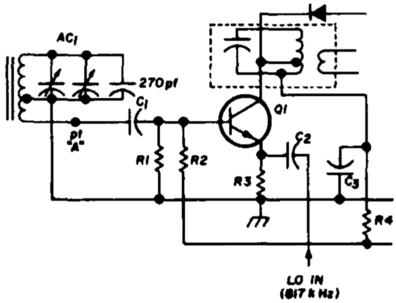
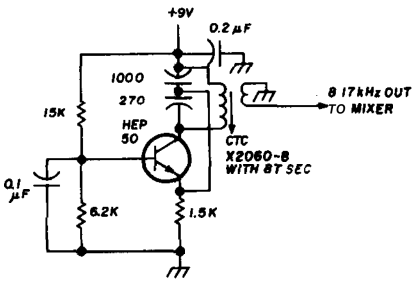


fig. 3. New oscillator stage for the modified broadcast receiver.



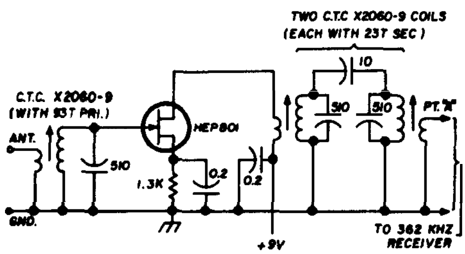
local oscillator. Tinkering with an autodyne led to instability in at least one case where I attempted to change the frequency of the receiver to a weather channel. It is my feeling that circuits such as autodynes which perform dual functions are inherently "fussy" and not amenable to casual modification.

By removing the oscillator coil (L2) from the circuit, the autodyne can easily be converted into a simple grounded-emitter mixer as shown in **fig.** 2. Local-oscillator injection must then be furnished by a separate oscillator. The local-oscillator drive should preferably come from a small link on the oscillator coil because it is through the low impedance of this link to ground that the emitter of the mixer is bypassed for the signal frequency.

The new oscillator circuit is shown in **fig.** 3. The finished weather receiver is shown in the photo. Note that the original miniature volume control has been replaced with a standard type on the front panel as has the on-off switch. The dual tuning capacitor is now adjusted with a screwdriver to peak the mixer input tuning. The local-oscillator frequency is adjusted by the slug in the local-oscillator tuning coil.

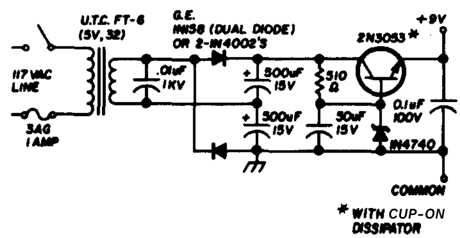
A word about old transistor BC radios is in order. The majority are discarded (and available inexpensively) because of one of two electrical faults: autodyne trouble, or electrolytic capacitor trouble, or both. Since this modification calls for destruction of autodyne action with a separate local oscillator, **that** fault is automatically cured. By replacing the four or five electrolytic capacitors in the circuit with tantalytic types, **that** fault can be eliminated.

fig. 4. Rf amplifier for the 362-kHz weather receiver.



The Kemet "C" or Sprague 150D series tantalytics are satisfactory for this replacement. Of course, a lot of transistor radios are junked for mechanical damage, battery-leakage corrosion and accidental circuit damage from reversing the battery polarity. This

fig. 5. Ac power supply for the weather receiver.



bunch may be harder to work with; but glue, sodium bicarbonate, and a few new transistors will usually do the job.

In my particular area the local weather station is about twenty miles away. This station (Oakland, 362kHz) provides weather reports for communities up to 300 miles away. If you live in an area which is not within the weather-reporting area, you can add an rf amplifier with an antenna and ground terminal as shown in fig. 4. The loopstick antenna is replaced with an rf interstage coupling circuit; a second weather receiver using an rf stage is shown in the lead photo. A complete list of FAA weather broadcasting stations within the United States is listed in table 1.

Of course, an external ac power supply can be built to power the weather receiver. The circuit of the one I built is shown in fig. 5. A simpler design would probably do, but I favor regulated supplies and this unit was built from "on-hand" and very inexpensive parts.

table 1. Continuous and scheduled aviation weather broadcasts on 200- to 400-kHz FAA radio facilities.

location	frequency (kHz)	type
Alaska		
Anchorage	338	S
Aniak	359	S
Annette Is.	266	S
Bethel	251	S
Bettles	391	S
Big Delta	347	S
Cold Bay	341	S
Fairbanks	260	S
Farewell	206	S
Galena	371	S
Gulkana	248	S
Gustavus	236	S
Homer	320	S
Illiamna	233	S
Kenai	379	S
King Salmon	400	S
Kodiak	394	S
Kotzebue	356	S
McGrath	350	S
Middleton Is.	260	S
Minchumina	227	S
Moses Point	263	S
Nenana	236	S
Nome	239	S
Northway	400	S
Petersburg	359	S
Skwentna	269	S
Summit	326	S
Talkeetna	305	S
Tanana	212	S
Unalakleet	382	S
Yakataga	209	S
Yakatat	385	S
Alabama		
Anniston	278	S
Birmingham	224	C
Mobile	248	S
Arizona		
Phoenix	326	C
Tucson	338	C
Arkansas		
Ft. Smith	223	C
Little Rock	353	C
California		
Blythe	251	C
Fresno	344	C
Los Angeles	332	C
Montague	382	S
Oakland	362	C
Redbluff	338	C
Colorado		
Denver (Englewood)	379	C
Trinidad	329	C
Connecticut		
Hartford	329	S
D. C.		
Washington	332	C
Florida		
Ft. Myers	341	S
Jacksonville	344	C
Key West	332	S
Miami	365	C
Pensacola	326	C
Tallahassee	379	C
Tampa	388	C
Georgia		
Atlanta	266	C
Hawaii		
Hilo	396	S
Idaho		
Boise	359	C
Burley	272	S
Idaho Falls	350	C
Illinois		
Carbondale	248	S
Chicago	350	C
Indiana		
Goshen	215	S
Indianapolis	266	C
Iowa		
Davenport	224	S

Kansas

Garden City	257	C
Wichita	332	C

Kentucky

Louisville	359	S
Paducah	356	S

Louisiana

Grand Isle	236	C
Lafayette	375	S
Lake Charles	242	S
Shreveport	230	C

Maine

Augusta	221	S
Bangor	239	S
Millinocket	344	C

Massachusetts

Boston	382	C
Westfield	230	S

Michigan

Detroit	388	C
Houghton	227	C
Sault Ste. Marie	400	C
Traverse City	365	C

Minnesota

Duluth	379	C
Minneapolis	266	C

Missouri

Kansas City	359	C
St. Louis	338	C
Springfield	254	C

Montana

Billings	400	C
Bozeman	329	C
Cut Bank	263	S
Dillon	379	S
Great Falls	371	C
Lewiston	353	S
Livingston	224	S
Miles City	320	C
Missoula	308	C

"May-day, may-day"

Nebraska

Chadron	275	S
Imperial	353	S
North **Platte**	224	C
Omaha	320	C
Scottsbluff	341	S

Nevada

Las Vegas	206	C
Reno	254	S

New Jersey

Millville	363	
Newark	379	

New Mexico

Albuquerque	230	
Roswell	305	

New York

Buffalo	260	
Elmira	375	
Syracuse	350	

North Carolina

New Bern	379	
Raleigh	350	

North Dakota

Bismarck	230	S
Fargo	365	S
Minot	209	S

Ohio

Cincinnati	335	C
Cleveland	344	C
Columbus	391	S
Findlay	209	S
Zanesville	275	S

Oklahoma

Ardmore	400	S
Oklahoma City	350	C
Ponca City	368	S

Oregon

Medford	263	S
Pendleton	341	C
Portland	332	C
Redmond	368	C

Pennsylvania

Allentown	321	S
Bradford	224	S
DuBois	356	S

Harrisburg	338	S
Pittsburgh	254	C
South Carolina		
Charleston	329	C
Spartanburg	248	C
South Dakota		
Huron	391	S
Pierre	347	S
Rapid City	254	C
Sioux Falls	245	S
Tennessee		
Knoxville	257	C
Memphis	371	S
Nashville	304	C
Texas		
Amarillo	251	C
Big Spring	326	C
Brownsville	388	S
Corpus Christi	382	S
El **Paso**	242	C
Ft. Worth	365	C
Galveston	206	C
Mineral Wells	284	S
Tank	388	S
Tyler	320	S
Utah		
Delta	212	C
Ogden	263	C
Vermont		
Burlington	323	C
Virginia		
Pulaski	272	S
Richmond	260	S
Roanoke	371	C
Washington		
Seattle	362	C
Spokane	365	C
Walla Walla	356	S
West Virginia		
Elkins	245	S
Huntington	323	S
Wisconsin		
LaCrosse	371	S
Milwaukee	242	C
Wyoming		
Casper	269	C
Rock Springs	290	C
Sinclair	368	S

• S Scheduled broadcasts
C Continuous broadcasts

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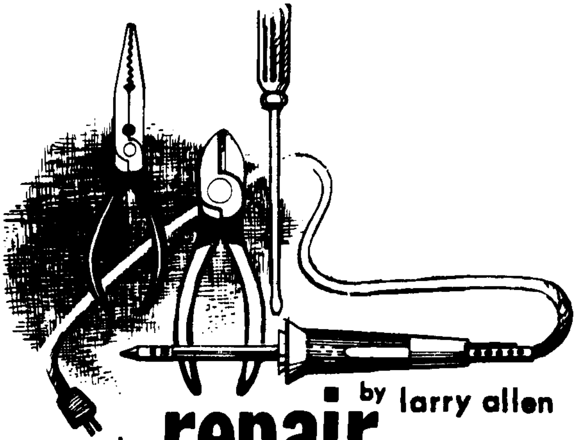


by larry allen

the repair bench

troubleshooting around fet's

Several articles about field-effect transistors have appeared in **ham radio,** and commercial equipment is beginning to show up with FET's in some circuits. Sooner or later, you'll build or buy a piece of new equipment using this special breed of transistor, and eventually you'll have to troubleshoot it. So, you ought to know a few of the idiosyncracies FET's display.

They're different from ordinary transistors. In some respects, they may remind you of tubes. But they **are** transistors and if you have one go bad in a receiver or something, you troubleshoot it much the same as any other transistor. Before I mention FET troubles and how to spot them, it seems like a good idea to recap how they differ from ordinary transistors.

two main types

Basically, a FET (rhyme it with "jet") starts with a **channel** of semiconductor material through which current can flow. Take a look at **fig. 1A.** The bar is N-type semiconductor material. Current flows in the direction of the arrows, carried through the N-material by electrons. The end of the channel the current enters is the **source;** the end from which it leaves is the **drain.**

Into the bar of N-material are diffused two spots of P-material. Connected together as in **fig. 1B,** they form a **gate.** The gate material makes a junction with the channel material, and the whole device is called a junction field-effect transistor, or just **JFET.**

Current through the channel in **fig. 1A** depends on the resistivity of the material and the voltage of the battery. In **fig. 1B,** however, the current is controlled by the negative voltage applied to the gate. Battery 2 is the gate-bias supply and reverse-biases the PN junction: negative voltage is applied to this P-material, and positive voltage would be applied to N-material. Reverse bias on the gate elements tends to **pinch** off the flow of current through the channel from source to drain. The higher the negative gate-bias, the less the current from source to drain. Naturally, if you override part of the battery-2 voltage with a signal, the signal makes corresponding fluctuations in channel current. (This is the way a grid controls current in a vacuum tube.)

The JFET is only one type of field-effect transistor. The other type is called **MOSFET,** an acronym for metal-oxide-semiconductor field-effect transistor. In a MOSFET (**fig. 2A**), the gate doesn't form a junction . . . at least not like in a JFET. The gate is a metal electrode, and between it and the channel is an insulating coating of oxide. The channel itself (an N-channel is the one shown) is surrounded by opposite-type material, called the **base—or** sometimes, substrate.

In a JFET, there may be slight leakage

across the gate-to-channel junction; there's none at all in a MOSFET, because the gate is entirely insulated by the oxide layer. Another name for the MOSFET is IGFET, for insulated-gate FET. The JFET has a high input impedance because a reverse-biased PN junction is the input element between gate and source. The MOSFET input exhibits almost infinite impedance since the gate cannot draw any current at all.

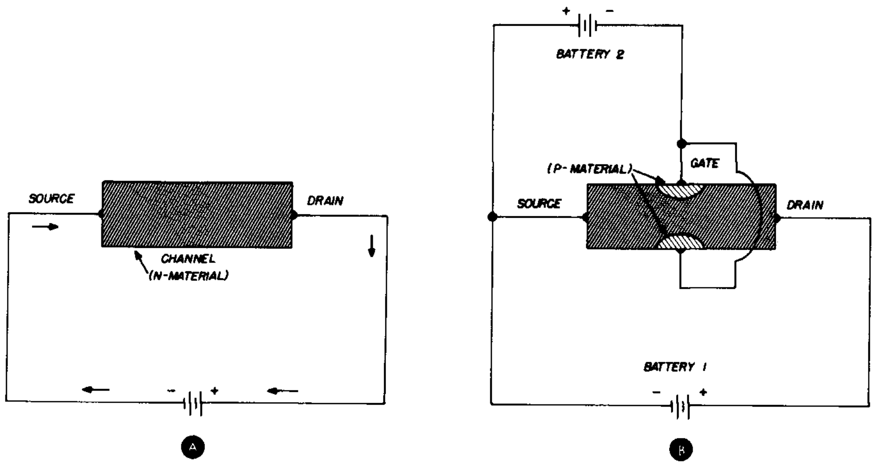
The MOSFET channel current is pinched off by a reverse gate bias. This action in the channel is called **depletion,** since the bias depletes all current carriers in the vicinity of the gate elements. A depletion MOSFET operates very much like a JFET. Its channel current is highest when there is no bias. Increasing the reverse bias cuts down the current in the channel. Enough bias drives the channel into cutoff—or, as it is called in a FET, into pinch-off.

The gate bias in a JFET cannot be permitted to get near the forward-bias mode. If it did, positive half-cycles of an input signal would make the junction draw current, and distortion would result. A depletion MOSFET doesn't draw gate current under these conditions, and can accept some forward bias. Nevertheless, it works best with some reverse bias.

Another type of MOSFET works only with forward bias. (In that respect, it resembles a bipolar transistor.) The structure of this MOSFET is shown in **fig. 2B.** The source and drain are separate, and the channel doesn't exist at all until a strong forward bias is applied. With no bias, there is no current from source to drain. As forward gate bias is applied, the substrate or base material changes character. The base is P-material, as it would be in an ordinary N-channel MOSFET; the section beneath the gate changes to N-material under the strong electric field set up by the gate. That section becomes the N-channel for this MOSFET. Because gate bias therefore enhances current flow instead of depleting it, the unit is called an **enhancement** MOSFET.

In operation, then, you find the JFET can be operated only with depletion bias and the MOSFET that can be designed for depletion or enhancement operation. Most depletion MOSFET's can work slightly into the enhancement mode, too. A circuit for this kind of MOSFET is shown in **fig. 2C.** Drain voltage is applied through a load resistor. Gate bias, a reverse bias in this depletion MOSFET, is developed across a resistor in the source lead (the way a cathode resistor develops bias for a vacuum tube).

fig. 1. Formation of n-channel junction field-effect transistor (JFET); semiconductor channel conducts electrons. With a gate element diffused into the channel material (B), voltage of battery 2 determines current flow from source to drain.



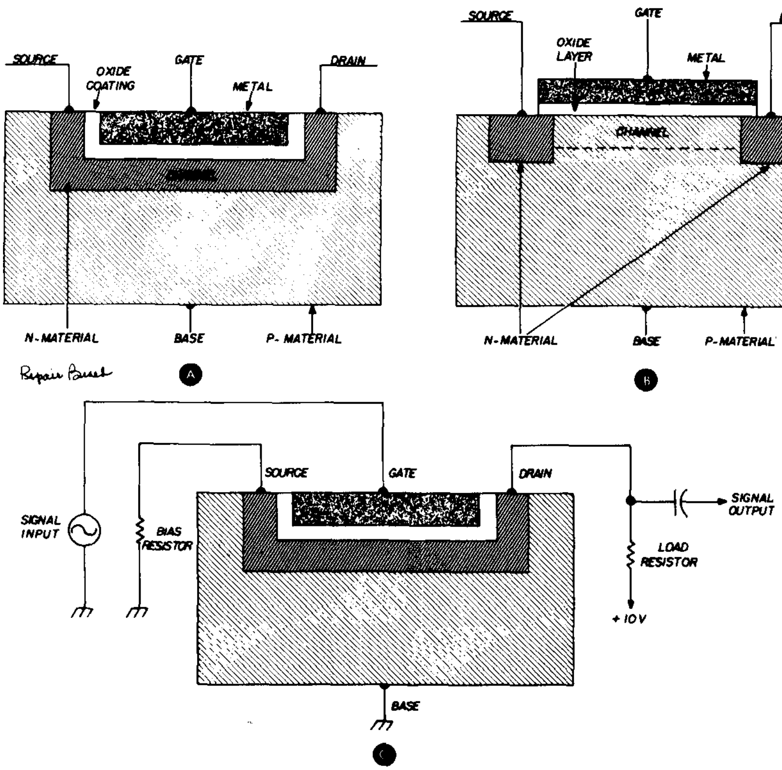
bipolar—unipolar

Terminology sometimes gets in the way of understanding. Special words fog up the very things they're coined to simplify. Conventional transistors, the ordinary junction kind that you find in profusion nowadays, are lately called **bipolar.** There's a reason.

Current flows in semiconductors by means of two kinds of carriers: holes and electrons. In N-material, the chief carriers (called majority carriers) are electrons; in P-material, the majority carriers are holes. In all ordinary junction transistors, the minority carriers contribute considerably to operation. Therefore, the transistors are said to be bipolar.

In FETs, the flow of current is not across junctions, but through the channel—all of which is (or becomes) one type of semiconductor material. Carriers, therefore, are of only one character. In N-channel FETs, the carriers are electrons; in P-channel types, the carriers are holes. FET's are therefore said to be **unipolar.**

fig. 2. Structure and operation of n-channel MOSFET. Oxide coating in A insulates gate from channel; base (substrate) is p-material. Enhancement type in B separates drain and source; channel forms in substrate under stress of bias voltage. Signal is fed to gate in C; bias in this case is from resistor; signal is taken from drain.



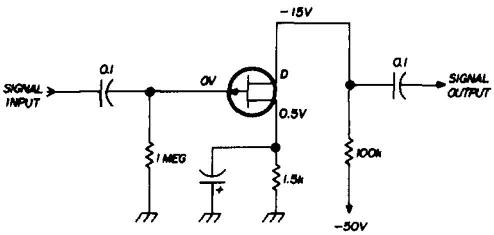
If you understand the operation of FET's, as I've already described it, the two terms have little meaning one way or the other. Nevertheless, you should know—if only for reference—that bipolar transistors are the ordinary kind and that unipolar transistors are a family to which FETs belong.

what happens to fet's?

Since FETs are so new, not much data has been gathered on their breakdown habits. Yet, enough is known to make it pretty simple to troubleshoot circuits that use them, and also to test them outside the circuit.

The most common fault in a FET is a short between gate and channel. In a JFET, forward bias for too long a time might damage the junction, or the reverse-voltage breakdown rating might be exceeded with the same result. Either way, the gate can no longer control channel current, and too much gate current flows. In a typical JFET, gate current should be only a few nanoamps; in a MOSFET, there should be none. Gate breakdown in a MOSFET is caused by over-

fig. 3. Simple audio amplifier using a p-channel JFET.



voltage in either direction, forward or reverse. The result is gate current, of which there would otherwise be none.

Excessive gate current in a JFET, whatever the cause, instead of shorting the junction, may burn it open. The effect is to prevent control of channel current although no gate overcurrent is noticeable. A MOSFET gate may open up, with a similar result, although excess gate current won't be the cause.

An open channel can develop in either kind of FET. It is usually caused by too much voltage between source and drain. The overcurrent that results just burns the channel open, usually at the narrow part near the gate. Afterward, no current can flow from source to drain. Depending on where the open is, there may or may not be slight gate current in a JFET that has an open channel; of course, there is never any gate current in a MOSFET, anyway.

The chief thing to guard against, then, is obviously overvoltage. If you're experimenting with FETs, pay attention to the voltage ratings. Next most detrimental is overcurrent. Even slight overcurrent may eventually overheat a JFET junction, resulting in permanent damage.

FETs are delicate in some respects. Even static that builds up on your body can blow the gate junction or gate insulation of a small-signal FET. Because of this, new FETs are "zot-proofed" by the simple expedient of twisting their leads together. You can keep them safe while you handle or install them by sliding a ring of solder or soft wire around all three or four leads. Bend the leads slightly outward to hold the ring in place. After you have the FET installed, clip or melt off the zot-ring.

finding faulty fet's

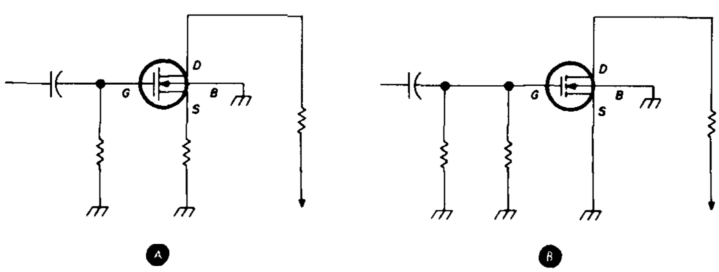
There are three ways to track down a FET that has given up. One is by signal injection or signal tracing—methods I discussed in the April and May 1968 issues. You simply pin down the faulty stage and then test it to see if the FET in that stage is at fault.

A second way is by checking voltages in the circuit with a high-impedance voltmeter. (A vom won't do; it loads down the circuit too much.) Once you know how a FET fails, you can 'figure out the dc voltage changes that result. Thus, measuring circuit voltages is another means of troubleshooting FET stages.

The third way is by removing the FETs from their circuits and testing them, one at a time. If you choose this method, take care to zot-proof the FET before you take it out of the printed board or socket. Also, use heat-sink pliers to unsolder and resolder; FETs are heat-sensitive, too.

You can see one ordinary FET circuit in fig. 3. It's an audio amplifier, using a P-

fig. 4. MOSFET supply circuits. Depletion device in A unit gets its bias from a source resistor. Enhancement type in B is biased by resistive voltage divider network fad from regular dc supply.



channel JFET (you can tell it's P-channel because the arrow points outward). Gate bias is developed in the source circuit, across the 1.5k resistor.

Suppose you are measuring voltages in this circuit and the drain voltage is –48 volts on your meter. Obviously, the FET isn't dropping supply voltage across the 100k resistor. You'd also find almost no voltage at the source terminal.

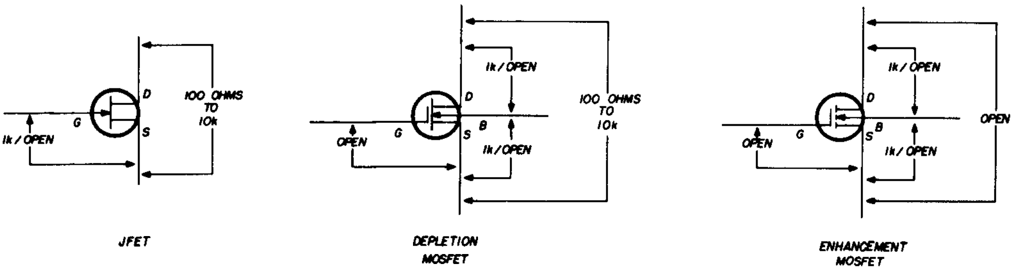
Suppose there is a very slight negative voltage at the gate terminal. You know that either the input coupling capacitor is leaky or the FET has gate-to-channel leakage. If the leakage is in the capacitor, and persists, the FET could be damaged. If in the FET, replace it. To find out which, disconnect the capacitor; if the voltage returns to zero, the capacitor was at fault.

Suppose the drain voltage is low . . . say, only –5 volts. The 100k resistor could have increased in value, but the FET is more likely drawing too much current. If so, the source-terminal voltage should be high. With the bias thus high, the gate should theoretically keep the FET from drawing much more than normal current; it might be open and having no effect. Finally, with drain voltage low, you might find source bias removed entirely or partially—probably the result of a shorted bypass capacitor or 1.5k resistor.

All those symptoms you could reason out for yourself, knowing the nature of FET breakdowns as you now do. Faults in MOSFET circuits present slightly different symptoms, but only because of the difference in structure. A simple MOSFET amplifier stage is shown in **fig.** 4A.

If you measured too much voltage at the drain of this N-channel MOSFET, it would likely be because the channel is open or the gate is overbiased. If it happened to be an enhancement-type MOSFET **(fig. 4B)**, even zero bias might cause the lack of current through the channel; you recall that an enhancement MOSFET needs forward bias to cause curent flow. Zero bias in either kind could be due to an open gate element. Only an enhancement MOSFET would need the bias arrangement in **fig. 4B;** either resistor might upset bias. A gate-to-channel short near the source end of the channel would eliminate bias, as would a gate-to-base short. A gate-to-channel short nearer the drain might cause excess current through the drain-supply resistor and thus lower the voltage measurable at the drain terminal.

fig. 5. Resistance readings you can expect with different types of n-channel FET's. Readings are the same for p-channel types except that the "diode" readings are reversed.



These are far from the only possibilities. Any incorrect dc voltage in a FET circuit can be reasoned out. You may find it quicker, though, merely to determine that the trouble is not in an associated component. Then you can remove the FET from the circuit for testing.

There aren't many FET-checkers around yet, but your trusty ohmmeter is an excellent tester provided its battery voltage isn't too high. The 1.5-volt type is best.

Diagrams that will help you check different FET types are given in **fig. 5.** The readings you should get are indicated clearly. Each type has it own peculiarities.

The source-to-drain readings of the JFET and the regular MOSFET are similar. You can connect your ohmmeter leads in either polarity; the normal reading lies between 100 ohms and 10k, depending on the particular FET. Source-to-drain resistance of an enhancement-type MOSFET is something else; because of the way the source and drain junctions are applied to the channel, they act as back-to-back diodes, and therefore you read an open between the two.

Both MOSFET's show open circuits from the gate to any other element, since the gate is insulated. The JFET, from gate to source, reads like a diode: about 1k in one direction and open in the other. A low backward reading indicates leakage. Open both ways means the gate is open. The JFET shown is an N-channel; the only difference in a P-channel JFET is in the polarity of the gate-to-source readings—they are just reversed.

The drain-to-base and the source-to-base resistance of both MOSFET's are similar in that they measure like diodes: about l k in one direction and open in the other. Which direction is which depends on whether the MOSFET is N- or P-channel. All types act as back-to-back diodes, though. If the negative ohmmeter lead is on the base of a particular MOSFET and the source checks about lk, so should the drain. If it checks open in that direction, it should check open to the drain also.

little oddities

There you have the basics of checking most FET's. Not that those shown are the only types—they're not. All those I've talked about this month are symmetrical types. That is, their gates are about half-way between source and drain. In most of them, drain and source are interchangeable. You can connect either end to function as either source or drain.

Nonsymmetrical types have the gate located elsewhere on the channel, near one end or the other. Ordinarily, the gate is placed near the source. Any faults that occur between gate and channel are likely to tie gate and source more closely together. The difference in result, as far as circuit voltages are concerned, may be noticeable. However, the resistance tests shown for symmetrical FET's are equally applicable to nonsymmetrical types.

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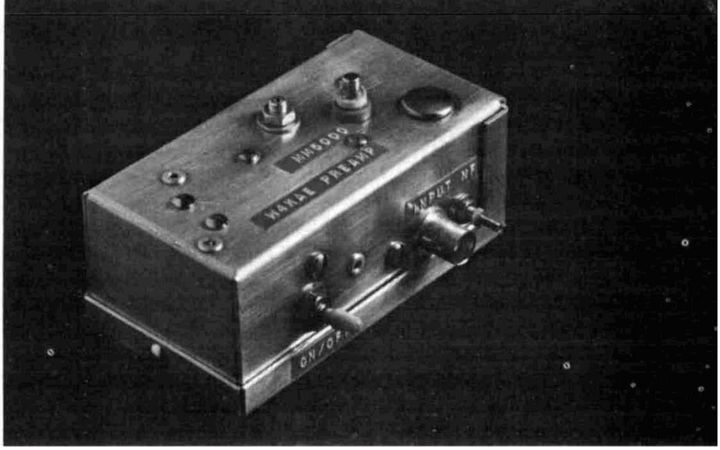


MM5000
2-meter preamp

Is this the last of the bipolar transistor preamps for vhf?

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky 40505

Here is a **two-meter** preamp that may be the last bipolar transistor design taken seriously by vhf amateurs. Motorola claims 1.6 dB maximum noise figure at 200 MHz for the MM5000; at 150 MHz or two meters I suspect we can expect 1.2 dB maximum. I have no sophisticated noise equipment at my station but I can hear the difference in noise between this transistor and an AF-239 (2.2 dB, typical). Hearing the difference should tell the difference!

I became interested in building preamps for fellow amateurs several years ago when I started offering 416B 432-MHz preamps. Little did I know that it would soon be obsolete because of the influx of new, low-noise semiconductors. Today, hams can buy low cost MOSFETS, JFETS and super low-noise bipolar—to make their own equipment. Except for noise figure, the MM5000 PNP germanium mesa bipolar transistor is a lot like many others; it overloads rather easily, but burnout in this circuit is unknown with my 500-watt linear amplifier—and all I use is a Dow-Key spdt coaxial relay.

If you have a quiet location away from commercial two-way equipment, it's possible to use this low-noise transistor to its fullest. As I said before, no burnout problems were observed. However, if you expect trouble because of high swr, try Microwave Associates MA-4850 Schottky-barrier diodes instead of 1N916's used here. The Schottky's are rated at 2 watts each! Price, $1.50.

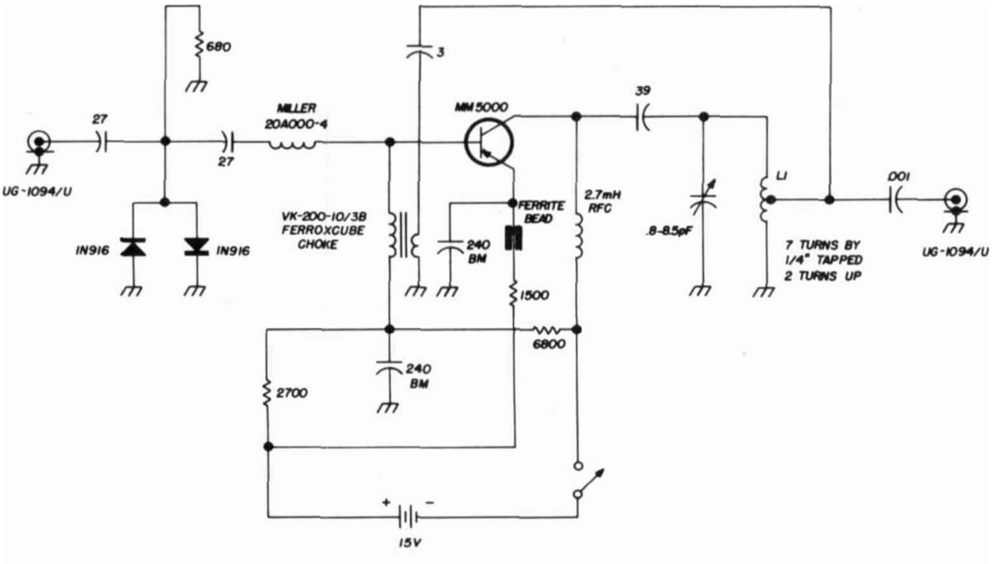


fig. 1. Schematic diagmm of the low-noise two-meter pmamp. Capacitors marked "BM" are button micas.

construction

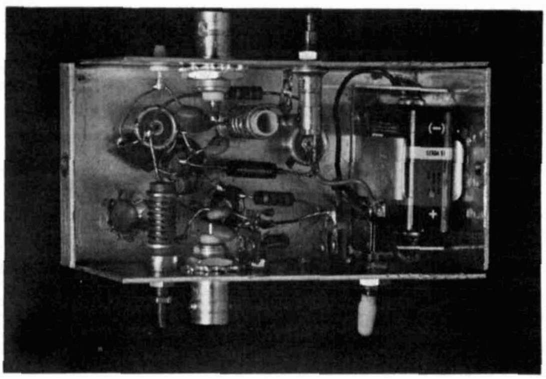
The MM5000 two-meter preamp is constructed in an LMB-421 flange-lock box chassis. It has what might be considered to be a floating ground although it's heavily bypassed to the chassis with button capacitors. A glass piston capacitor is used to tune the output tank. A series NF adjustment coil is provided on the input. The input coupling circuit is designed with an independent "T" section with separate return for the protective diodes and a 680-ohm resistor to dissipate overload power. The input discoidal capacitor values were chosen for series resonance at two meters. Output tank coupling uses tapped-coil impedance selection since it is parallel-resonant. The 0.001 disc capacitor at the output may not be required.

The neutralizing ferrite bead is a Ferroxcube VK 200-10/3B hf-vhf choke.* The extra hole in the bead is used to couple reverse-phase energy back to the base. The neutralization scheme shown costs about 1 S-unit per 4 S-units gain, but stability is excellent! Next to the output connector is a teflon standoff and a 3-pF discoidal coupling capacitor selected to be 5 times the internal feedback capacity of the MM5000.

* Ferroxcube VK200-10/3B chokes are available from the author for $.15 each postpaid.

It's interesting to note that most bipolar transistors have asymptotic noise characteristics. That is, as frequency of operation decreases, noise figure falls off along a constant ordinate of about 2.2 dB for the AF-239 or an estimated 1.2 dB for the MM5000—both maximum values at 150 MHz.

Internal construction of the preamp. The series NF coil is in the lower left next to the input coaxial connector.

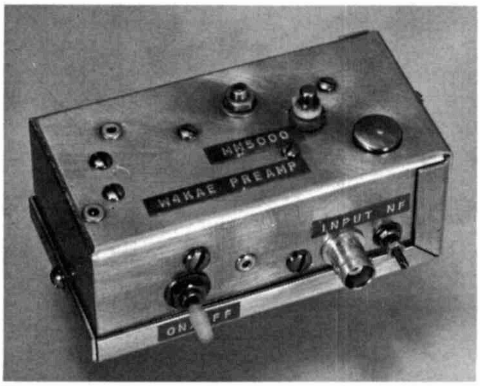


I found that noise figure is best adjusted through the use of series inductance tuned for a dip in the input signal or noise. Of course, the output tank circuit was previously set for maximum gain. I observed that noise reaches a sharp null when using a signal or a broader null when using antenna noise. Further work in peaking the output piston brought the gain up on DX signals. However, the procedure of dipping the signal (or antenna noise) on the input resulted in best noise figure.

This preamp will surprise you! A signal-to-noise comparison with nuvistors shows you can hear a lot more with the MM5000. Going from a manufacturer's spec of 3-dB NF using 6CW4's to about 2 dB with the AF-239 was more noticeable than the change to the MM5000. Noise fall-off with the AF-239 made fairly loud signals louder; the MM5000 made conditions improve. The second change was more subtle since it affected weak signals most.

This is a low-noise bipolar transistor in a ferrite-neutralized common-emitter circuit. Adequate gain with hand-picked values of important components resulted in what I feel is a good design. Outside or chassis grounding is separated from biasing circuitry; one cause of noise entry. Although some of the bipolar disadvantages are present, a quiet location and a well-matched and tuned antenna can provide perfectly satisfactory reception. Burn-out has been a myth with 150 watts output to a flat line.

The complete preamp is housed in a small LMB flange-lock chassis.



The MM5000 is available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Part number 49E26 MM5000; $4.40.

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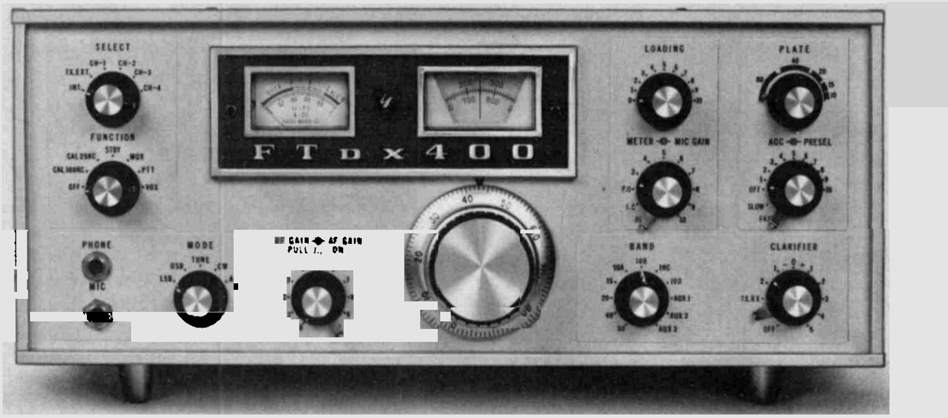
NOT FOR THE NOVICE



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

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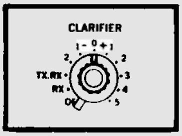
SPECIFICATIONS: Maximum input: 500 W PEP SSB, 440 W CW. 125 W AM. Sensitivity: 0.5 uv, S/N 20 db. Selectivity: 2.3 KHz (6 db down), 3.7 KHz (55 db down). Carrier suppression: more than 40 db down. Sideband suppression: more than 50 db down at 1 KHz. Frequency range: 3.5 to 4, 7 to 7.5. 14 to 14.5. 21 to 21.5, 28 to 30 (megahertz). Frequency stability: Less than 100 Hz drift in any 30 minute period after warm up.



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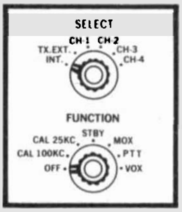

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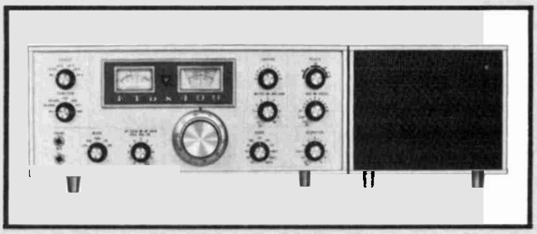

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During recent years, amateur equipment has improved to the point where l-kHz frequency accuracy is expected, and may well be essential for many reasons. Several receivers, exciters and transceivers are available which have a specified dial accuracy of 1 kHz **provided** the calibrator is freshly checked against WWV, and the dial hairline (or other equivalent adjustment) is set properly.

Presently, sub-band problems are arising which make it desirable to take a fresh look at calibrators and more sophisticated means of making frequency measurements. After much consideration of alternate means, it appears that only two of them are worthy of serious consideration: the quick and easy way, and the accurate way. Either of these can be simple enough for the amateur and financially within his reach.

quick and easy

The easy way is to accept some inaccuracy and use existing equipment with little or no change. However, it must have a suitable calibrator that can be checked frequently with WWV and linear dial calibration with l-kHz marks. Let us look at the calibrator first.

Most calibrators use a 100-kHz crystal. This is not the only choice. It's possible to use a 50-kHz crystal, and even a 25-kHz crystal, if you are willing to alter the circuit to make the crystal oscillate. However, in view of considerations which I will discuss later, this approach is not worth a great deal of expense unless it's designed into a new receiver.

The receiver and calibrator most certainly should be well warmed-up so they are reliable within half a kHz. Many people forget this. In Southern Florida, temperature changes are not great, but in California where unheated "shacks" are common, there may be a 50-degree temperature change within the receiver during the first few hours it's turned on in the morning.

It is desirable to minimize the change of calibrator frequency with temperature. Also, you should get a very smooth adjustment which will turn as much as 90 degrees through zero beat. This can be done by using small series and parallel capacitors in the adjustment circuit and choosing zero- and negative-coefficient capacitors to minimize temperature effects. A hair dryer and a towel hood have proved effective for making quick tests of temperature compensation. Unfortunately, most calibrator crystals are on one side of the chassis while the temperature-correcting units are on the other, so the effectiveness of this procedure is somewhat limited.

With an accurate and stable calibrator, the next step is to calibrate dial error. The receiver can be checked against a 10-kHz source to produce dial corrections which apply equally to every band. I have found that the Collins S-line has a sinusoidal error curve, being from a half to a full cycle off from one end of the dial to the other. The error ranged from 300- to 7000-Hz maximum after the end points were set. This can be reduced in some cases by misadjusting the hairline to make the error extend half way on each side of the median frequency.

By using this method during ARRL Frequency-Measuring Tests, it was possible to keep within about 200 Hz of the correct frequency anywhere on the dial. A slight improvement was made by setting the receiver dial exactly on a l-kHz mark (which can be done to within 50 Hz; then the beat note from the unknown signal was measured by matching it with an audio oscillator. The oscillator was calibrated up to 1 kHz from multiples and submultiples of WWV's 440 and 600 Hz tones.

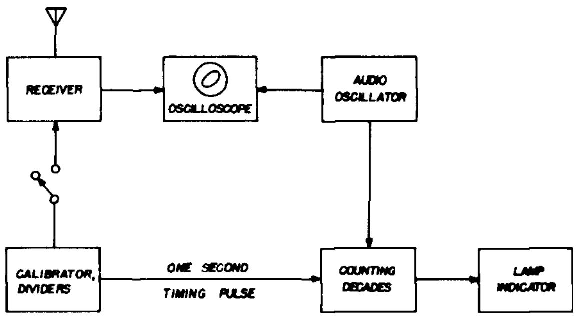
With this approach, measurements were made as close as 17 Hz but others were still 100 Hz off. For most purposes, this accuracy is adequate. It will keep an Official Observer well within the Class-1 category with average errors in the Frequency-Measuring Tests of about ten parts per million. This is provided you avoid 3.5-kHz signals and concentrate on the higher bands where the percentages are in your favor.

the old way

In the past, the accurate way to measure frequencies usually involved the production of a spectrum of accurate 10-kHz harmonics. Then the beat note between one of these and the unknown frequency was measured by using an accurately calibrated 5-kHz audio oscillator.

This presents some problems. Today's receivers cannot always get an audio beat between signals 5 kHz apart. And, the calibration of the audio oscillator from 1 Hz to 5 kHz was difficult until electronic counters became available. When the beat note is low, the receiver may not produce output for measurement purposes. Also, it may be troublesome to determine whether the unknown frequency is a few cycles above or below the calibrator harmonic. Ways were developed to avoid these problems.

fig. 1. Basic circuit used by K6KA for making amateur frequency measurements.



Another approach is to use a separate signal-frequency oscillator (or its harmonic on the higher bands). This can be counted directly at its fundamental when it is in zero-beat with the signal to be measured if a suitable high-frequency electronic counter is on hand. On ssb signals, the receiver can be placed in the a-m position with the external oscillator to restore the carrier. However, there is a slight error in counting which is magnified if harmonics of the oscillator are used on the higher bands.

the right way

There are amateurs who haven't missed Frequency-Measuring Test signals by as much as one hertz. Accuracies approaching this are not difficult today if you go about it right. The first thing is to stabilize the temperature of the calibrator crystal. Home methods might work. A commercial unit with a quiet mercury-solid-state thermostat and a 1-MHz output costs less than $100.00.* The solid-state switching has produced no radio interference here, and the life is long. It is possible to obtain a "proportional" solid-state control on the oven which does not vary at all at a higher price, but the difference in the two is not important. These units have an internal solid-state oscillator circuit.

The crystal and oven I use have a 115-Vac oven, a 12-Vdc oscillator and a built-in divider which comes out at 100 kHz. This is never used for accurate work without a buffer, but it turns out that it hasn't been necessary to use the 100-kHz output. The output is several volts of saw-tooth waveform and is used to drive integrated-circuit JK flip-flop frequency dividers directly.

The accurate adjustment to WWV presented some problems until W6EF suggested that the 12 volts on the crystal oscillator be varied as the "fine" adjustment. A variation of 9.5 to 12.5 volts, with a transistorized regulated power supply, changes the frequency of the 20-MHz harmonic about ten cycles. This is a great convenience for checking that the crystal is oscillating and on frequency. When you get a zero beat only once every ten or twenty seconds, normal fading may obscure it! The potentiometer is then restored approximately to its former position. The frequency stays for weeks within much less than one cycle in twenty million provided that the oven is left on.

When I started, there were two questions. First, would circuitry have to be added to drive the IC flip-flops used in the decade divider? The Fairchild JK flip-flops which I used could be driven directly from the saw-tooth output, with no circuitry between, so this was no problem. Secondly, would any circuitry have to be provided to make the decade outputs audible at 30 MHz? However, the three-volt square-wave output produced an S-6 signal, even from the I-kHz decade which is its 30,000th harmonic! So, no problems here either.

Note that I haven't used the word "multivibrator." They are usually troublesome. The only really stable one I have found is in the Racal RA-17 receiver, similar to the one described in QST for October, 1965. It uses two tubes to produce a 10:1 division and has stayed in adjustment for years until one tube had to be replaced.

It costs about $4.00 a decade to divide frequencies—a little more for the first decade if it must count or divide well up in the megahertz range. One decade divider, or two dual JK flip-flops, or four single ones, will divide by ten, usually with no other parts and no adjustments. So, exit the multivibrator. Even Racal has replaced theirs in current production with a little black box.

With a series of frequency-dividers, it is generally advisable to provide harmonics at 5 kHz for easy use in ssb receivers. Also, it's easy to feed the supply voltage to a decade divider's feedback connection so that it will shift to dividing-by-two twice, producing harmonics of 2½ kHz for those few cases where neither the lower- nor upper-sideband setting provides a satisfactory beat note.

With these dividers in operation from a stable-frequency source, it's an easy step to provide several more decade dividers to count the audio interpolation oscillator frequency accurately, and to indicate it. This refinement is well worth the few extra dollars of cost.

Look at the block diagram in fig. 1. Normally, the unknown is measured approximately by the receiver dial, as a later check on the arithmetic. Then. with the receiver in the a-m position, a beat note is obtained between the calibrator harmonic and the unknown frequency. An audio oscillator is tuned to zero beat with this beat note and its output is counted. This count is added to the lower harmonic or subtracted from the upper harmonic, whichever is being used.

An oscilloscope does a fine job of indi-

* Type SO-1171 from W6EF's Monitor Products Company, 815 Frernont Street, South Pasadena, California 91030.

cating the zero beat. Sometimes it is difficult to determine zero beat by ear with weak signals or in the presence of interference. You must use the same ear to hear both the beat note from the receiver and the output from the audio oscillator. There is no suitable mixer in the human brain to produce a beat note between what is heard separately in each ear.

With the receiver output fed into one scope amplifier and the audio oscillator into the other, a stable elliptical pattern is displayed when both signals are in phase. The ellipse may be at the end of a "cylinder" in the presence of an interfering beat note or noise, or it may be seen during gaps in the interference. Foreign signals may produce problems because of fading, but measurements may be made to within several cycles at worst.

In some cases, it is easier to use the receiver in USB or LSB so that a suitable beat note is heard from the unknown frequency when it's connected to the antenna input of the receiver, and from the calibrator when it is turned on. Two separate measurements are taken with the audio oscillator. One subtraction and one addition must be made at most to obtain the frequency of the unknown; the receiver must remain stable between measurements. This method can be used in a large number of cases where weak signals, noise or interference is present. This method completely eliminates problems which occur between the unknown and the calibrator harmonic near zero beat.

results

What can be expected from this set-up? By measuring up from a calibrator harmonic just below WWV, all measurements have been within 1 Hz of the WWV frequency. This is usually close enough. A ten-second timing gate instead of a one-second gate in the counter will eliminate this. In a recent Frequency Measuring Test, two participants subsequently exchanged their measurements of three transmissions on unknown frequencies. In each case, the results agreed to the hertz. Both of these amateurs used home-brew calibrators and home-brew electronic counters.

ham radio

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designing single-tuned interstage networks

If you do any of your own design work, you've probably had to design interstage networks; here's an approach that takes all of the variables into consideration

Bob Nelson, K6ZGQ, 110 Morning Valley Drive, San Antonio, Texas 78227

Resonant circuits are used extensively in amateur radio equipment — perhaps the most difficult to design are the interstage networks between stages of converters, i-f and rf amplifiers and preselectors. Amateurs design and build this type of equipment every day, but unfortunately some of it doesn't work because of poor interstage network design. This article outlines a fairly simple, yet effective, method of designing one particular type of interstage — the single-tuned tapped-coil. This is probably the most often used circuit, although there are many other single-tuned interstage configurations.

the interstage network

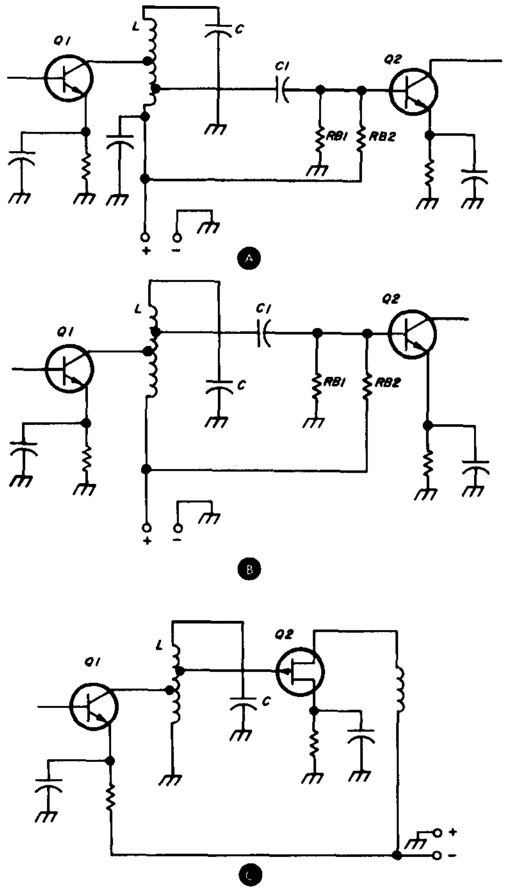
The resonant interstage network has three normal functions: a dc path for the bias current to either the preceding or succeeding stage, or both; impedance matching between stages; and frequency selectivity.

Fig. **1** shows three ways of using a tapped-coil interstage between two transistors. Two NPN transistors are matched by the tapped coil in **fig. 1A;** the output impedance of Q1 is higher than the input impedance of Q2. C_1 provides dc isolation between the collector of Q1 and base of Q2. The coil carries bias curent to the collector of Q1. The circuit of **1B** is similar except that the input impedance of the second stage is higher than the output impedance of the first; thus the base coil tap is higher than the collector tap.

The circuit of **fig. 1C** shows the tapped-coil interstage network at its simplest. Since an

NPN transistor is driving a P-channel field-effect transistor, bias for both transistors can be run through the coil. The three circuits shown here are far from the limit—many other configurations are possible.

fig. 1. Three **examples** of the **tapped-coil interstage network.**



equivalent circuits

The ac equivalent circuits for all the interstage circuits of fig. 1 are identical as shown in **fig.** 2. In this circuit, R1 is the resistive part of the output impedance of Q1 and C1 represents the capacitive part. Similarly, R2 and C2 represent the input impedance of Q2. Note that the parallel combination of Rbl and Rb2 of **fig. 1** should be large in comparison with R2 to avoid loading the input of Q2 and reducing the working Q of the tuned circuit. L represents the coil, Cx its distributed and stray capacitance, and Ro its residual loss resistance. Ca represents any padding or tuning capacitance which may be used.

The equivalent parallel-resonant ac circuit is drawn in **fig.** 3. Here N1 is the ratio of the total turns of L to the number of turns between the bottom of the coil and tap 1; similarly for N2. Since impedance matching is desired, the turns ratios of the two taps must be arranged so that R1 and R2 are matched through the coil. This requires that:

$$N1^2R1 = N2^2R2 \qquad (1)$$

Two Q's should also be defined for future use. Qo is the natural unloaded Q of the coil itself:

$$Qo = Ro/6.28\ fo\ L \qquad (2)$$

Where fo is the resonant frequency in hertz. The other Q is the operating or loaded Q which is called Q_L:

$$Q_L = \frac{\text{(parallel combination of Ro, } N1^2R1 \text{ and } N2^2R2)}{6.28\ fo\ L} \qquad (3)$$

$$Q = 1/(6.28\ fo\ L\ [1/Ro + 1/N1^2R1 + 1/N2^2R2]) \qquad (3a)$$

The heart of the design lies in the power transfer curve shown in **fig. 4.** It is a sad fact in any resonant interstage network that even though impedances are properly matched, not all the power is transferred to the succeeding stage. This is because part of the available power is inevitably consumed by the residual loss resistance of the coil.

The power that is transferred to the next stage depends on the ratio of unloaded Q to operating Q of the coil. This is shown graphi-

the design problem

When starting an interstage network design problem, you'll know six quantities: R1, C1, R2, C2, fo and Pav. The first five quantities were defined before; Pav is the power available from the driving stage. The other eight quantities, N1, N2, L, Qo, Q_L, Cx, Ca and Pt have to be determined. Pt is the power transferred to the driven stage.

cally in **fig. 4**. Mathematically:

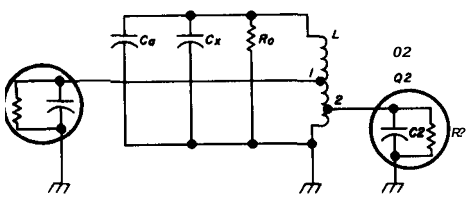
$$P_t/P_{av} = (1 - Q_L/Q_o)^2 \qquad (4)$$

From fig. **4** it can be seen that the loaded Q (Q_L) is always less than the unloaded Q (Q_o). When the unloaded Q is twice the loaded Q, only 25% of the power is transferred; when the unloaded Q is ten times the operating or loaded Q, about 81% is transferred. It should be pretty obvious from this curve that it is important to use coil materials that result in high unloaded Q.

The power transfer curve can be used in several different ways when designing single-tuned interstage networks. The most straightforward design occurs when you have four of the unknowns: P_t, Q_L, C_x and C_a. You can find the power transferred to the driven stage (P_t) from the over-all power gain you're looking for in the amplifier. The loaded Q is specified from the desired selectivity; C_x can be estimated since it's not critical.

You can check C_x out later by finding the natural resonant frequency of the coil with a grid dipper and using the resonance formula. The value of C_a you use in the calculations depends on the circuit; possibly zero if you use slug-tuned coils, or, the median capacitance of the tuning capacitor.

fig. 2. Equivalent circuit of the three circuits shown in fig. 1. The positions of taps 1 and 2 determine N1 and N2, respectively.



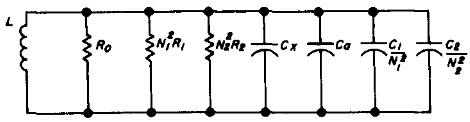
With these four quantities in mind, you can use fig. **4** to find the required unloaded Q of the coil. Inductor manufacturers usually provide this information in their catalogs or you can measure it with a Q meter. Now the inductance of the coil can be calculated from the following formula:

$$= \frac{2Q \quad Q_o - R1\ 6.28\ f_o\ (Q_o - Q_L)\ (C1 + C2\ R2/R1)}{78.88\ Q_L \quad Q_o\ f_o^2\ (C_x + C_a)} \qquad (5)$$

Don't be scared by the arithmetic — it is only simple addttion, subtraction, multiplication and division plus a little basic high-school algebra. When you use this and the other formulas given here, remember that C is always in farads, L in henries, R in ohms and fo in hertz. The values of Q and N are unitless.

If the inductance calculated from eq. **5** results in an unwieldly value, you may want to adjust Q_L, P_t or C_a slightly.

fig. 3. Simplified equivalent circuit.



The next step is to calculate the turns ratio, N1:

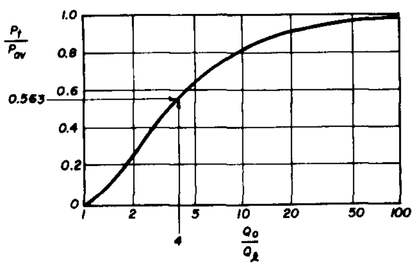
$$N1 = \sqrt{\frac{12.56\ Q_L \quad Q_o\ f_o\ L}{R1\ (Q_o - Q_L)}} \qquad (6)$$

This turns ratio should always be one or greater. If it's not, you'll have to adjust the knowns slightly. Now you can determine the turns ratio N2 from:

$$N2 = N1\sqrt{\frac{R1}{R2}} \qquad (7)$$

This completes the design. You may want to tweak some of the values after construction, but you can be confident of a sound design.

fig. 4. The power transfer curve. This curve illustrates the compromise that must be made between operating Q and power transfer.



method two

This approach is similar to the first one except that the turns ratio N1 is specified instead of Ca. Pt, Q_L and Cx are known and Qo is obtained from **fig. 4** as before. The inductance L is found from:

$$L = \frac{N1^2\ R1\ (Qo - Q_L)}{12.56\ fo\ Q_L\ Qo} \qquad \textbf{(8)}$$

Then Ca is calculated from:

$$Ca = \frac{1}{39.44\ fo^2\ L} - \left\{ Cx + \frac{1}{N1^2}\left[C1 + C2\ (R2/R1)\right]\right\} \qquad \textbf{(9)}$$

The turns ratio N2 is found from **eq. 7.** This design approach will probably find its greatest application where the turns ratio N1 is unity, thereby eliminating one coil tap when R1 is greater than R2.

method three

This design method has a little more of the amateur essence to it. Suppose you have a junk-box coil with known or reasonably accurately estimated Q. Then you can specify Qo along with Ca and an estimated Cx. Now you can determine Q_L and Pt from **fig. 4.** This will require a compromise since you can't have high Pt and high Q at the same time.

After you have found Q_L from **fig. 4,** calculate values for L, N1 and N2 from **eq. 5, 6** and **7,** respectively. Once again, some adjustment of the specified values may be necessary to avoid unwieldy values for the calculated quantities.

Method four is simply a combination of methods two and three. Here you use the basic approach of method three but substitute the turns ratio N1 in place of Ca. The loaded Q and Pt are determined as in method three; L, Ca and N2 are found from the formulas outlined under method two.

practical example

Suppose you want to design an interstage network similar to the one shown in **fig. 1C.** Assume that the input impedance of transistor Q2 is 2000 ohms in parallel with 5 pF. The output impedance of Q1 is 100 ohms in parallel with 2 pF. The power available from Q1 is 2 mW and the operating frequency is 144 MHz. Therefore, R1 = 100 ohms, C1 = 2×10^{-12} F, R2 = 2000 ohms, C2 = 5×10^{-12} F, Pav = 2×10^{-3} W and fo = 144×10^6 Hz.

In looking through the junk box, you find a brass-slug form that has an unloaded Q of 100 on two meters. Also, since R2 is larger than R1, it's desirable to make N2 equal to one. The capacitance Cx is estimated at 2 pF. A quick survey of the known quantities indicates that method four is probably the best approach.

First of all, choose values of loaded Q and Pt from **fig. 4.** You find you need a loaded Q of 25 to obtain the desired selectivity. Since the unloaded Q of the coil is 100, you can see from **fig. 4** that Pt/Pav = 0.563. Therefore, Pt, the input power to Q2, is 0.563 x 2 mW or 1.125 mW.

Next, calculate the turns ratio N1 by rearranging **eq.** 7:

$$N1 = N2\sqrt{\frac{R2}{R1}} = \sqrt{\frac{2000}{100}} = 4.5$$

The inductance is found from **eq. 8** to be 33×10^{-9} henry or 33 nH. Ca is calculated from **eq.** 9 and found to be 30 pF.

Now all you have to do is wind enough turns on the coil form to equal 33 nH. If you have a grid dipper, this is a snap. Since 33 nH will resonate with a 39-pF capacitor at 140 MHz, you can spot solder a 39-pF capacitor across the coil form and look for resonance at 140 MHz when the coil is completed. This completes the design.

Actually, this design method takes more time to tell about than to use. The fastest way to learn it is to use it once or twice in your homebrew work. For those of you who are interested in the mathematical derivation of the graph and formulas, I will send you a copy of the derivation if you enclose a self-addressed, stamped envelope with your request.

ham radio

Cliff Watson, K6WC, pounding the key at the Dewey Mine Station near Grangerville, Idaho in 1906.

early wireless stations

Here is a brief history of early wireless and some of the amateurs who lived it

Ed Marriner, W6BLZ, 528 Colma Street, La Jolla, Californi[redacted]

The old time wireless stations, like the operators who manned them, are gone. The golden age of the sea-going wireless operators who operated the rough notes of spark transmitters or nostalgic musical notes of the arc, made the blood flow in any young man with a wanderlust. The old quenched spark gap with the pickle-jar muffler had a far away sound and lured many an operator off to sea.

In the very beginning I can imagine the young operator, his first time on board ship, with a new transmitter resting in front of him, getting the fragrant smell of lacquers and phenolic compounds enclosed in the tight, stuffy wireless shack. Outside, the smoke, stack gas and carbon grime covered the bulkhead. The canvas lifeboat cover outside the porthole was encrusted with a combination of salt spray and soot.

Perhaps the new operator would familiarize himself with his new treasure before the ship got underway. He might turn on the switch and press the key as they taught him in the Marconi School in San Francisco. Maybe he would take a pencil and draw an arc from the antenna leadin, or watch the meters flick a few times to instill confidence in himself (this was before the time of radio inspectors).

Once back from a long voyage, Sparks would come into port looking for a new berth, spend his money and be out of work. What would he do? Casserly's Bar on Market Street was the most likely place where he could get a free hardboiled egg and a ham sandwich for the price of a five-cent beer.

Portland. Oregon's own retired radio inspector, Joe Hallock, W7YA, on Board the SS Alaska in 1917.

Next in the order of things, the wireless operator had to check in with Malarin, the hiring agent. Malarin would generally tell young Sparks to wait in the static room. Hours would go by. Finally, the young man would stick his head out the door to find Malarin had forgotten him and gone off to the ball game.

Eventually he would be on board another ship with a little more experience. He might have picked up a bag of silicone so he could pick out some good hot crystals for the detector in the ship's receiver. Some of the time might be spent building a receiver from army surplus audio tubes or fixing the spark gap by putting a 30-30 shell case over the gap for a better sounding note. There was also that little trick of dropping the helix to broaden out the signal. Once out of port, he could contact a navy station on 2300 meters using this modification.

On the return trip, young Sparks might have gathered a few bottles of "Old Crow," because prohibition was in effect, and hide them away for his friends. The stowage problem was always solved by putting a few bottles from Canada in the transformer oil or behind a high-voltage fuse panel.

How did wireless start and lead up to glamorous sea-going jobs? There were many tinkerers and experts like Loomis, Tesla, Preece and others fussing with wireless before Marconi. Professor Amos A. Dolbears, of Tufts College, attested to the successful experiments of shipboard wireless by Lt. Bradley A. Fisk prior to August, 1888. Lt. Fisk wound a number of turns of insulated cable around the USS Newark lying at the New York Navy Yard and likewise around a yard tug. He could receive signals a short distance away with a telephone receiver. The system, however, was called induction wireless. and he couldn't claim the invention of wireless.

Nothing really happened in the way of commercial communication until Marconi connected an antenna to his transmitter in 1895. The libraries are full of books about Marconi and his early experiments. It is noteworthy to say in passing that while others dabbled, Marconi had vision and did something about it! On June 2, 1896, he obtained a patent and took his apparatus to England to obtain commercial backing.

Things began to happen fast, and in just a few years, wireless communications were a reality. In 1897, Marconi was operating his own company and was transmitting signals 12 miles away. He reported that Kingstown Yacht Regatta for the British newspapers ashore as a publicity stunt in 1898. The next year, he was able to increase his transmitting distance to 66 miles. The same year he founded the American Marconi Company in Wellfleet, Massachusetts.

The U.S. Navy first tried wireless when

Marconl installed sets on three naval vessels. The first official naval message actually took place on September 30, 1899, when Marconi sent the following message:

Via Wireless Telegraph:

To: Bureau of Equipment, Washington, D.C.
From: USS Connecticut

Under way in Naval parade via NAVESINK station. Mr. Marconi succeeded in opening wireless telegraphic communication with shore at 1234 PM. The experiments were a complete success.

Signed Blish, Lt. USN

George S. Hubbard, the wireless operator who **flashed the first SOS from the Pacific** Mail **Liner Asia** whan **she ran** aground on **Finger** Rock on **the** China **coast.**

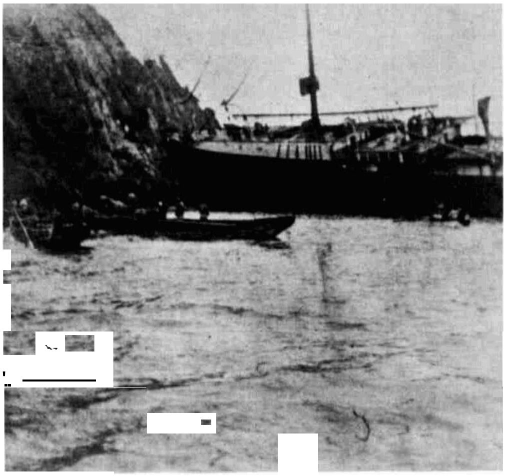
This message was received at the Highland Station on the New Jersey coast. By 1901, all major ships in the U.S. Fleet had been equipped with German-made wireless equipment after three U.S. Naval officers had been sent to Europe the year before to examine various equipment.

On December 12, 1901, Marconi sent a signal across the Atlantic Ocean. His signals were also reaching the Hawaiian Islands, and the army became interested. During 1902, the navy was installing Slaby Arco, Brau-Siemens-Halske equipment designed by Rochefot and Ducretet of France and equipment made by Lodge Muirhead of England. They also purchased DeForest equipment and quenched gaps of American design, including the Lowenstein gap, Simon and others, but it was not until 1909 that the USS Connecticut and USS Virginia had wireless telephone.

Military communications really started in 1903 when the first real message was sent across the Atlantic Ocean and the U.S. Army established communications in Alaska. The first message in Alaska was transmitted on August 7, 1903. At this time the navy had only six wireless stations. Because of foreign control of Marconi equipment, a complete change was made to the Slaby-Arco equipment.

The first International Wireless Conference was held in 1903. At this conference, CQD was added to the operators' signals for distress; however, the Germans continued to use SOE. The New York Navy Yard had a wireless school established with 13 students. DeForest went to England to demonstrate high speed Morse sending. Pop Athern and Harry Brown, **two DeForest** men, set up a station in Shantung, China. Romance had **begun!** A wireless net from Lake Erie to Buffalo, New York was set in operation—a full 180 miles. The operators were known for their Lake-Erie swing, a term which has been handed down and puzzled many over the years.

Imprisoned SS Asia on **the jagged** rocks. **surrounded** by a mob of **maddened** and **blood-thirsty Chinese pirates.**



By **1904,** the navy had eighteen shore stations and thirty-three ships equipped with wireless. Nine ships of the Asiatic Squadron also had wireless. The Saint Louis Fair exhibited a 20-kW transmitter in contact with Chicago, **300** miles away!

The navy completed the West Coast wireless chain of stations in **1905.** The same year SOS became the international distress signal. Lee DeForest sent the Institute of Electrical Engineers his first paper on the audion tube, and the first voice transmission by wireless was made.

In **1906,** the United Wireless Company started spreading out over the U.S.A. Teddy Roosevelt's Great White Fleet was outfitted in **1907** and started on its way around the world. Twenty ships had DeForest equipment on board which was used to contact naval stations up and down the West Coast. The next year, the USS Connecticut, en route to Hawaii and New Zealand, contacted the naval wireless station high atop Point Loma, California, expanding the communications distance to **2900** miles.

In **1910,** the Ship Act required all ships carrying **50** souls, including the crew, to have wireless, although no license was required. On June **30, 1911,** the young United Wireless Telegraph Company hung out the "Out of Business" sign. The officers of the company pleaded guilty to Marconi infringements and were convicted of selling stock under false pretences. The company was purchased by the Marconi Company on June **29, 1912,** the same year the Radio Act required operator and station licenses.

In **1914,** V. G. Ford Greaves compiled a chart showing the the average age of the seagoing wireless operator was **19.** Several operators were listed who were only **15.** They could be found in the shacks of the SS Asuncion, SS Yale and SS Harvard, cruising up and down the West Coast for United Wireless Telegraph Company.

In the Northwest, a lad could always find a berth on the Rose City H2 or stay ashore at 0-2 in Portland, S-2 in Seattle, or take a spin at some of the fish cannery stations in Alaska. It was a great life and a thrill to listen to the rotary spark gap and fog horns when coming up the Northwest Coast. Alas, those days are gone forever—just a dim memory for a few of the old timers who are left.

Syd Fass, W6NZ, on the left, and W7QY, with an unknown operator on board the SS General Lee near KPH about 1912.

ham radio

next month in ham radio magazine:

Transistor Transmitter

Six-meter transverter

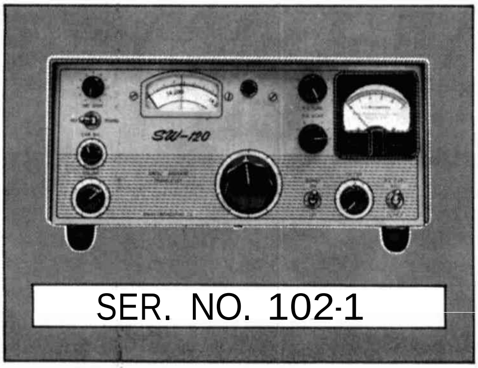
QRP Operation

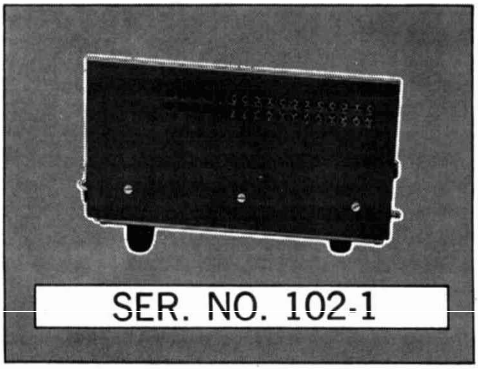
Pi Networks

Stub-Switched Antennas

Plus many more . . .

WANTED





Model **SW120-Swan** Single Bander manufactured in April, 1961 in a garage in Benson, Arizona. Grey, enameled cabinet, clear, anodized panel. Known to frequent the 20 meter band, probably working DX. Height: 5 1/2 inches. Weight: 14 lbs.

REWARD: One new Swan 500C Transceiver with 117-XC power supply.

Swan Electronics began some 7 years ago as a one-man operation with Herb Johnson, then **W7GRA**, building the first 10 single band Swans. At that time the only other SSB Transceiver on the market was the well known Collins KWM-2, selling, of course, for considerably more money. During the intervening years Swan has consistently offered top quality products at the lowest possible cost and backed them up with customer service that is unparalleled in the industry. As a result, Swan is now a team of 160 skilled craftsmen who are justly proud of their position of leadership in the sale of single sideband Transceivers to the Amateur Radio Service.

The first ten transceivers were serial numbered from 101-1 to **110-1**, with the first nine being SW-120's operating on 20 meters, and the tenth, 110-1, being the first SW-140 operating on 40 meters. The company retrieved Serial No. 101-1 about 5 years ago from the original Ohio owner, and have it in our display case. Unfortunately, we have lost the name and call of the original owner of this one. We're wondering now where the other 9 are, and will offer the following rewards for news of them:

(A) A new 500C Transceiver with 117-XC power supply in exchange for the lowest serial number identified by Nov. 1, 1968. This number must be one of the nine from 102-1 to 110-1. We reserve the right to make positive identification before making the exchange.

(B) A new 117-XC power supply will be shipped to each of the other eight early series owners who write in with positive identification by Nov. **1**, 1968. If there is any question concerning serial number verification, Swan will pay shipping costs to the factory and return.

You may be interested to know that not only will the current 117-XC power supply run the early model Swan, but the cabinet on the current 500C Transceiver is interchangeable with the one on the earliest models. You might call this being consistent.

SWAN
ELECTRONICS
Oceanside, California
A Subsidiary of Cubic Corporation

propagation predictions for october

October
is a good month
for DX;
during the rest
of your life
you may not see
MUF's climb
as high as they will
this month and next

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

There are some indications that solar cycle number 20, which began in October 1964, may have reached its peak and that the smoothed sunspot numbers have leveled off and will begin **a** slow descent during 1969. The observed smoothed sunspot number for October 1967 was **94.** The predicted (ITS*) sunspot number for October 1968 is 100. As a result, high-frequency propagation conditions during October 1968 are expected to be only slightly improved over those of October 1967.

The slight improvement may bring quite a few more 50-MHz openings this October, however. Most of these openings will be to the south of east and west and will peak in the midafternoon early in the month—in both the midmorning **and** afternoon later in the month. Chances of direct-path openings between the U.S. and Europe or Japan are slim. Much more likely are side-scatter openings from common directly-illuminated areas much to the south of the direct path. During this solar cycle, the 50-MHz operator aspiring to WAC will wish he had not only a kilowatt and a big beam, but a hilltop location as well.

*** Institute for Telecommunication Sciences of the Environmental Science Services Administration (ESSA).**

Last month I mentioned some observations of MUF's vs ionospheric disturbances that were scaled from ionograms taken at 35.5° N. While it is true that F2-layer MUF's are reduced at this and higher latitudes during ionospheric disturbances, I neglected to point out that F2-layer MUF's to the south are frequently **increased** greatly during disturbed conditions. Moral: when in doubt, turn your antenna south.

Other than F2-layer propagation, you can expect a few auroras, a few good tropospheric openings during Indian summer and a dozen TE (Transequatorial Forward Scatter) openings, at least to the southern states, during the month of October. Sporadic-E will be rare and probably coincident with ionospheric disturbances or meteor showers.

maximum usable frequency

I have been presenting MUF data as a time chart and have noted in past columns that the time chart of MUF(4000)F2 may be considered to first order as a fixed pattern relative to the sun with the earth rotating underneath. This month I am presenting three MUF charts to show the variation of MUF with longitude. **Fig. 1** is a time chart of the median MUF derived from ITS predictions for October 1968 for 45° W longitude; **fig. 2** is for 90" W and **fig. 3** is for 135° W. **Fig. 1** is the more accurate chart for the East Coast looking east, and **fig.** 3 is the more accurate chart for the West Coast looking west.

The greatest differences occur near the geomagnetic equator where there is a trough of MUF—associated, by the way, with one form of TE propagation. Use of these charts has been covered in past columns. First a control point is found for any particular direction 1250 miles away from your station. The MUF at the control-point latitude and local time is the MUF for that direction. For

fig. 2. Time chart of predicted muf for October 1968 for 90° W longitude.

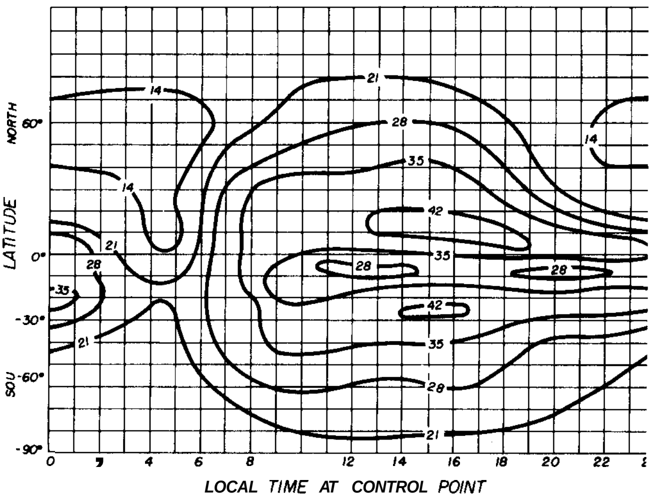
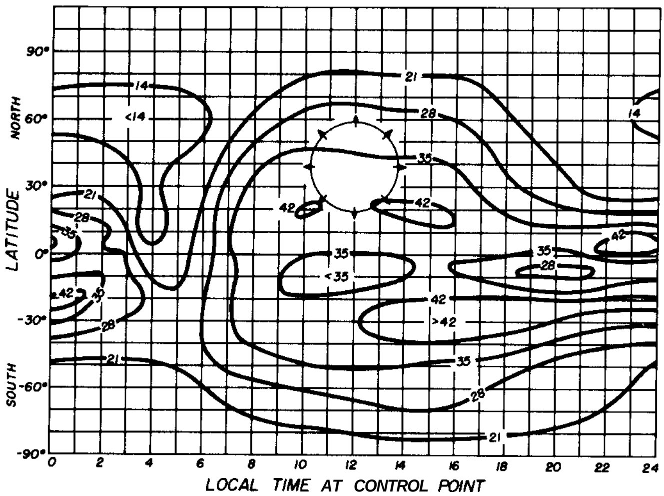
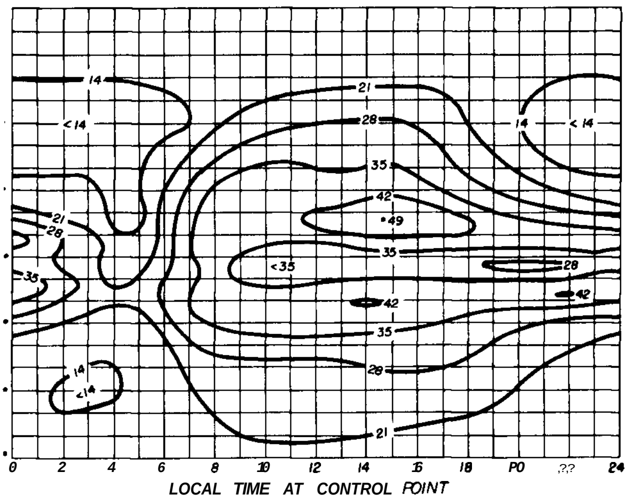


fig. 1. Time chart of predicted median muf for October 1968 for 45° W longitude.



your convenience, a circle of 1250 miles radius centered on 38" N latitude and local noon has been added to **fig. 2.** This circle may be moved (mentally or with an overlay)

fig. 3. Time chart of predicted muf for October 1968 for 135° W longitude.



to your local time and latitude. All your control points then lie on the circle.

maximum range

Time charts of maximum range for 160, 80, 40, and 20 meters as determined by absorption and atmospheric noise levels are shown in **figs.** *4* to **8.** They assume an output power of 100 watts CW or 800 watts ssb on 80 to 20 meters, and an output power of 12 watts CW or 100 watts ssb on 160 meters, with combined receiving and transmitting antenna gains (over an isotropic) of –12 dB for 160 and 80, 0 dB for 40, and 12 dB for 20. An increase of system parameters by a factor of 10 will approximately double the 160-meter nighttime range. It is not known at this time just how significant those peaks in the 80- and 160-meter range near twilight are, since they involve some questionable assumptions regarding the variation of atmospheric noise at these times. There is another factor, not taken into account, which increases the strength of 80- and 160-meter near dawn—the horizontal gradient of ionization found at the dawn line—which makes it a time worth monitoring anyway.

propagation summary for october

160 and 80 meters will awaken for DX purposes. Transcontinental QSO's should be possible on 160, but communications much beyond 3000 miles will require decent antennas, some form of blanker (for Loran—only atmospheric noise was included in the predictions) and careful digging. Eighty should have about 15-dB less noise than 160, no Loran and a 1-kW power limit. Nevertheless, the average station will have trouble working much beyond 4000 miles, and the antenna's as much to blame as anything else. A full-sized dipole up one-half wave-length, or even one-quarter wavelength, is probably at least 6-dB better than what the average station has.

40 meters will have very good propagation across the pole to the sun line. If the European broadcasters don't get you, then the Asian broadcasters will. It would appear that 3 to 5 AM local time might be reasonably quiet and that you could work Antarctica,

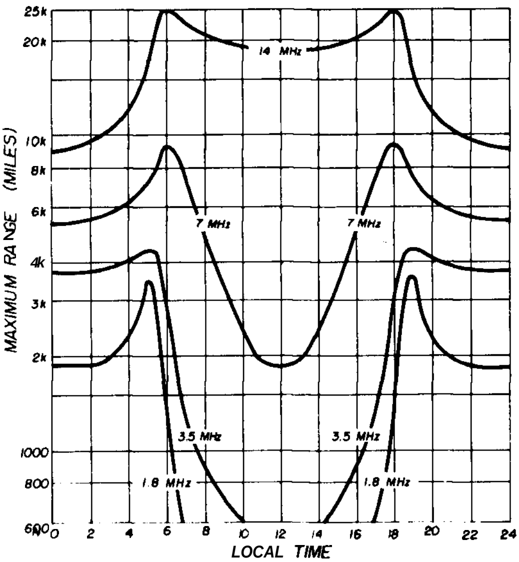


fig. 4. Maximum range vs local time for propagation to the North from midlatitude United States (38° N).

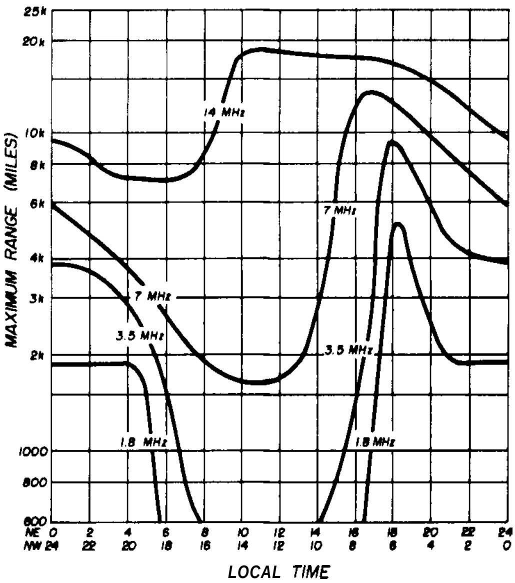


fig. 5. Maximum **range vs** local **time** for propagation to **the Northeast** (top time scale) **and** the Northwest (bottom **time scale).**

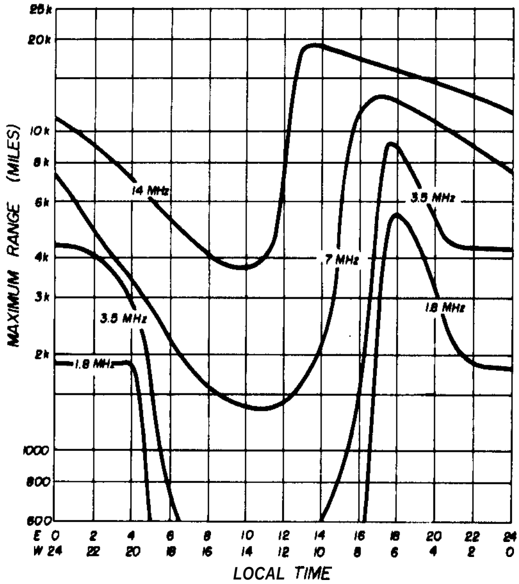


fig. 6. Maximum range vs **local** time for **propagation** to the **East** (top **time scale) and the West (bottom** time **scale).**

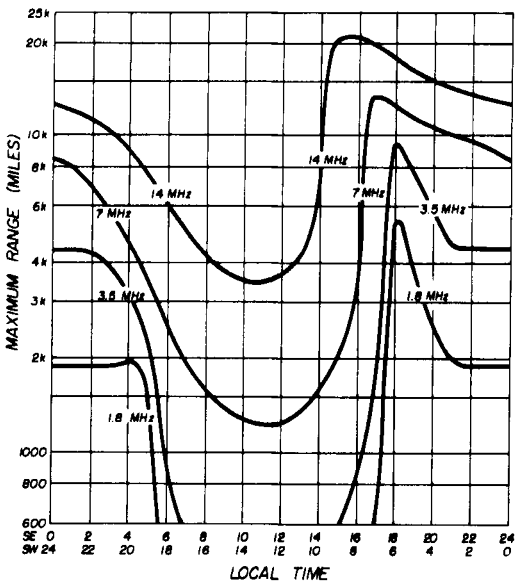


fig. 7. **Maximum range vs** local **time** for **propagation** to **the Southeast** (top **time scale)** and **the Southwest** (bottom time **scale).**

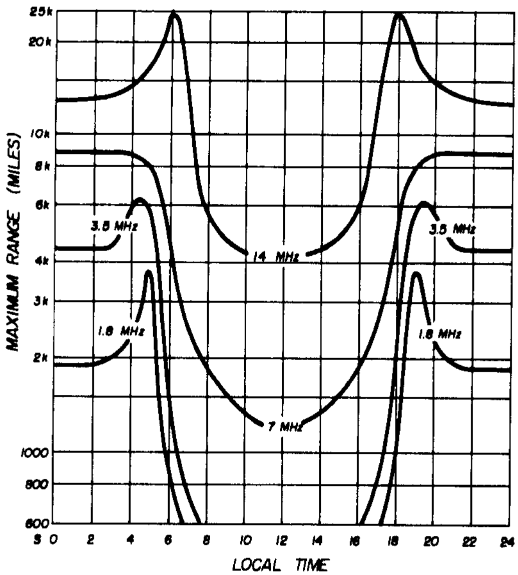


fig. 8. Maximum **range vs** local **time** for propagation to **the** South from 38° North.

how to use these propagation charts

1. To find the maximum usable frequency (F2-4000 kM) In any particular direction from your location, find the latitude of the control point from table 1. The control point will be 1200 miles (2000 kilometers) from your location.

The curves in fig. 1, 2 and 3 show the MUF for the latitude and local time of the control point. Since the control point Is 1200 miles away, local time there is 45 to 90 minutes later then your local time if it is to the east, and 45 to 90 mlnutes earlier if it's to the west. Unless your station is located in the middle of a local time zone, the standard time for your area is close enough for these calculations. Remember that standard time Is the time at 75° W (EST), 90° W (CST), 105° W (MST) and 120° W PST). For accurate time at your location, add four mlnutes per degree longitude west of the longitude that determines the time zone for your area.

Example: Your statlon Is located at 34° N, you want to work east (90° beam heading) or west (270° beam heading). Whet would be the best operating times on 15 meters?

First, find the latitude of the control point from table 1—32° N. From the MUF curve, you can see that *21 MHz will be open for distances* 2500 miles (4000 kilometers) and beyond between MOO and 2030 hours control-point time. The bend will be open to the east between 0430 and 0030 hours, and to the west between 0730 and 2130 hours local time.

2. To find the maximum propagation distance because of absorption, refer to fig. 4, 5, 6, 7 or 8 depending on the direction you want to work. Note that the time scales are reversed for westward propagation in fig. 5, 6 and 7. These curves are based on unity signal-to-noise ratio in a 6-kHz bandwidth with 100 watts output power (100 watts CW or 800 watts ssb), with combined receiver and transmitter antenna gains of —12 dB on 1.8 MHz and 3.5 MHz, zero dB on 7 MHz, end $12 dB on 14 MHz. On ten and fifteen meters, the communications range should not be limited by absorption to less than one transit around the earth. However, anytime you expect minimum range on 14 MHz, round-the-world propagation will be minimal on 21 MHz.

3. To find the MUF for a particular path in the northern hemisphere, locate the other station's control point. Remember that it Is 1200 miles (2000 kilometers) toward you. The MUF curve may then be used to make a crude approximation of his control-point MUF. The MUF Is the lower of the two control-point MUF's—yours and his. These curves are not useful for the southern hemisphere.

The MUF curves should be accurate within a couple MHz between 45° and 135° west longitude. They were prepared from basic propagation predictions published monthly by the Institute for Telecommunications Sciences (ITS), Boulder Colorado end available through the U.S. Government Printing Office. The maximum distance curves were derived from standard formulas at 1000-mile intervals in each of 8 directions from 38° N latitude.

Australia and eastern Asia at this time. The Asians seem to pick up about 6 to 8 AM local time (probably the horizontal gradient effect noted on 80 and 160, except later).

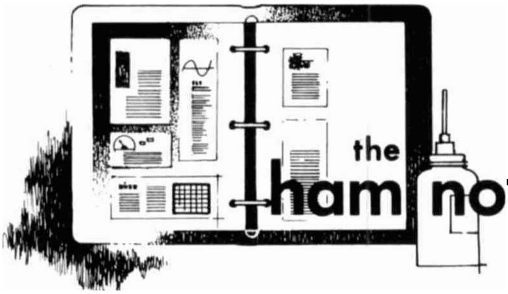
table 1. Control-point latltudes (degrees N).

your latitude (degrees N)	N	NE/NW	E/W	SE/SW	S
24	42	36	23	10	6
26	44	38	25	13	8
28	48	40	27	15	10
30	48	42	28	17	12
32	50	44	30	18	14
34	52	48	32	20	16
36	54	47	14	22	18
38	I 6	48	36	24	20
40	I 8	51	38	26	22
42	60	53	40	28	24
44	82	55	41	30	26
46	84	57	41	32	28
U	66	58	45	34	30

20 meters will begin closing down between 10 PM and 5 AM local time, at least for directions north of east and west, as October wears on. On the nights when the MUF does remain over 14 MHz to the north, however, look for excellent openings to Europe and Asia.

15 and 10 meters will open to the southeast near sunrise and as far north as northeast within an hour later. The bands should be open to most parts of the world that are in sunlight. Ten will close to the NW shortly after sunset, fifteen will stay open to that direction until a couple of hours later. Ten will close to the SSW about 10 PM, and fifteen will follow about an hour later. Early in the month these bands may remain open longer —later in the month they will close earlier.

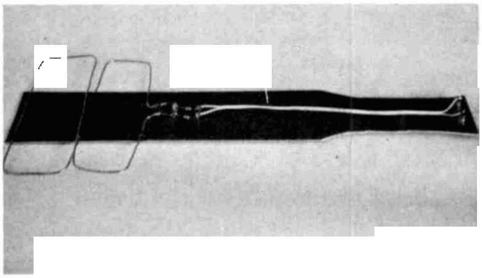
ham radio



the ham notebook

rf current probe

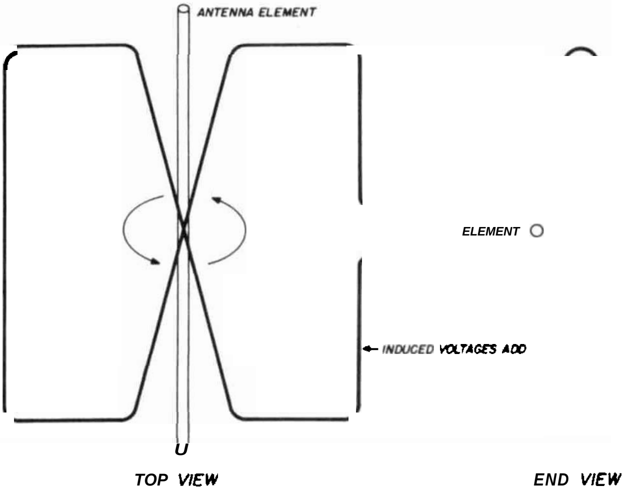
When working with antennas, you often need a way of checking relative current magnitudes in individual elements. Here is a gadget that will do this without breaking the

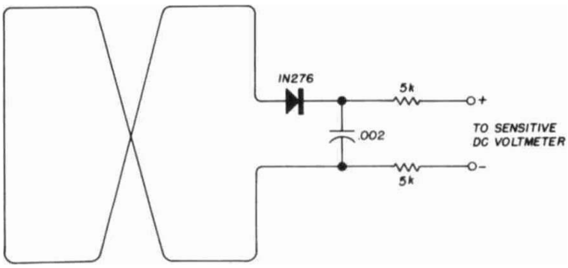


element. It is essentially a loop connected to an rf voltmeter which permits inductive coupling to the antenna element. The voltmeter gives a reading which is proportional to the current in the conductor.

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.

Construction details should be obvious from the photograph. This model was built on a 2-inch wide strip of rnasonite. No dimensions are critical except that the two loops should have approximately the same area.



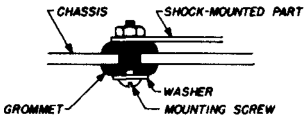


This device proved very useful in checking out the performance of the three-band ground plane described on page 32.

Fred Brown. W6HPH

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

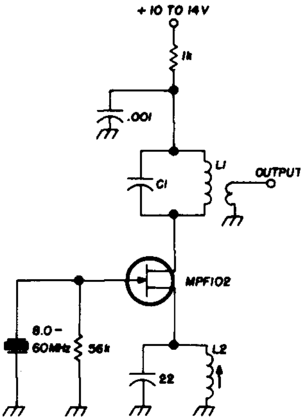


drilled for the grommet; then the relay, etc. is mounted. The grommet acts as a shock-absorbing washer. The larger the grommet, the greater the shock absorbtion. If the inside diameter becomes too large to hold the head of the mounting screw, a flat washer should be added.

D. E. Hausman, VE3BUE

overtone oscillator

Here is a reliable overtone oscillator using a field-effect transistor that may be used with crystals from 8 to 60 MHz. 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

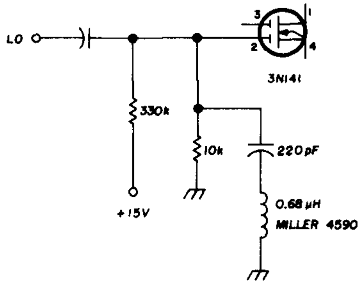


to the third, fifth or seventh overtone of the crystal. Although the schematic shows a link-coupled output, a small capacitor connected to the drain may be used to couple rf out.

George Tillotsen, W5UQS

neutralizing the two-meter mosfet converters

An interesting addition to the MOSFET 2-meter converters described in the August issue is a neutralizing circuit in gate 2 of the mixer. The circuit is series resonant at the 14-MHz intermediate frequency. Since the signals are increased by 1 S-unit with this simple change, it makes the single-rf-stage converter a bit more attractive.



The neutralizing circuit will not work on the 6-meter converter described in the June issue unless the resistor-inductor biasing network in the gate-2 circuit is changed to a resistive divider as was used in the 2-meter units.

Don Nelson, WB2EGZ

space bibliography

Here are some late additions to the space bibliography which appeared in the August, 1968 issue of ham radio. Unfortunately, these additions were received too late to be included with the original bibliography.

W. Browning, G2AOX, "Keeping Track of OSCAR," Radio Communication (formerly RSGB Bulletin), June, 1968, p. 376.

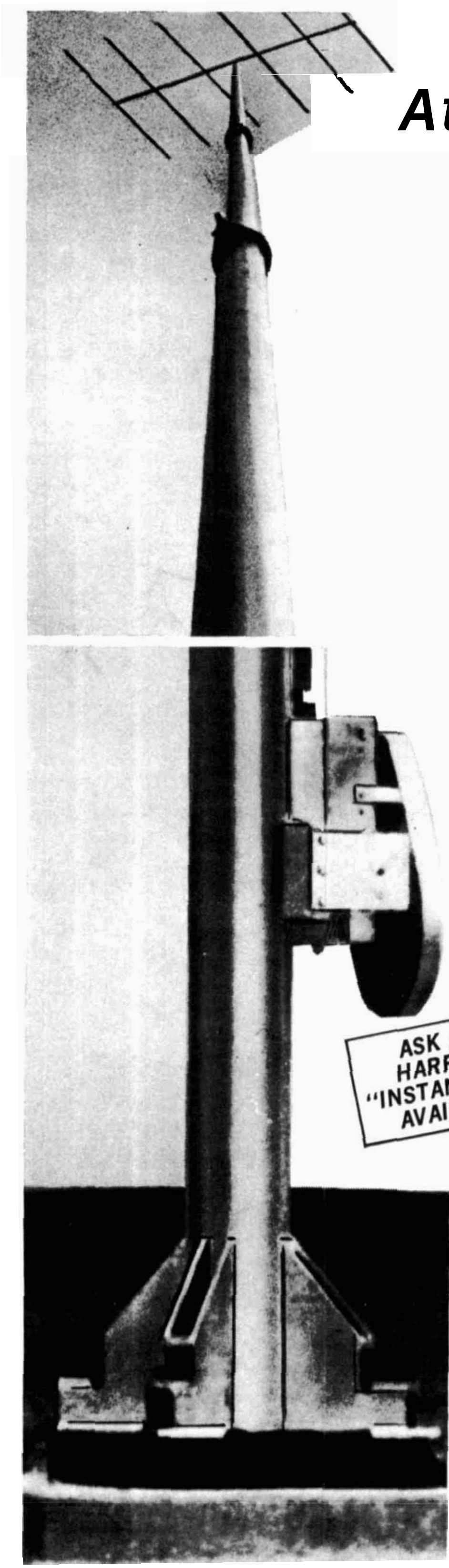
"NASTAR Receives Teleprinter," CQ, May, 1968, p. 46.

"Moonbounce in the U.S.A.," VERON VHF Bulletin (Netherlands), May, 1968, p. 3.

A. Hart, "Space Communication in Australia," Amateur Radio (Australia), September 1962.

"Wide Interest in OSCAR II," Radio, Television and Hobbies (Australia), 1962.

ham radio



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new products

amateur radio circuits book

Most hams spend a certain amount of time building, but most of the gear that is built isn't a Chinese copy of someone else's design. Usually, the builder takes a look at several different construction projects, picks the circuits that appeal to him and puts them together in his own gear.

The new **Amateur Radio Circuits Book** published by the Radio Society of Great Britain is geared to the amateur who is looking for new circuits. This book, compiled by G. R. Jessop, G6JP, contains all types of circuits covering the range from audio to uhf. There is an abundance of transistor circuits, as well as a nice selection of vacuum-tube schematics. There are sections on antenna matching and T-R switches, receivers, transmitters oscillators, power supplies and test equipment. Many of the circuits have never been presented in the United States and represent unique solutions to amateur communications problems.

Each of the various sections covers a wide variety of circuits. In the receiving section, for example, there are preamplifier circuits, converters, local oscillators, i-f filters, Q-multipliers, i-f amplifiers, ssb and a-m detectors, audio agc circuits, noise limiters and noise blankers. The circuits cover the complete spectrum from 160 meters to 432 MHz. Other sections in this handy little book are just as diverse.

The **Amateur Radio Circuit Book** is available in the United States and Canada for *$2.00* postpaid from the Book Division, Comtec, Inc., Box 592, Amherst, New Hampshire 03031. An insert included with each book contains a substitution guide for replacing English vacuum tubes and transistors with types available in the U.S.

cubex tenna switch

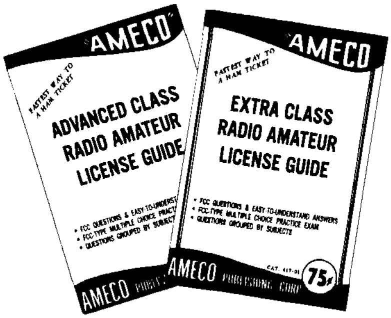
The Cubex TS-4 Tenna Switch is a remote switching system which will let you use up to four separate remotely-located antennas with a single feedline from the operating position. It is ideal for remotely switching bands on multi-band cubical quad antennas or multi-band arrays of yagi antennas. In the Tenna Switch, both sides of the transmission line are switched—this results in more isolation. A single feedline, either coaxial or balanced, is run from the operating position to the Tenna Switch, which may be mounted on the boom, mast or tower. Separate short lengths of feedline are used to connect up to four individual antennas to the switch.

The actual switching function within the Tenna Switch is taken care of by two low-loss ceramic switch decks. The remote control head is designed for 117 Vac, 50- to 60-Hz

current and provides the necessary voltages for operating the Tenna Switch. Lightweight control cable, AWG number 22, is used to connect the remote switching unit to the control unit at the operating position.

The Tenna Switch is capable of handling up to 2 kW PEP or 1 kW on CW. $15.95 from your local dealer, or order from the Cubex Company, Post Office Box 732, Altadena, California 91001.

aerotron radio amateur license guides



Aerotron has announced the availability of two new AMECO books for the amateur radio operator. These books are designed to aid the amateur in upgrading his license in accordance with the latest FCC incentive-licensing program. Book # 16-01 is designed for the general class amateur who is interested in upgrading himself to the advanced class. Book #17-01 is designed for the advanced class licensee who is interested in the extra class license. In addition to book #17-01, a 33-1/3 rpm long-playing record will shortly be available with code-practice text. This will permit the ham to prepare himself completely for the increased code speed requiremnt for the extra-class exam. Both the #16-01 and #17-01 books combine FCC questions with easy to understand answers, practice examinations with FCC-type mutiple-choice answers as well as questions grouped by subject for easy study. The #16-01 is priced at 50c; #17-01 is 75c. Both of these books are available for immediate delivery from numerous radio distributors throughout the country.

NEW FET SIX METER CONVERTER

The Horizon VI incorporates the latest in solid state VHF techniques. Field-effect transistors are used throughout the unit to provide excellent protection against overload and cross modulation.

All power is provided by a built-in 115 volt AC power supply – (no need to buy extras). An additional crystal position has been provided to allow the owner the option of expanded frequency coverage. Local oscillator output is accessible from the front of the unit for those who desire transceive operation.

The low noise figure of the Horizon **VI** assures the operator of excellent performance when **stations** are weak.

See your local dealer for a demonstration of the **Horizon VI;** you'll be impressed by its exceptional performance.

SPECIFICATIONS

Freq. range: **50** – 54 MHz; **I.F. output:** 14 – 18 MHz; **Input** impedance: **50** ohms; Output Impedance: **50** ohms; **Noise** figure: 3dB typical; Gain: 15 **dB** nominal; One 36 MHz crystal installed; **Built-in** power **supply:** 115 volts AC; **Weight:** 18 ounces; Dimensions: L-6¼" x W-3¾" x D-2"; Price: ***$59.95.***

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NEW FET SIX METER CONVERTER

CONVERTER SPECIFICATIONS

Freq. range - 50 - **54** MHz; **I.F.** output. 14 - 18 MHz; Input **impedance - 50** ohms: Output **impedance** - 50 ohms; **Noise figure** - 3dB **typical; Gain - 15** dB **nominal;** One 36 MHz crystal **installed; Built-in power** supply. 115 volts AC; **Weight:** - 18 **ounces; Dimensions** - L **6-1/4"** x W **3-3/4"** x D **2".**

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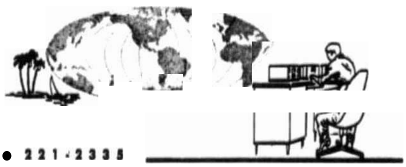

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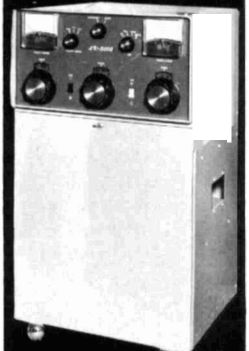

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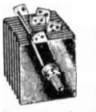

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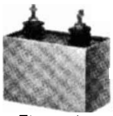

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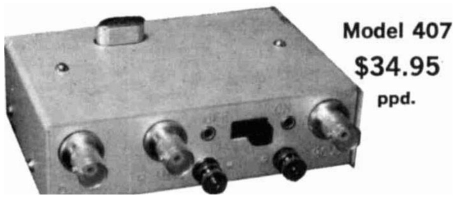

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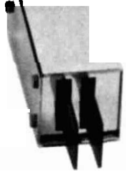

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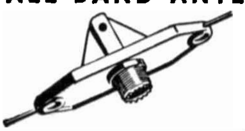

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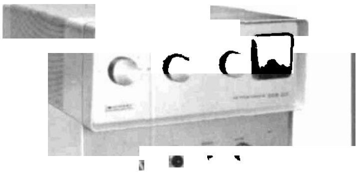

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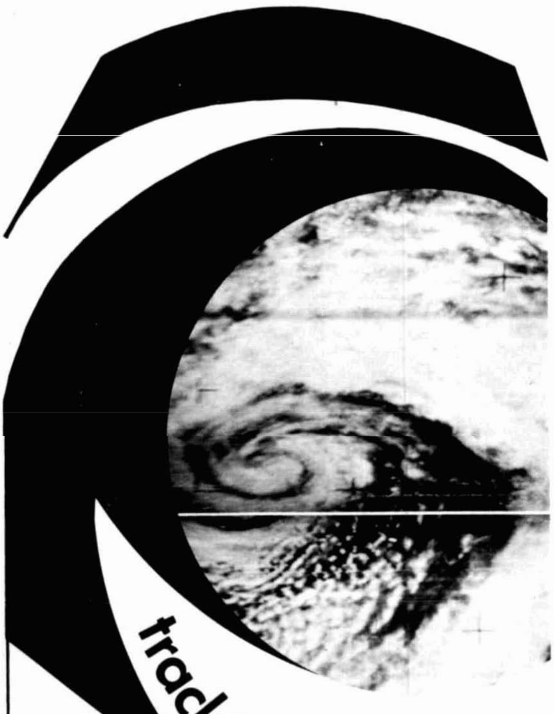

tracking and recording
satellite transmissions

page 6

this month

Ham Radio at its best

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312-B4

516-F2

DL-1

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Adding Collins S-Line components to a Collins system gives you ham radio at its best.

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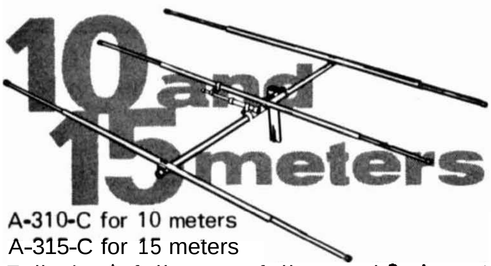

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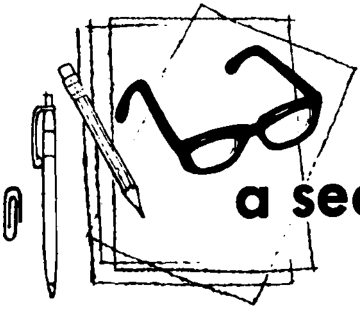


a second look

by jim fisk

If you **asked** the **man** on the street what area of electronics had the most impact during the past decade, he would probably mention color television or computers—they've certainly affected his life the most. To the amateur it might be the widespread use of single sideband or high power on uhf—depending on his interests and point of view. The engineer would no doubt mention lasers and computers, not necessarily in that order. All of these things are important of course, but in the future, the laser will probably have more impact on all our lives than any of the others.

Since the first working laser was put together by Dr. Maiman in 1960, it has captured the attention of scientists and the imagination of the public. To the general public, the laser is a zap gun that will cut through metal like it was a cube of butter. It is much more to the engineer: ultra-fast computers that use optics in place of electronics, three-dimensional TV pictures, and relief to our crowded spectrum by permitting millions of messages to be transmitted on a single light beam. This is just part of its potential; it has already been used in medicine, precision machining and welding, optical gyroscopes, optical recording that is 100 times faster than magnetic tape and optical memories for computers.

This doesn't mean that the laser has taken industry by storm—far from it. But advances are being made every day and it has been predicted that in the future, the lines between electronics, optics and quantum mechanics will be blurred through laser technology. Research is progressing slowly and it will be a good many years before your telephone calls will be transmitted over a laser beam, but advances are being made. Consider the number of materials that have been made to lase—over 2000; more than 1000 of these are semiconductors, with solids, liquids and gases making up the remainder. Solid crystals, carbon dioxide, neon, argon, helium-neon, organic and inorganic liquids and doped glass are just a few that have been made to work.

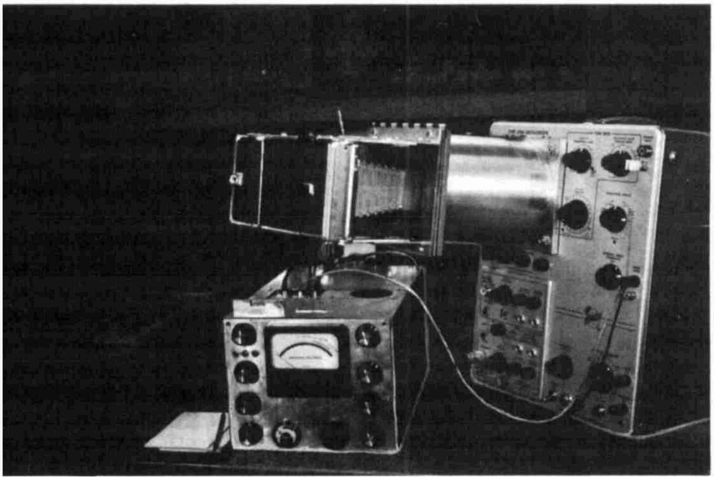
You're probably wondering what all this has to do with you. just this: here's an area for the basement experimenter who has run out of worlds to conquer. So far as I know, the only amateur laser communications were conducted in 1963 by members of the Electrical-Optical Systems Amateur Radio Club. Although the output power of their laser was only 125 μW, modulated at 28.62 MHz with a Viking II, they managed to transmit over a 118-mile line-of-sight path in Southern California. Possibly other amateurs have been working with lasers during the ensuing five years, but I haven't heard about them.

In any event, lasers have come a long way since this early experiment in the San Gabriel mountains. For one thing, costs are down. If you're interested, you can buy a precision helium-neon laser for under $200 from University Laboratories in Berkeley, California. If you're interested in a semiconductor laser, try Allied Radio—they list a 15-watt unit (pulsed) in their 1968 industrial catalog for $95. If you want to build one yourself, consult the Scientific American—they had a complete construction article a couple of years ago.

If you do decide to try the laser, use caution and do a lot of reading first. They can be dangerous if you don't know what you're doing. Relatively small lasers have been used to bore holes in diamonds, so you can imagine what one would do to you if you get in the way of the beam. Protect your eyes particularly; just looking at a laser beam can damage the retina.

If you have a working laser system or are working on one, I'd like to hear about it; I'm sure some of our other readers would too.

Jim Fisk, W1DTY
Editor



tracking and recording satellite transmissions

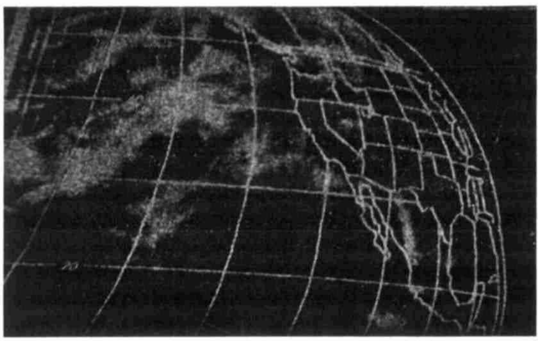
A project that will appeal to the experimenter, builder, vhfer and camera enthusiast

Greg Toben, W6CCN, 1336 Marilyn Dr., Mountain View, California

Automatic picture transmission (APT) is moving into its third year of continuous coverage. Prospects for the increased use of these versatile and extremely useful weather satellites look good.[1]

Two satellites, Essa IV and VI, are now transmitting overlapping APT pictures daily covering nearly the whole earth. An applications technology satellite (ATS-1) in a stationary orbit over the

Photograph of cloud cover over the Northern Hemisphere from the ATS satellite. This picture was gridded and re-broadcast from Mojave, California.



mid Pacific has broadcast pictures of the whole earth and rebroadcast pictures from other satellites, both in the APT mode (table 1).

Experiments will be continued by ATS-3, which is in stationary orbit over South America. A third in the Nimbus series will

in the many experiments is stimulating and rewarding.

This article describes an all-electronic system (fig. 1). It is based on a system used by WSM-TV, but good pictures can also be produced by other means.[2,3,4]

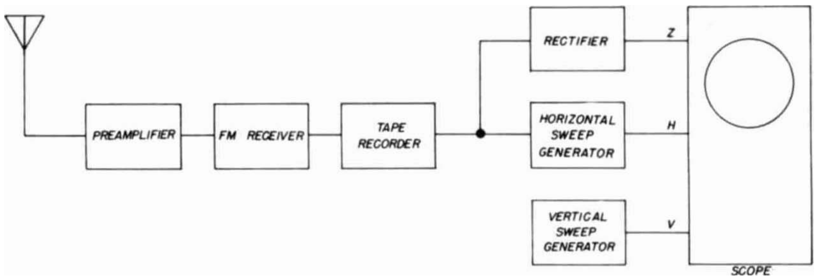


fig. 1. The complete **receiving,** display and **recording system** used by **W6CCN** for automatic **picture trans-**missions. The x-axis mtifier is part of fig. 2; the horizontal **sweep** logic diagram is **shown** in fig. 3.

probably also transmit night-time infrared pictures early next year. A panorama of good pictures is available to users daily from Greenland, the North Pole and Siberia, to Florida, Central America and Hawaii. Coverage includes most U.S. latitudes. NASA encourages you to use this service, and participation

the antenna

Anything from an fm antenna to a stack of multi-element yagis can be used. A four-element yagi from the Handbook probably gives the best results for the least effort, but the picture steadily improves as antenna gain and signal-to-noise ratio go up. An S-meter is adequate

Panoramic view of North **America** as **transmitted** by ESSA **VI June, 1968. It's difficult** to make out the outline of the United States because of cloud **cover,** but the Gulf of **Mexico, Baja, California,** and the West Coast **are** in the clear. Note **the** many **storm** systems that **are active.**

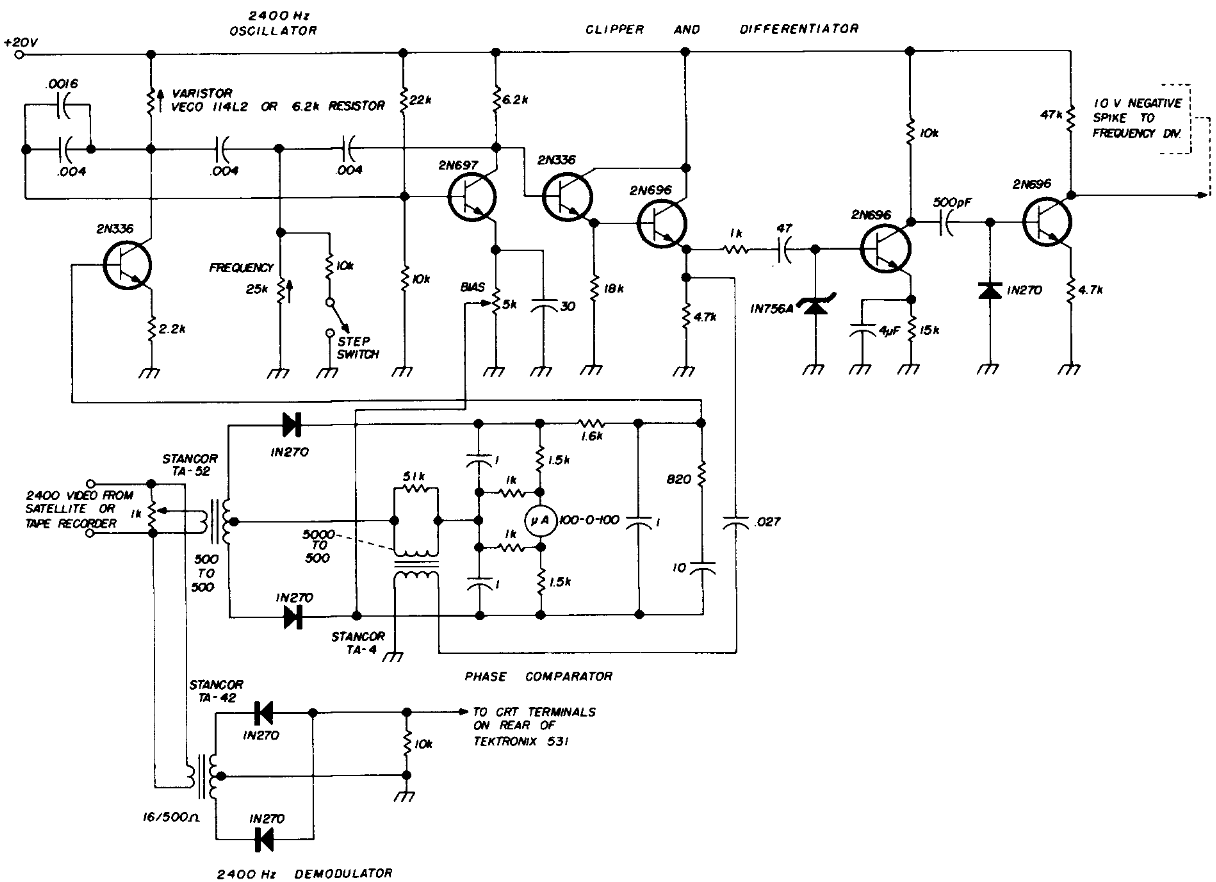


fig. 2. Phase-lock oscillator, clipper/differentiator and phase comparator used in the solid-state sweep generator used to display satellite weather pictures.

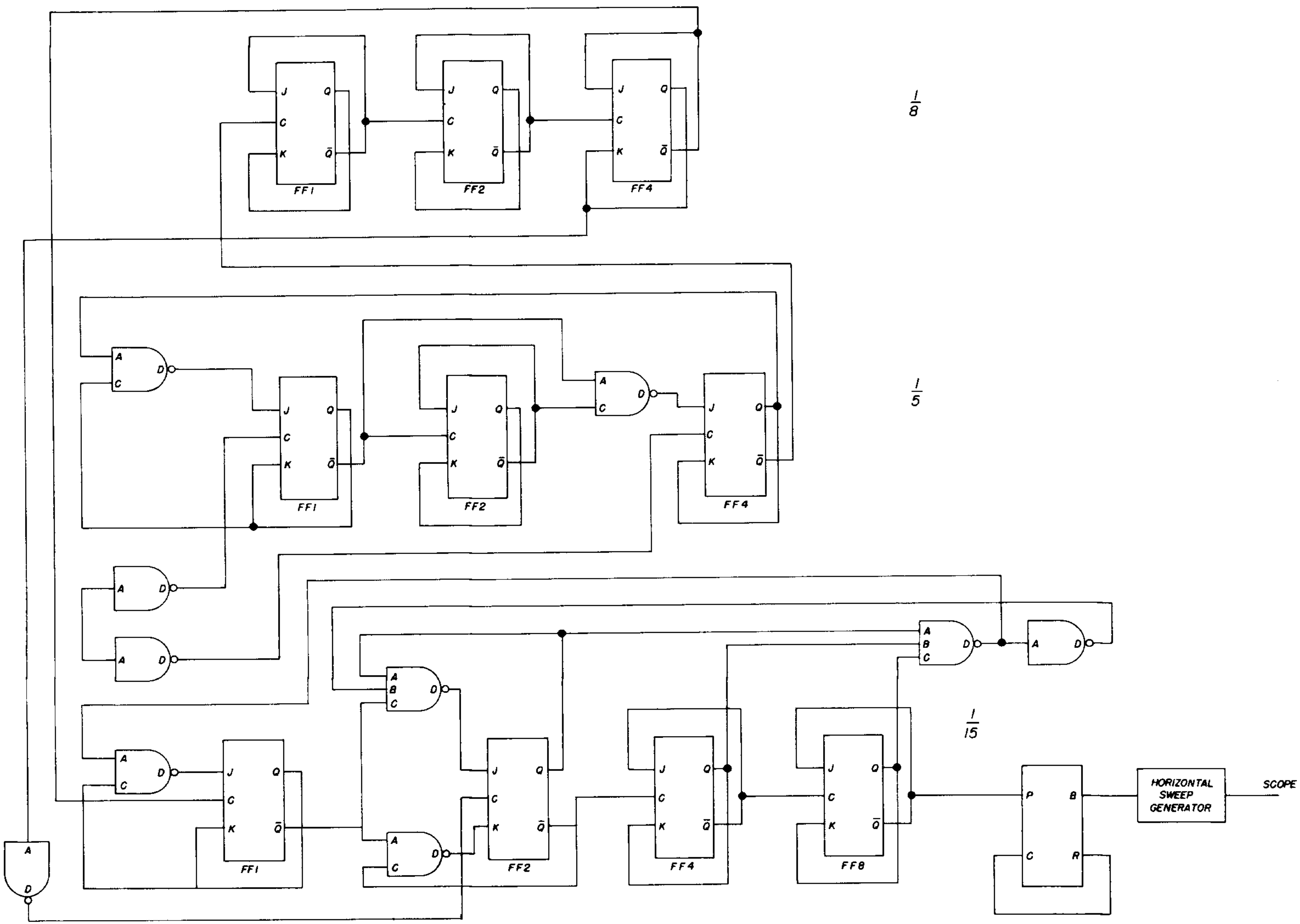


fig. 3. Logic diagram for the horizontal-sweep generator. Although the transistor circuits shown in fig. 4, 5, 6, 7 and 8 were used here, integrated circuits would be ideal for this job.

for monitoring, but a panoramic adapter is even better.

The ear is the most sensitive monitor and can follow the signal right down into the noise. Motor-driven antennas are necessary for remote control, but they are noisy, and need a wide speed range as well as some means for rapid search. Use a manual control if at all possible. So far, all transmissions have been within 1 MHz in the 136-138 MHz satellite band.

The antenna transmission line should have a flat response, and everything should be peaked for maximum gain. Rotation around the horizon (azimuth) is necessary on all but the overhead passes. Nimbus satellites come up in the south and disappear almost due north, while Essa satellites come up from the north and disappear to the south. Tilting up (elevation), is less critical, but rotation around the axis of the antenna (polarization) is important.

So far, the satellite signals have been linearly polarized, and reception on a circularly-polarized antenna has meant some loss of signal. There is a choice, then, of using a helix or crossed yagis and losing some signal, or rotating the antenna on its axis with the attendant mechanical problems.

The plane of polarization rotates as the satellite passes over. It may rotate several complete turns during a pass. The resulting fading is both sharp and deep (20dB). It is hard to adjust crossed yagis for truly circular polarization, and some fading always results.

the receiver

A receiver with 1-μV sensitivity can get a good picture on the overhead passes, but a good preamp is necessary to get clear pictures on the early and late passes. Overloading of the input stage is a problem in metropolitan areas where fm, TV, police and commercial fm stations bathe the antenna in a mish-mash of strong signals night and day. A coaxial cavity between the antenna and the preamp is some help.[5]

fig. 6. Single-shot multivibrator used in the APT sweep generator shown in the logic diagram in fig. 9.

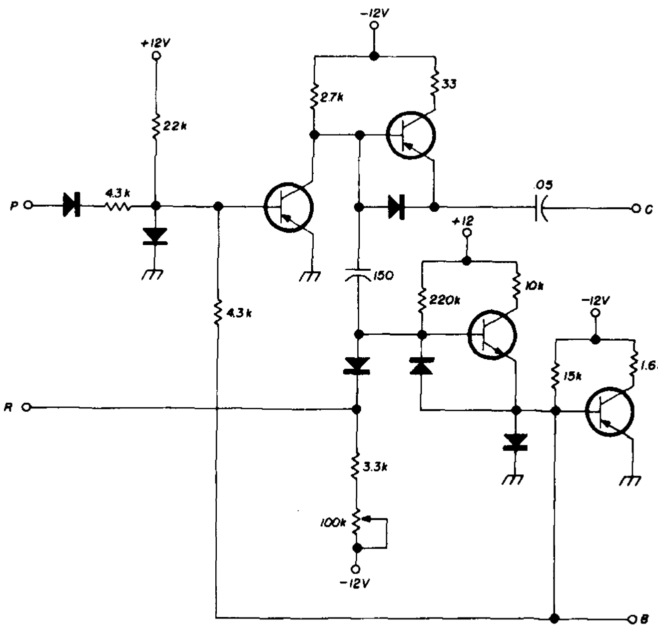


fig. 4. Three-way NAND gates used in the sweep system.

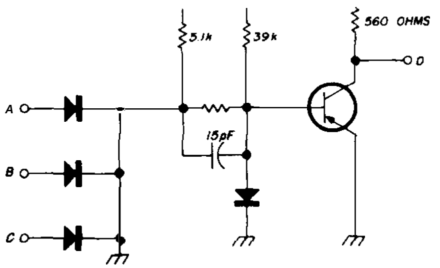
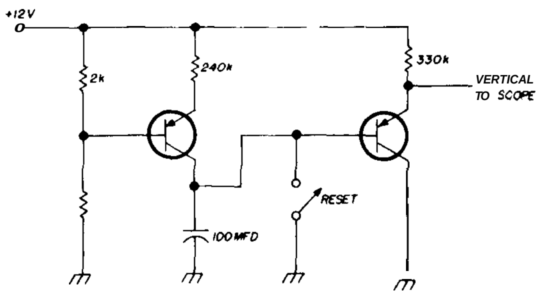


fig. 5. Vertical circuit.



Several receivers are possible, but the very excellent Motorola 148-174 MHz Sensicon "A" receivers are so good and reasonably priced that there isn't much point in discussing alternatives. These are receiver strips from commercial two-way radio equipment such as that used on trucks, taxicabs, etc. They are double conversion, single-frequency superhets and have five cavities in the front end, sharp filters in the i-f strip, good squelch and good over-all stability. The rf stages can be tuned to 137.5 MHz, and a type RM16 crystal (26.4-137.5 MHz) completes the conversion. The output is taken from the discriminator through a volume control to avoid overloading the input stage of the tape recorder. The Perma-kay i-f filter can be removed unless interference is a severe problem.

fig. 7. Horizontal sweap.

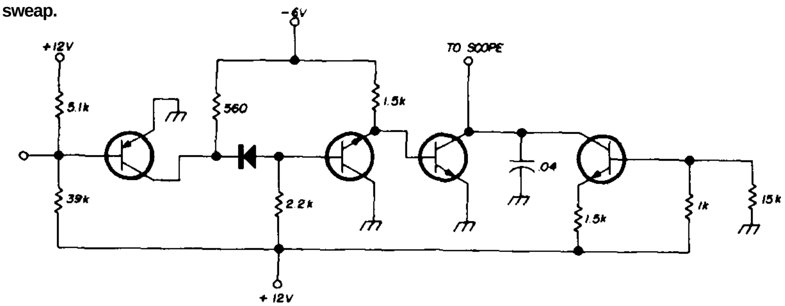


fig. 8. Flip-flop circuit used in the horizontal-sweep generator.

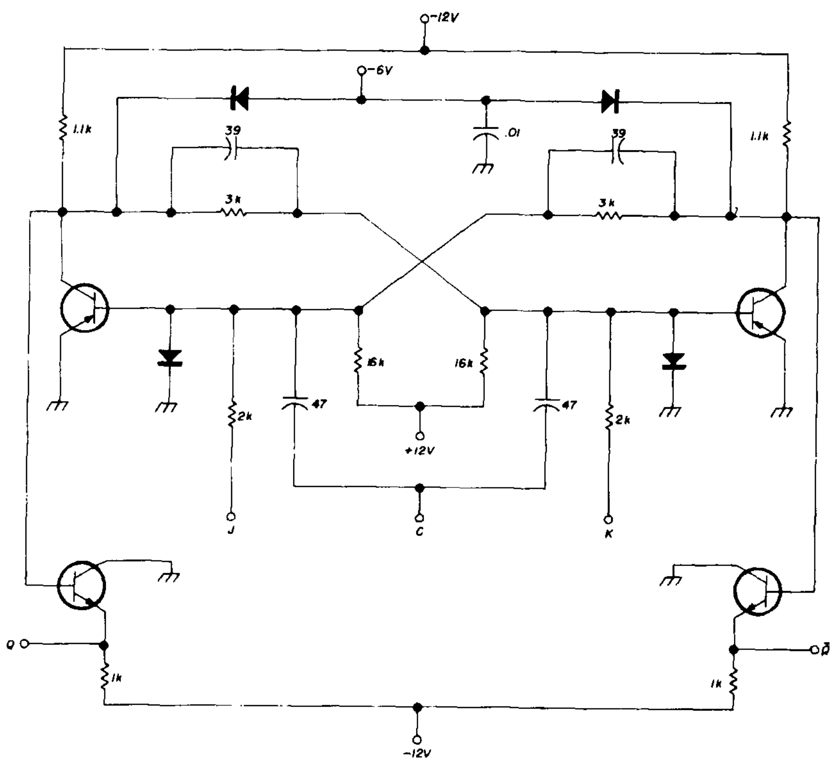


tabla **1.** satellite automatic **picture** transmission systems

ATS	**135.6 MHZ**
Nimbus	**136.950 MHZ**
ESSA	**137.500 MHZ**
modulation	**fm**
deviation	**10** kHz
subcarrier	**2.4** kHz
modulation	a-m
maximum amplitude*	white **level**
minimum amplitude	black level
black-to-white ratio*'	**26** dB
video	dc to **1.6** kHz
lines per picture	800
line rate	four lines per second
frame **time**	200 seconds

Frames repeat every 208 seconds **from** Nimbus and about **every 360** seconds from ESSA.

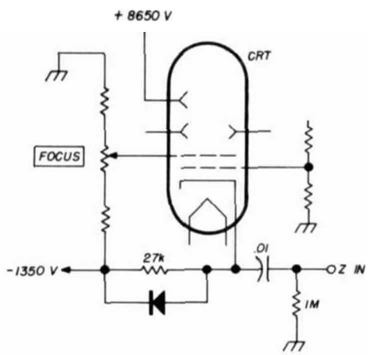
* A whita **level** calibration signal is transmitted between frames.

** Equal voltage steps for **equal gray-scale** stops.

tape recorder and sync system

The tape recorder can be almost anything with 3-kHz response and reasonable speed control. This eliminates only the small battery-powered jobs. A 7-inch reel of 4-track tape will hold pictures gathered in four two-day weekends (1 month). The pictures can be viewed directly, but this takes the operator's attention away from the antenna, and some pictures will be missed while changing film. The high cost of the film dictates that only choice shots be preserved,

fig. 9. Dc-mstom **circuit** for the **Tektronix 531 oscilloscope;** circuits for other oscilloscopes would **be similar.**



and the tape is used for previews and replays.

The output of the tape recorder will vary from the recorded speed if the line voltage or the mechanical friction has changed since the recording was made.

Storm pattern transmitted by the ATS **satellite, gridded** and re-broadcast from **Mojave,** California.

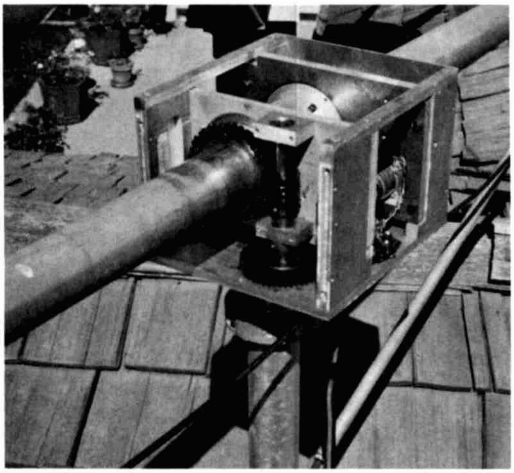
A phase locked voltage controlled oscillator (VCO) keeps the scope synchronized despite these changes. The circuitry for this "black box" is the only homemade electronic equipment necessary.

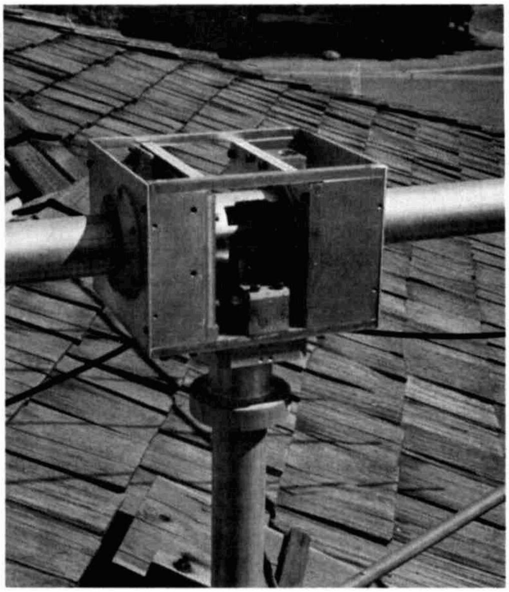
Again, there are many ways to do the job.[1,2] In my system, the shaped 2400-HZ pulses from the **VCO,** which is driven by the 2400-Hz subcarrier from the discriminator, are fed to a count-down chain. The first three triggers count up to 5, then reset and pulse on to the next three. These reset after 15 pulses and drive a single shot. The output of the single shot, through an inverter, shorts out the charge on the horizontal sweep capacitor in a few milliseconds; thus a new sweep is initiated after every 600 pulses. The time constant of the VCO is such that it will supply pulses and maintain sync for a few seconds if the signal fades out.

the scope

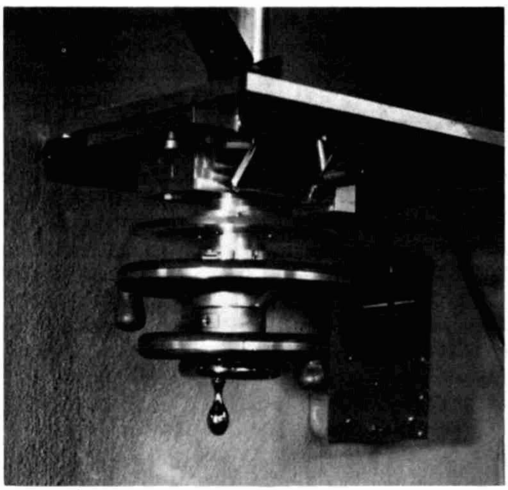
The spot size and focus pretty much determine ultimate picture quality. I use a Tecktronix 531, but I have seen good pictures on a 5-inch Paeco and on a TV tube. The intensity is set so the trace is just barely visible before the Z modulation is applied. At this low intensity, spot size is small and focus is better.

Some of the constructional details of the **azimuth/elevation** mechanism used at the base of the antenna. This mechanism is driven through the roof of the house by the controls shown in the photograph just below.





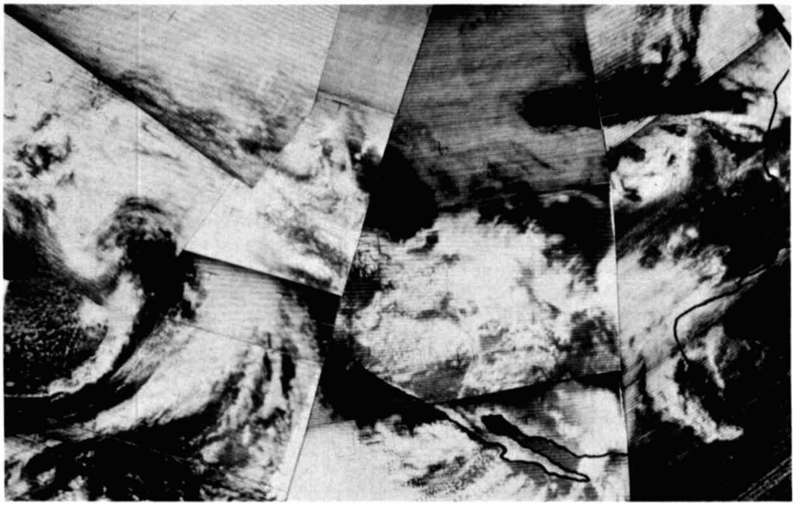
Antenna controls inside the house. From the bottom up is the antenna polarization control, elevation control, azimuth control, azimuth indicator dial, minor readout and clamping mechanism. The selsyn elevation indicator is on the wall to the right.



The vertical sweep is started manually and allowed to run off the bottom of the picture (or top).' Reversing switches on the deflection plates allow the picture to be painted from the top to bottom on n-s (Essa) passes and from bottom to top on s-n (Nimbus) passes. A diode restores picture reference dc to the CR-tube grid. P2 phosphor is about the right persistence, and the green gives a realistic tint to the pictures which will last for about three minutes in complete darkness.

the camera

The camera was made by machining a piece of 6-inch aluminum pipe. The tube is lined with black felt and used as a hood when viewing in daylight. A used Polaroid No. 800 3 x 4-inch camera was used for its developing box and bellows. The lens, iris and remote-operated shutter are from an old Kodak. The sliding mount, as well as horizontal- and vertical-gain controls on the scope, give a complete freedom of picture sizes. Polar-



Panorama of North America transmitted by tha ESSA satellite in April, 1968. if you look closely, you can **see** soma familiar **land-marks** in tha lower-righthrnd **corner—par-ticularly** the Florida peninsula end **Baja,** California.

oid type film 42 at f24 has been used for the past two years, but the newer rapid-processing films look promising for the future.

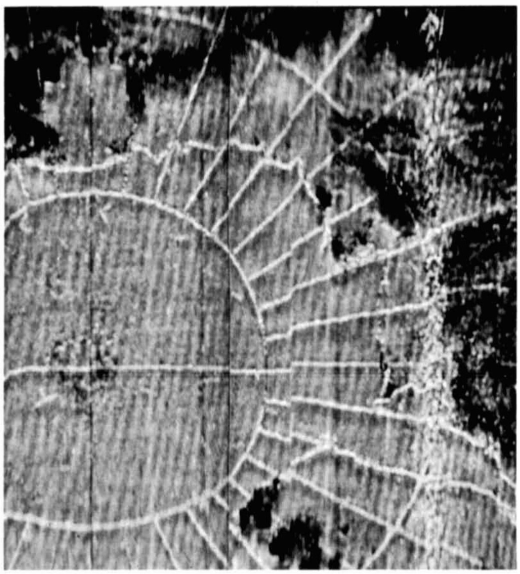
transit times

It takes about 1 hour and 50 minutes for these satellites to make a complete trip around the world. This works out so that Essa VI arrives about an hour earlier each day. If you hear it at 2 o'clock to-day, you could expect it at 1 o'clock to-morrow, or at 8 o'clock six days from now. On a given day the other passes can be expected every two hours less ten minutes, just about time enough to run to the store or cut the lawn.

In an **APT** household everything runs on satellite time. Three passes are available every day, sometimes four, with the mid-dle pass near noon. The longest pass lasts fifteen minutes (three pictures), but the length of each pass depends on where it cuts your receiving range circle. The range depends on the height of the trans-mitting antenna (600 miles) and receiv-ing antenna, and also on hills, build-ings, etc., on the horizon.

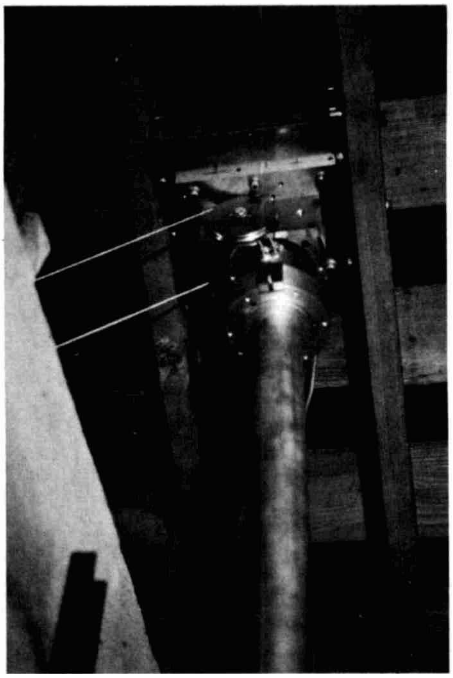
W1AW broadcasts accurate equatorial crossing times, but it works out fairly well to just turn up the volume and squelch and go on about your business until the satellite comes over. It won't be long. The Essa signal is a distinctive 2400 Hz with

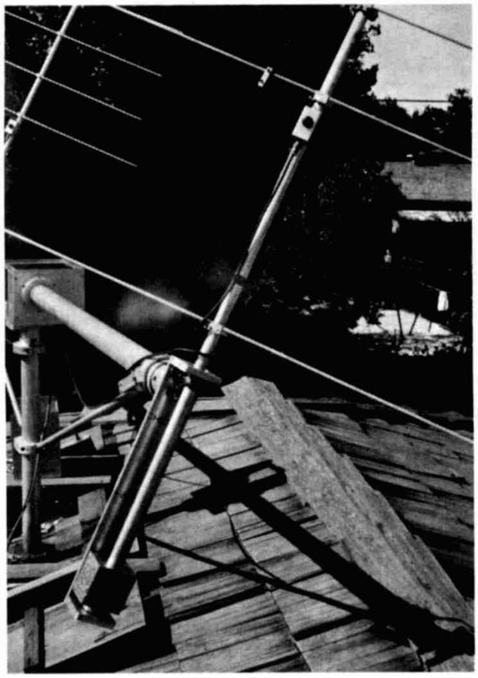
South polar region.



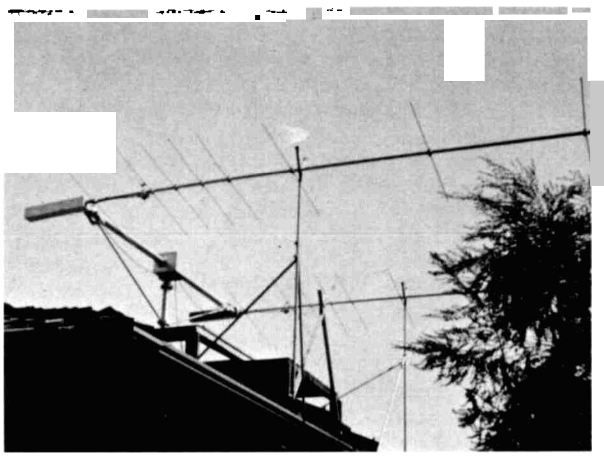
4-Hz modulation for 208 seconds and a steady unmodulated tone for two min-utes, and repeat. Nimbus continuously transmits 4-Hz modulated signals with no space between pictures.

The **ATS** satellites carry a variety of other experiments; however, their power is limited and their time is scheduled to avoid interference with Essa and Nimbus picture-taking operations. They are there-fore on at times that are inconvenient





More details of the antenna system used to track ATS satellites. Above is the mast which goes through the roof to the antenna. Above right is a **closeup** of the antenna mount; two bevel gears are **used** for polarization control; **counterweight** at bottom. To the right is the complete antenna system.



for amateurs. A schedule can be had from NASA when your station is ready.

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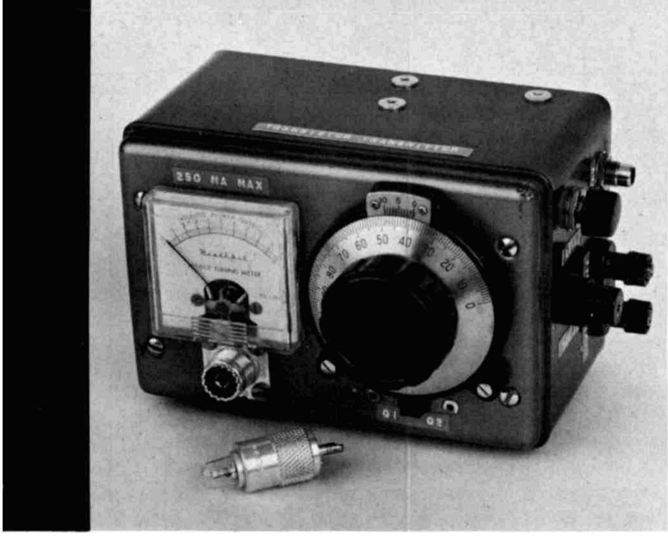
ham radio



broadcast engineer's transistor transmitter

Although the upper frequency limit of this solid-state transmitter is 1600 kHz many of the ideas are directly adaptable to our lower amateur bands

Ralph W. Campbell, W4 33 1 Mariemont ve, Lexin tucky 40505

Notwithstanding the fact that high-power tubes are here to stay, transistors are available at last for a few dollars with a respectable amount of output power. My broadcast engineering friends tell me vacuum tubes are obsolete; this may be true—if you're designing receiving circuits. On the other hand, they may be right on some power applications too.

They wanted me to build a solid-state power oscillator capable of two watts output over a design range of 550 to 1650 kHz. The completed unit had to be absolutely stable with provision for external modulation. The unit was needed to drive a General Radio 916A Radio Frequency Bridge used for testing broadcast antennas after sign off. An audio oscillator is used for modulation.

Several watts of rf power must be developed to override West Coast broadcast signals which are picked up and confuse measurements. The usual method of measuring common-point antenna resistance and reactance is inside the transmitter building where all the phases and loops and feeds converge. However, without a power oscillator, it's impossible to correct mismatches at each tower base which send false indications to the common point.

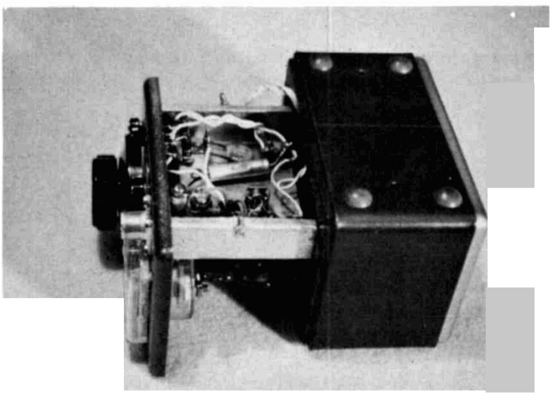
Power is a problem for vacuum-tube equipment because the ac mains to a tower house may fail just when you're all set up. With transistors, a car battery is all you need. At 2 o'clock in the morning, reliable battery power is your best friend. The hunt for a small high-power TO-5 transistor was on.

circuit

This power oscillator uses two 2N2631 NPN overlay transistors. The first stage is a series-tuned Colpitts oscillator operated in class A, and the second transistor is operated in class B. The class-A operating point and loadline is shown in fig. 2. An emitter-follower output stage, which is normally biased "off," is well into the class-B region when an output of approximately 1.7 or 2 watts is obtained, since there is only current gain with this type of circuitry.

1 chose a common-emitter oscillator with an emitter-follower output stage to match the low (and varying) impedances encountered in broadcast engineering work. Oscillator ouput impedance was optimized, mostly by experimentation; frequency compensation over the broad range of the broadcast band was provided (almost accidentally) by using the base bypass value of 0.005 μF with a 200-ohm resistor. This was chosen to equal the measured Colpitt's output-coupled impedance. An "anti-squegg" resistor was added to prevent reflexing in the common-emitter oscillator on its low-frequency range.

A slide switch selects either of two expanded ranges to cover the broadcast band. Two high-Q Carbonyl SF toroidal cores keep frequency stability within tolerable limits, although a trade-off was made in circuit *Q* because of my inability to find high value silver mica capacitors for the low collector and base swamping reactances. An unloaded *Q* of 230 is easily achieved in practice.



As a result of W1DTY's article[1] on transistor oscillator design, I undertook the challenge to design this piece of equipment! No one else has devoted much to this topic. Other articles might have been of help, but the average collector characteristics on the 2N2631 data sheet showed the operating area to avoid second-breakdown. A glance did reveal the ac beta to be equal to 8. I don't know what the dc beta is, but I've found a stability factor of twice this (16) to be handy when designing the bias circuit.

A TO-5 type transistor was preferred after I discovered Wakefield 254SI insulated heat sinks were available for $.25.* If you look inside a 2N2631, you'll see why the chip is so easily heat sunk. All of this leads us back to the large amount of power dissipation required for a class-A stage delivering 1.7 watts. Strictly speaking, 1.75 watts was the goal with good frequency stability and isolation from loading effects. Several transistors, one hopeless 2 AM session and a good car battery—later—I succeeded.

development

The front of the completed transistor transmitter is shown in the photographs. Looking at the surplus meter you can see that an arbitrary maximum limit of 250 mA has been placed on this movement. The meter shunt with this instrument is approximately 1/3 ohm. However, it's best to cut and try even with figured data on hand, because even the best ohmmeters are crude instruments with which to measure shunts. Use a series-dropping resistor and a 1½-volt dry cell along with an instrument of known accuracy in series and adjust the full-scale value for correct current.

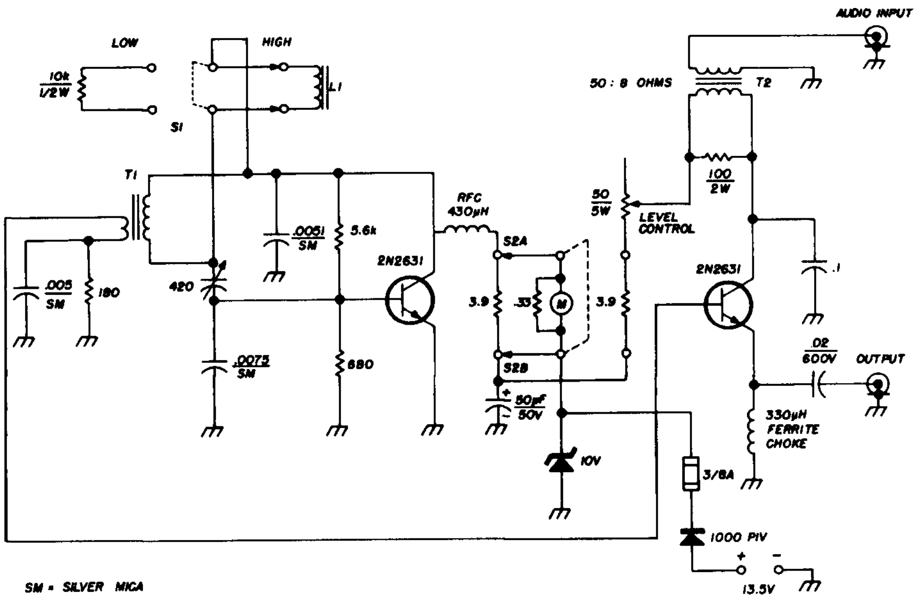
In fact, if the 318-amp fuse holds, use a fast ACX 318-ampere as opposed to the AGC or 3AC type. This is important since the ACC fuse may not be fast enough to save those

*** Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 60E6541. $.25 plus postage; shipping weight, 2 ounces.**

overlays! The ACX fuse is only one inch in length but will fit the usual 1'14-inch fuse-holders.

The 2N2631 overlay transistors have a low collector-to-emitter breakdown voltage. Although the usual rule is to take 1/4 BV_{CBO} as a safe value (assuming 100% modulation), in this case it's safer to use 13.5 volts instead of 20 volts as would normally be chosen. First of all, a safety factor of 75% is necessary to prevent excessive rf swing at some points on the dial (such as when squegging); secondly, voltage at 75% of 1/4 BV_{CBO} works out to be 15 V. I went above this voltage, and my school-of-knocks observations showed frequent failure of the class-B output emitter follower.

Luckily, with modulation I had no problem. But after seeing how hard equipment can be used by professionals in the consulting field, I slapped a +13.5 V limit on the unit. Also, the discussion on modulation should be added to. You can see in the illustrations that I used a small 3-watt modulation



L1 Close wind number-22 win on Permacor 57-1516 ferrite con until 85% of the circumference Is full.

T1 Primary: close wind numbar-24 enamel wire to cover 20% of the circumference of a Permacor 57-1516 ferrite core; secondary is close-wound number 26 to fill up remainder of circumference.

T2 50:8-ohm transformer (Knight 54B1493).

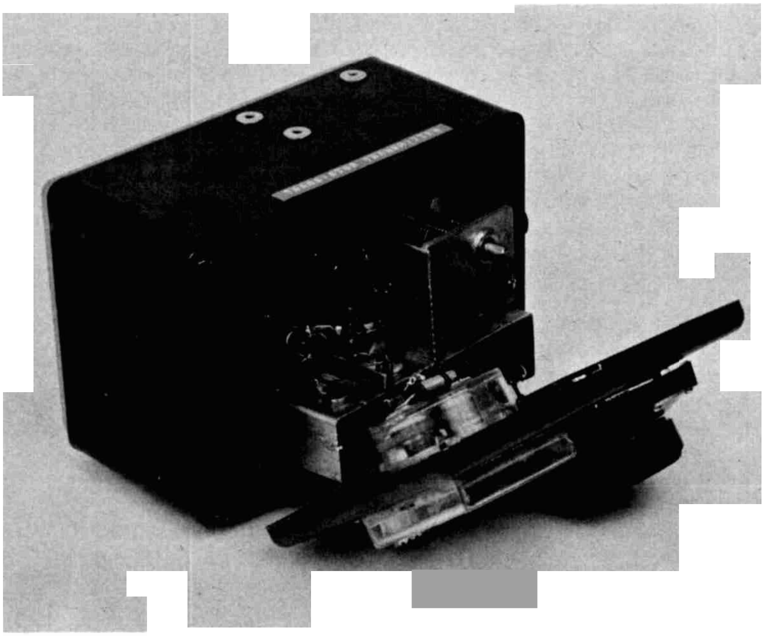
fig. 1. **Schematic diagram** of the **broadcast engineer's transmitter.** Power output is 2 watts. Both transistors are mounted in Wakefield 254S1 heat sinks.

you must allow for peak maximum oscillator base reverse voltage. By this I mean that for the power oscillator stage, a maximum reverse voltage swing of –1.5 volts is all that can be withstood from base to emitter.

With these conditions, even a stable operating point of 10 volts V_{CE} and 55 mA is not held perfectly, because this excursion must exist in any power oscillator. The limiting transformer with a 100-ohm 2-watt composition resistor to swamp hookup transients. This is because transients generated in connecting the unit by moving leads to and from the terminal strip and the battery can cause trouble.

To stabilize the device from such troubles, I installed a 50-µF electrolytic. I got oscillator decoupling from the modulating voltage as well, since these circuits are in tandem if not truly parallel to the supply battery. Behind the upper terminal is a "goof-proof" 1000-PIV diode. This prevents using the wrong polarity from the battery terminal and vaporizing the transistors.

The internal view shows one of the main features: the Permacor 57-1516 Carbonyl SF toroidal core wound with adhesive tape and dipped in INSL-X™ high-voltage coating. The coating and taping were necessary when too many long-nosed pliers pierced a thinner undercoat and a solder-splash adhered to the wire insulation and caused a short.



Also, the engineers had an infield short using the old frequency dial which put collector voltage on the base of the oscillator transistor. Problem: rotor to stator plates touching. This all cleared up with the new Midland dial; rigidity is important.

It is important to mention that the stability provided in this toroidal inductor is higher than would be obtained using a ferrite-cored solenoid. The reason is that copper losses, stray coupling and the Q of the variable 420-pF air capacitor used with it offset capacity considerations when figuring from the charts.[1] The reason this is true is that an extra band or range must be provided, even with 420 pF, to cover the entire BC band. The shunting reactance provided by the second toroid permits coverage from 940 to 1550 kHz.

Complete coverage of the BC band was not obtained at first. It was necessary to remove a blocking capacitor originally installed between the variable rotor connection and the underchassis return to the base of the bipolar oscillator. This spread the lower limit to about 535 kHz and moved the upper limit from 960 kHz down to 940 kHz on this lower band. I found no trimming adjustments were needed, but it was necessary to unwind several turns from the shunting core-wound inductance to achieve the upper limit of 1650 kHz on the upper range. Over-all frequency response was finalized from 535 kHz to 940 kHz (lower position, with anti-squegging resistor in-circuit) and 950 kHz to 1650 kHz (upper position).

The subchassis is an aluminum CB-1625 on top of another CB-1625, cut down to fit the Bud CB-463 utility cabinet. Use of two subchassis was for good heat-sinking purposes mainly, and both Wakefield 254SI insulated sink studs are firmly attached.

The Philmore 1945T variable capacitor shaft is cut very short so that it will fit properly with the Midland dial. Also, space must be allowed for slide switch clearance by cutting a half-circle with an inch-and-118th Greenlee punch on the lower lip of the utility cabinet. Spade bolts hold the chassis as shown, and 118th-inch steel pop-rivets hold the spade bolts. Speaking of rivets, those you see on top are holding in the modulation/intercom transformer. Although pop-rivets are time savers, screws and bolts are really best.

I looked through catalog after catalog for a 50- or 100-ohm low-impedance and low-dc resistance power transistor transformer without luck, except for the very large Stancor TA-11. It was so large it wouldn't fit the cabinet without drilling into the laminations, removing the crimp-clamp housing and twisting heavy-gauge stainless-steel wire around what was left! This was intolerable and could've fallen apart in use if I hadn't found the identical electrical equivalent from Allied.*

There it was, 3 watts, same ac resistance as the TA-11, and I could mount it. I recommend getting two if you can afford it, because you may want to build a transistor a-rn rig sometime after getting your feet wet using 2N2631's. In all fairness to the TA-11, I must admit the extra size may be necessary for low-frequency hi-fi performance in other applications.

The high-range toroid is, as I said above, wound with heavier wire switched in shunt with the low-range one. The output coupling to the emitter-follower is the same as for the first case. Some possibility for squegging occurred, so a 10k half-watt resistor was used to swamp the low-range coil.

Silicone rubber is the chassis mount for the higher frequency toroid. Both of the cores use this as an encapsulant as well as a mount. With the first coil, two 1'12-inch fiberglass circles were cut from 1116-inch sheet stock. A 9164th-inch hole was drilled in the center of the inner disc, and 6-32 binder-head machine screws and nuts were tightened down on the pretinned Beldenamel wire from the inside.

design

Probably most difficult and most important is finding the correct operating point. Since there is little information in simple form available to the amateur designer, you must rely on load lines and select an operating point so the game fits the rules (to explain operation, if not predict it). Fig. 2 shows the load line I used. With the manufacturer's data sheet in hand, I took the V_Q, or voltage coordinate point, from the maximum supply voltage of nominally 13.5 V, allowed for a 3.5-volt drop in the combined circuit and saturation resistances (R_t), and established the quiescent current at the measured value of approximately 55 mA.

fig. 2. Load line used in designing the transmitter.

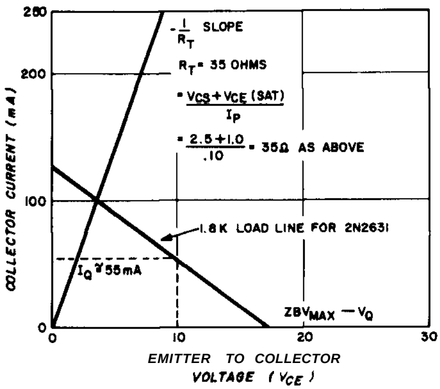


Fig. 2 shows the collector voltage swing from V_{CS}, the collector voltage at saturation (not neglecting circuit losses) to a peak value of twice the supply voltage (V_{CCmax}) or 27 volts minus the V_Q value, ten volts. Looking at the graph, the 17-volt swing should be the first thing you see. Any convenient point on the load line yields a value of about 1800 ohms for the load resistance.

Load lines can be drawn in any part of the operating area, but I suspect that rf base current is so difficult to measure that you should use the cut-and-try approach. Since the overlay transistor is priced below $5.00, this may be the most feasible anyway.

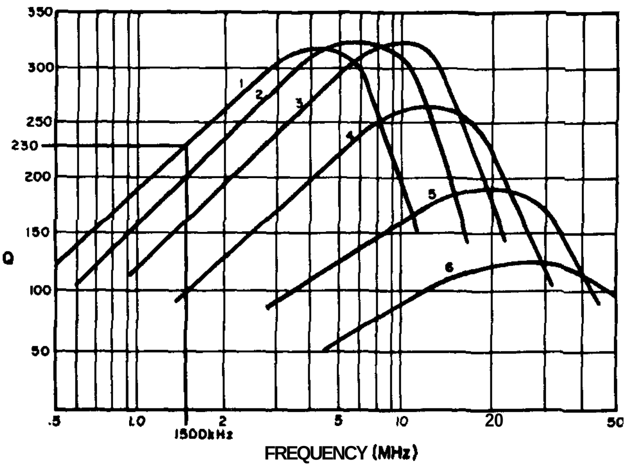
* Catalog number 54B1493. $2.03 plus postage; shipping weight 8 ounces.

The schematic diagram of the transistor transmitter is shown in **fig. 1**. The biasing resistance was chosen to equal approximately three times the load line value or 5.6k. Cut-and-try methods resulted in about 820 ohms form R_b, and R_a equal to 3.9k. A more reasonable criteria might result in higher dc stability by finding the stability factor. However, I felt that since this was a power oscillator, selection of minimal biasing resistance for R_b was more important.

At three times the load value, only 25% of the oscillator power is consumed in the biasing resistance. Another consideration is the Q with the Permacor 57-1516 core. The specifications for the Permacor core are shown in **fig.** 3. The inductor I used is 14.2 μH; from fig. 3, the unloaded *Q* is 230 at 1500 kHz.

fig. 3. Specifications for the ferrite Permacor cores used in the transmitter.

1. 48 TURNS .020 WIRE = 14.2 μH
2. 33 TURNS .020 WIRE = 6.7 μH
3. 24 TURNS .020 WIRE = 3.9 μH
4. 16 TURNS .020 WIRE = 1.6 μH
5. 9 TURNS .020 WIRE = 0.5 μH
6. 4 TURNS .020 WIRE = 0.18 μH



The output 2N2631 is operated in class B. Input pulses turn the emitter follower on and off with a residual bias voltage stored in C_3, which is also a bypass; the time constant is chosen to compensate over the lower range of oscillator frequencies from 535 to 940 kHz. It turned out that the same values worked well on the higher range from 950 to 1650 kHz.

Before going on, a word about the biasing of the oscillator transistor is necessary. At first R_a and R_b were chosen experimentally to turn-on the transistor. The values were 3.9k and 820 ohms. This wasn't very scientific. So, using these values to explain the results, we computed the stability factor. Results: stability factor of almost exactly 16! Twice the ac beta. The network provided adequate dc stability (something pretty hard to find in germanium bipolars) and no overheating.

conclusion

In this article I have shown how an overlay transistor can be used as a power oscillator for consulting broadcast engineering. I am directing this effort to the ambitious engineer who is also a competent amateur and hope that we'll be getting people interested in transistor uses by showing how a commercial piece of equipment can be made to work reliably and furnish power. Not much has been said in periodicals or outlines, so I assume the empirical method to be as effective as any. The biggest hitch, I think, in rf calculations is the difficulty of measuring rf base currents where needed and deducing class of operation.

Class of operation is sometimes difficult to determine even with tubes, since curves don't hold in every case. With a beam pentode I am currently using, different screen voltages and drive can push an AB_1 linear into AB_2 or B. In my own uses with tubes I have found that by simply listening to the linear output as detected audio, you can learn quite a bit. And, with constant current curves (or presumably constant collector curves), output power and tank-circuit efficiency can be calculated.

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ham radio

single-sideband detectors

How the single-sideband signal is demodulated in your receiver

Forest H. Belt, 2610 Whittier Avenue, Louisville, Kentucky 40220

The air is full of single-sideband signals these days. Up and down the ham phone bands, a-m holdouts can hear the donald-duck chatter of their modern-minded cohorts QSOing away—squeezing every last decibel of usefulness from every watt—on sideband.

It doesn't take much to turn that chatter into plain talk. Just a special detector will do it. At least those guys with a-m sets could listen.

The sideband operator already has that special demodulator, built right into his ssb receiver. It goes by many names, but the one used most is **single-sideband detector.** Other names come from the method of operation. Product detector, heterodyne detector, carrier-insertion detector, bfo detector—are among the terms that describe typical ssb demodulators.

The basics of a sideband detector are simple. The signal your sideband receiver picks up is nothing but one sideband of some operating frequency. To recover the voice modulation which created that sideband signal, you need a carrier for the sideband to heterodyne with. (That's how an a-m detector works; the sidebands heterodyne with their carrier in a nonlinear detector — usually a diode.)

A single-sideband signal has no carrier of its own; that was removed at the transmitter. So a carrier has to be added at the receiver. Then the carrier and sideband can be fed together through an ordinary diode detector, and the voice signal recovered.

The i-f amplifier is the best place to mix the carrier and sideband signals. The frequency there is always the same, no matter what band is tuned in up front. A single-sideband detector mixes the i-f sideband signal with a signal at the frequency the i-f carrier would be if there was one. The steady signal is then called the carrier, since its purpose is to supply a signal against which the sideband can beat for demodulation — the purpose of a carrier.

The steady signal in receivers is most often supplied by the bfo that is used for code reception. Adjusting the **bfo pitch** control lets you control the timbre of the demodulated voice. In transceivers, the signal more often comes from the carrier oscillator; it's common practice to generate the initial carrier (before balanced modulation, sideband filtering, and up-translation) at the same frequency as the i-f. Whatever its source, the fixed frequency is fed to the demodulator system along with the sideband.

You can see one oversimple system in fig. 1A. (Don't bother copying it though; it's inefficient.) For distortion-free detection in any mixing-type ssb demodulator, the carrier signal must be much stronger than the sideband signal. One way is to attenuate the sideband signal; that's why the variable attenuator is included.

The arrangement in fig. 1B is a little more effective. The improvement comes from isolation provided by an amplifier between the carrier source and the mixer. The amplifier also gives that needed boost to the carrier signal.

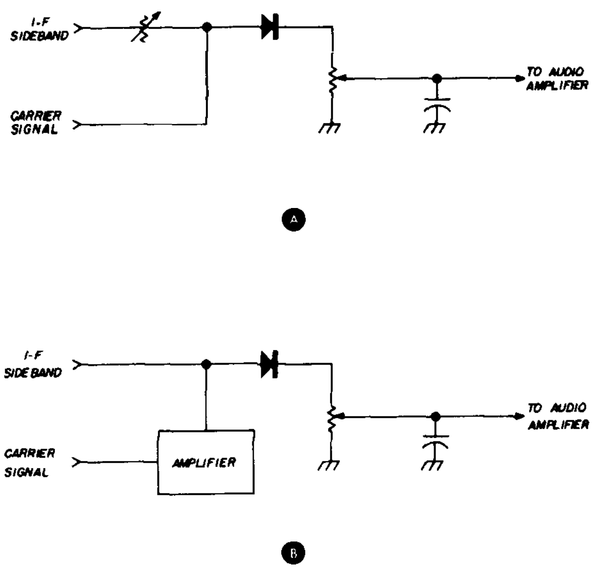


fig. 1. The **simplest principles of ssb demodulation.** There is no isolation between signals in A; **isolation** plus **carrier amplification are provided** in B.

toward a better way

Extra care must be taken with single-sideband detectors. Distortion is always a possibility, unless each signal is handled so that the only nonlinearity is in the detection circuit itself. Applying the signals to a single detector diode is not the most desirable way to get this particular job done.

Better efficiency can be had from the improved version in fig. 2, using two diodes. The carrier signal is applied to them in a parallel mode (its coupling capacitor is connected between their cathodes). The i-f sideband signal, on the other hand, is in series with both diodes. This parallel-series hookup lets the two signals mix in the special way that produces an audio signal.

The special way mixing takes place in **fig.** 2 as the result of how the signals are brought together. The carrier signal is fed to the stage in a mode different from that of the sideband. The mixing generates a product of the two signals instead of sums and differences. (That's where the name product detector comes from.)

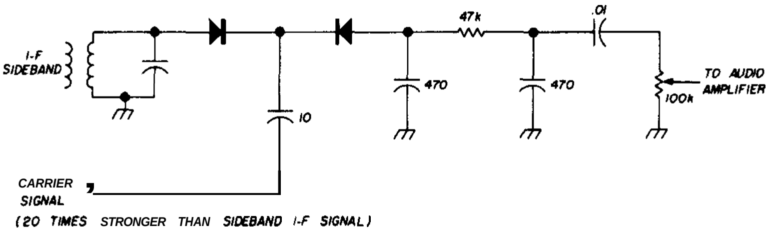
Furthermore, the output signal is taken from the stage in series—a mode opposite to the carrier input mode. This encourages cancellation of the input carrier, keeping it from the output. The product of this mode of mixing, therefore, is a relatively pure audio signal—the recovered voice signals that originally formed the sidebands. Any slight remaining carrier or sideband signal is eliminated by the 470-pF capacitors and 47k resistor.

(The parallel/series method of feeding the two signals into the stage—and of taking the output—should sound familiar if you've read earlier articles in this series. Beginning on page 24 of the May issue, I described balanced modulators in ssb equipment. They also use this two-mode way of handling input and output signals.)

Tubes offer a better means of isolating and mixing (see fig. 3). Furthermore, the tubes can build up the carrier-signal strength. The sideband signal from the i-f amplifier is applied to a cathode follower; that isolates the signal source from the mixing circuit, without adding any gain. The sideband is then cathode-coupled to the mixing tube. Meanwhile, the carrier signal is also fed to the grid of the mixer, and is amplified.

These signals mix within the tube. The output is a product of both signals—a heterodyne product that includes the original modulation that has been carried by the sideband. All rf is filtered out by the pi-network, and clean audio is sent to the audio amplifiers.

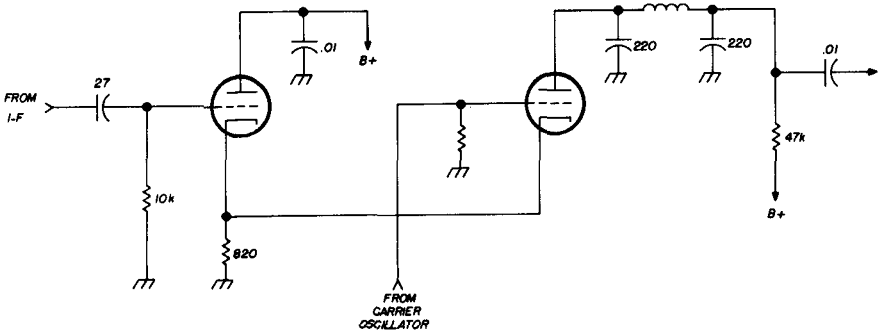
fig. 2. More elaborate le circuit for ssb de-ulation; circuit has fundamental resemblance to balanced demodulators.



balanced ssb detectors

You've already seen a simple single-sideband detector with characteristics ap-

fig. 3 Two-triode version exemplifies principles of sideband detection and is used commercially.



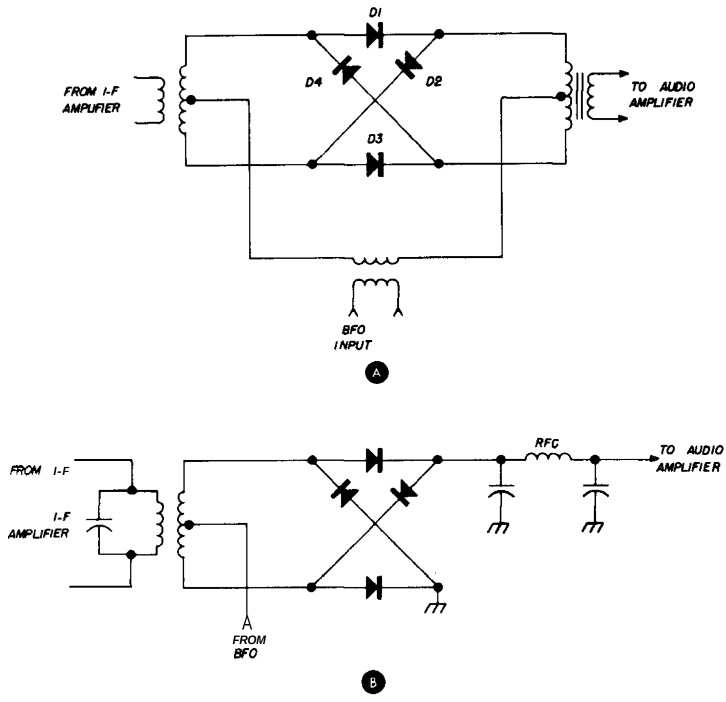
proaching those of a balanced modulator. The fact is, you can use a circuit very like a balanced modulator to demodulate sideband signals.

If you study the stage in **fig. 4A,** you will see that it differs only slightly from a balanced modulator. Both input transformers are rf types, whereas in a balanced modulator one of them would be an audio type. The output transformer in **fig. 4A** is an audio transformer; in a balanced modulator it would be an rf type. What you see in **fig. 4A,** therefore, is a balanced demodulator.

The diodes are in what's called a **ring** arrangement; if you trace through them, you'll see they are essentially in series— 'round and 'round. The name of the stage is **ring demodulator.**

Its operation is exactly what is needed to recover audio from sideband signals. It accepts the sideband i-f signal and the carrier signal (from a bfo or a carrier oscillator), reinserts the carrier so the signal can be demodulated, and then couples out the resulting audio signal while eliminating the carrier. The action in a balanced detector is thus very like the action in a balanced modulator; whatever signal is fed into the stage in a mode opposite from the output mode is canceled. This helps considerably in a ssb detector, since the carrier must be applied at a level so much higher than the level of the sideband.

fig. 4. Ring type ssb demodulators. Primary version in A Is balanced demodulator; eliminating the costly and bulky transformers doesn't alter stage operation (B).



The ring demodulator can be simplified. Transformers are costly and bulky, and any circuit alteration that eliminates them has an advantage. An altered version is shown in **fig. 4B.**

Major characteristics remain. The solid-state diodes are hooked in a ring, the i-f sideband signal is applied in push-pull, and carrier signal is applied in parallel. With the bottom of the sideband-input transformer grounded (instead of the center tap), ground is made one side of a push-pull arrangement; the output is therefore effectively in push-pull, even though it is single-ended for any circuit following. The effect is thorough demodulation of the

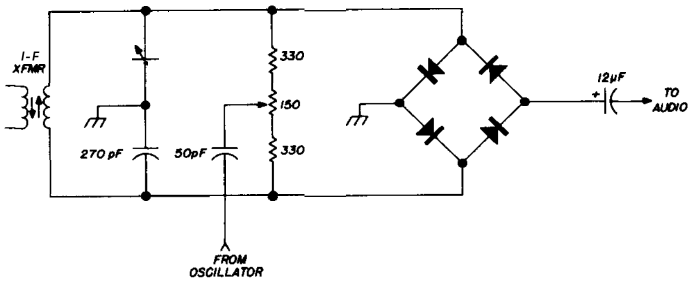


fig. 5. Commercial version of ring demodulator looks slightly different, but uses all principles of others.

sideband signal, with the carrier canceled in the output. The pi-network eliminates any slight rf that remains.

Does anyone use the balanced sideband detector? Yes. One version is part of the Sideband Engineers SB-34 transceiver. There's a schematic of the stage in **fig. 5.** I've redrawn the ring circuit to simplify the looks of the stage for you, but operation is the same as already described. The carrier signal, which in this case comes from the oscillator that generates the initial carrier for the transmit function, is fed to a resistive balancing network; the resistors also isolate it from the ring diodes.

The carrier is applied in the parallel mode, as you can see; the sideband input is push-pull, because of the "phantom" center-tap ground point offered by the ground connection between the two capacitors. In the ring circuit, input and output connections are the same as in **fig. 4B;** you'll see it if you trace them carefully, even though they may look different at first glance.

a one-transistor version

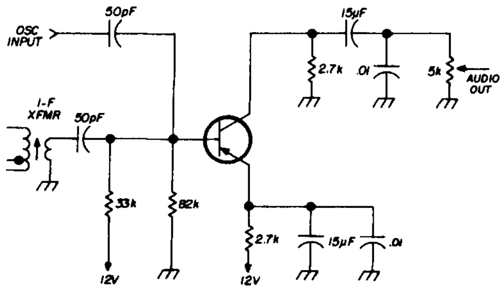
Diode sideband demodulators are all solid-state, since almost no manufacturer uses vacuum-tube diodes today. Semiconductor diodes are more efficient and less expensive. When you talk about solid state, though, you must include transistors. At least one manufacturer uses a transistor ssb detector.

You can see the circuit in **fig.** 6. This stage is from a Gonset Sidewinder transceiver. The pnp transistor is biased in a way normal for negative-ground power supplies—the emitter goes to the power-supply bus, and the collector goes to ground through its load. The two inputs are not isolated in this particular demodulator. The carrier signal is already amplified before it is applied to the transistor base (through the 50-pF capacitor). It and the sideband signal mix in the transistor.

What keeps this from being a simple amplifier for both signals is the bias level chosen for the transistor. The base-emitter junction is strongly backward-biased; the heavy carrier signal then is amplified class C, which is nonlinear.

Mixing in the base resistor as they do, these two signals generate considerable cross modulation. When the cross-modulation products are amplified by the Class-C

fig. 6. Transistor product detector, with no isolation between input signals. Bias of the transistor is what makes it a demodulator rather then amplifier.



transistor, the audio is easy to separate from the other products of this nonlinear mixer. The 0.01-μF capacitor across the volume control eliminates most of the rf signal that is left over. The original modulation, which has been masquerading as a sideband, is thus recovered.

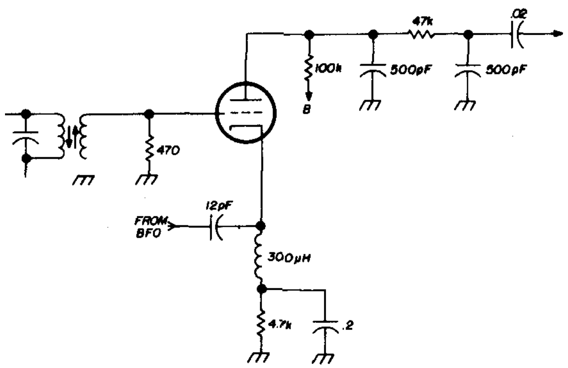
This transistor sideband detector hasn't become popular; no other set uses it that I know of. But transistors can be substituted in any triode-tube demodulator, provided you consider their dc supply requirements and their low impedance.

ssb detection with tubes

In any single-sideband detector system, isolation of the two input signals is desirable. One way to achieve this is in a simple triode product detector—fig. 7. The high-level signal from the bfo (or from the carrier oscillator) is fed to the cathode, using a 300-μH choke as high-impedance—and therefore efficient—input load. The i-f signal, which is the sideband to be demodulated, is applied to the grid, across a low-impedance load: the 470-ohm resistor. This disparity between the two input-load impedances goes part-way toward setting the 10-to-I ratio you want between these two signal strengths.

This triode stage is another reminder of an important principle of product detectors. It isn't always the circuit arrangement that makes a stage detect sideband signals; it is the way the stage is operated. Without the high bias developed by the

fig. 7. Triode product detector with some isolation for signals. Same idea could be used with a transistor.



4.7k cathode-bias resistor, the triode would be nothing more than an amplifier. It would transfer both signals to its output, amplified but otherwise unaltered. It is the nonlinear operating characteristic that permits product detection—and therefore sideband demodulation.

The pi-network in the output of this triode single-sideband detector consisting of two 500-pF capacitors and a 47k resistor eliminates whatever rf products get through the detection process. Good rf filtering is more important in a detector stage like this than in a balanced type, simply because the balanced stage inherently keeps most rf from reaching the output.

The **fig.** 7 circuit is popular because of its economy and simplicity. You'll find it in several Heathkit sets and in the Hallicrafters SR-2000 transceiver.

A tube version like the one I described in **fig.** 3 is part of the Hammarlund HQ-180. The stage configuration is the same; the only differences are in parts values. A triode Colpitts bfo is used in the HQ-180 to furnish the carrier.

The other Hammarlund models revert to the single-tube product detector using a pentode: the HQ-110 and the HQ-145. There is no isolation between the two inputs; both signals are applied to the control grid. Some isolation is achieved by the weak coupling used for both signals. The strong carrier signal is applied through a 3-pF capacitor, small even in this service. The i-f sideband signal is coupled only by a twisted-wire "gimmick" capacitor offering less than 1 pF of coupling capacitance. The high gain of the pentode makes up for any expected weakness in the output—the demodulated voice signal.

using special tubes

You read earlier that product detection is more how the tube is operated than what kind of circuit it's in. That being the case, imagination suggests that tubes with certain special operating characteristics could do an efficient job of demodulating single-sideband signals. That's right. One such tube is the gated-beam detector, a tube with pentode qualities and special construction that makes it particularly suitable for product detection. The beamed electron stream in a gated-beam tube is controlled by both the control grid and a special "gating" grid near the plate. Both grids have exceptionally linear control over the electron stream, and very little effect on each other.

Combine these characteristics into the stage in **fig. 8** and you have a better-than-passable product detector. The sideband signal is applied to the control grid, or G1. The carrier-oscillator (or bfo) signal is applied to the special grid, G3.

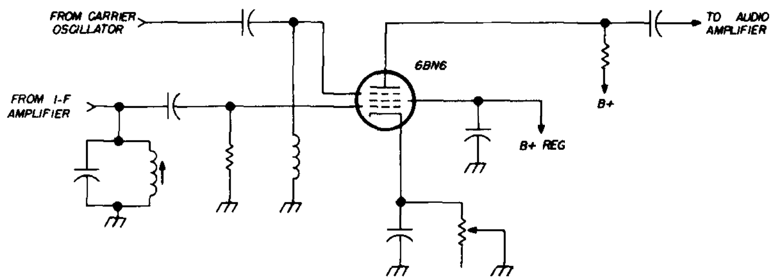
The gated-beam stage is a little tricky to adjust. Unless the bias is just right for each particular tube, considerable output distortion is common. Designers also must carefully work out the strength ratios of signals applied to the two grids. Properly designed and adjusted, though, the gated-beam ssb detector does a good job.

An offshoot of the gated-beam idea is used in the Galaxy V Mark 2 transceiver. The circuit is shown in **fig. 9.** The tube is a 6GX6, a pentode specially designed for broadcast-receiver use in fm detectors. Its non-interacting quality between grid 1 and grid 3 serve sideband detection admirably. As you can see from the diagram, a crystal-controlled oscillator is formed by the cathode/grid 1/screen portion of the tube. The 6GX6 thus provides its own insertion carrier. Sideband signals from the i-f stages are applied to grid 3.

This tube, like the gated-beam detector tube, is touchy. Signal levels must be guarded to avoid crossmodulation that might upset output clarity. The diode between the i-f transformer and the tube input acts as something of a safeguard, to prevent overdriving grid 3. (The diode can't act as a detector because there is no carrier with the i-f sideband signal.)

The connection going to the balanced modulator is shown because the oscillator portion of the 6GX6 circuit doubles—during transmission—as the carrier oscillator. That connection has no bearing on detector operation during recepiton.

fig. 8. Gated-beam detector, used for years in tv and fm receivers, can also make a good ssb detector.

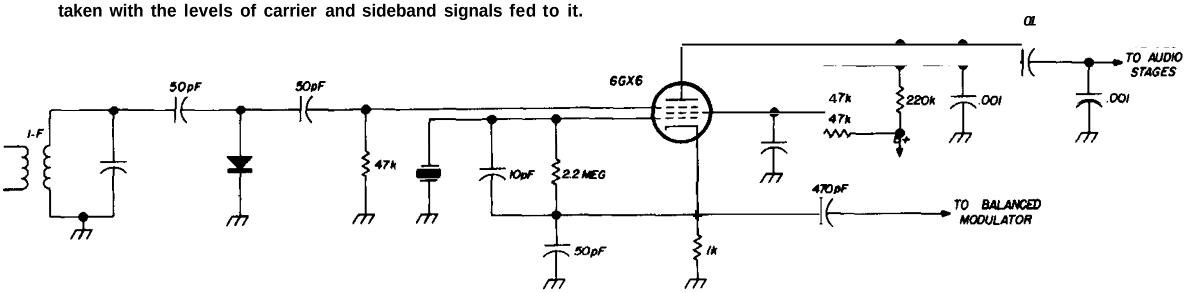


beam-deflected ssb detection

If you did read the earlier article on balanced modulators, you may remember a rather unusual stage using a **beam-deflection** tube. The tube is an RCA 7360, and makes an efficient—though expensive—balanced modulator. This circuit can be altered slightly to become a balanced demodulator, as can other balanced modulators.

The sidebands are applied to the beam-deflecting plates in push-pull. (Supply circuits are not shown to keep the diagram

fig. 9. Another special tube, the 6GX6, makes an excellent ssb demodulator if care is taken with the levels of carrier and sideband signals fed to it.



simple.) The carrier signal is applied to the control grid, which puts that signal in parallel insofar as the rest of the stage is concerned. The demodulated audio output is taken off in push-pull by the simple device of grounding one output plate for audio through C4. (C5 is a different value because its purpose is to eliminate any stray rf signal that might be left over after demodulation.) Output is from the other plate, coupled to the audio amplifiers by C6.

in carrier and sideband signals instead of carrier and voice signals. A balanced modulator becomes a balanced demodulator by that simple switch; of course, you have to change one input transformer from an audio type to an rf type and change the rf output transformer to an audio type.

In demodulators of the product-mixer kind, you have to be sure the carrier signal is 10 or 20 times as strong as the sideband signal. That's because the carrier is always much more powerful than the sidebands in an ordinary a-m signal, and that relationship must be maintained for proper demodulation. If you're working up your own demodulator circuit, you should adjust the ratio between the two signals until you get the best output signal-to-noise quality; at the same time, keep both signals low enough in strength that one doesn't overload the tube(s) you're using.

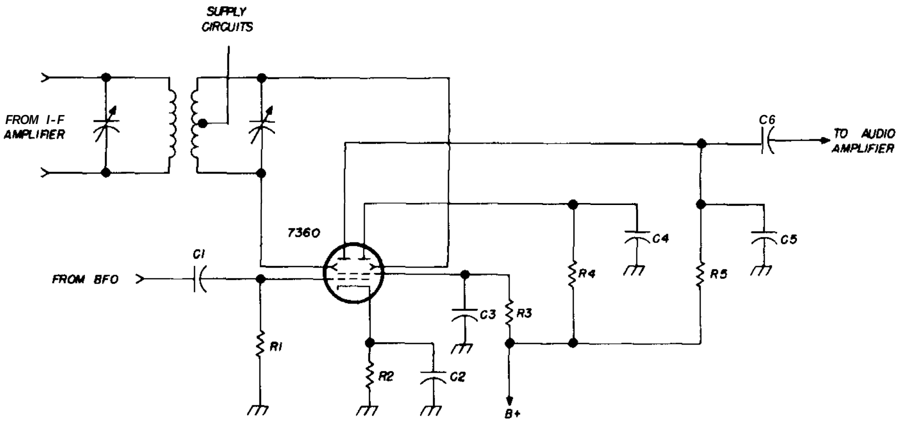


fig. 10. Tube developed especially for use in ssb communications; can be either a modulator or demodulator.

The beam-deflected method of ssb demodulation hasn't been used in any commercial ham equipment I know of. Its expense, though not great, is more than for diode balanced-demodulator systems; cost is always a deterrent to inclusion in factory-built equipment. The circuit is showing up occasionally, however, in home-brew designs. It is efficient, and you might want to try it in a receiver of your own. Operating characteristics for the 7360 can be found in The Radio Amateur's Handbook, in the Special Receiving Tubes table. From those, you can work out parts values.

sideband detectors in general

Summing up the characteristics of various single-sideband detectors, you can draw certain general conclusions. First of all, any ssb modulator can be altered to become a demodulator. You merely feed

With the information here, you should find that sideband detectors have few secrets anymore. You've seen both tube and transistor types, as well as solid-state diode types.

And—speaking of transistors—that's what we'll go into in the next article: transistors in single-sideband equipment. More transistors are being used, so there should be a lot of dope that will help you in your future ssb activities.

ham radio

solid-state antenna switching

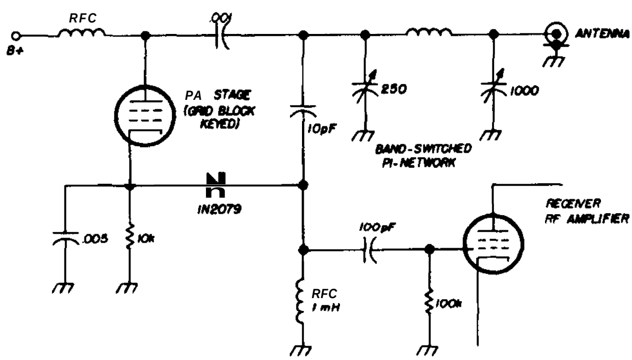
Solid-state antenna switching and how to adapt it to your own rig

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

Solid-state antenna switching has been widely advertised as a feature in a number of commercial transceivers. The advantages are numerous and the schemes used in transceivers can generally be adapted to your station.

One of the advantages of solid-state antenna switching is that the noise and contact problems are eliminated. Also, since the antenna transfer between transmitter and receiver is almost instantaneous, break-in CW is available with the

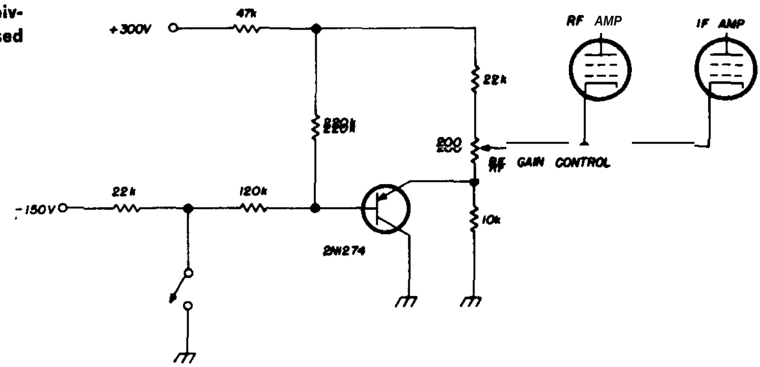
fig. 1. Basic circuit of the antenna-transfer circuit used in the Heath HW-16. The diode is a standard silicon rectifier rated at 500 V and 750 mA.



addition of a simple receiver-muting circuit. As a side benefit, the solid-state antenna switching circuit automatically protects the receiver front end from excessive input which may burn out the rf stage.

This article explains some of solid-state antenna switching circuits which have been used in commercial equipment. The circuits are relatively simple, and once their operation is understood, you should be able to develop a circuit that will operate with any transmitter and receiver combination.

fig. 3. Simple receiver muting circuit used in the Heath HW-16.



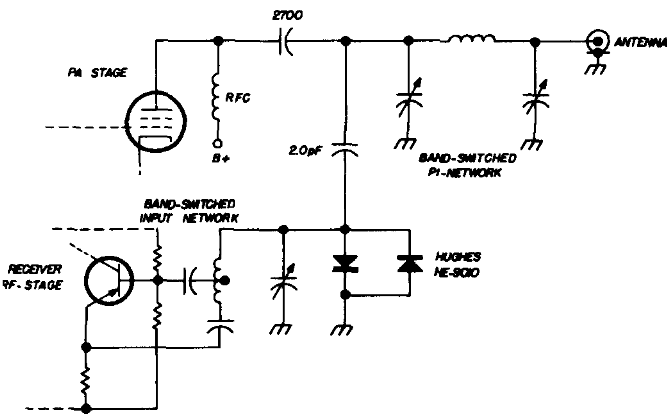
antenna switching

When you hear the term "antenna switching," you naturally think in terms of a relay where the center arm is connected to the antenna and completes the electrical path to the transmitter or receiver. However, this type of antenna switching is rarely used with solid-state equipment. Instead, the receiver is connected to the transmitter and antenna so that when the transmitter is not operating, a low-impedance path exists between the antenna and receiver. When the transmitter is turned on, the receiver is shunted to ground through a low-impedance path that does not affect transmitter loading.

The following circuits should make this clear. However, the idea of shunting rather than switching should be remembered to understand circuit operation clearly. In these circuits, the transmitter remains continuously connected to the antenna, and the receiver input is switched to either the antenna or ground.

If you don't want to use solid-state switching with diode switches and transistor muting circuits, a relay could be used across the receiver antenna terminals. Since the transmitter output is not switched, a small (and inexpensive) relay can be used, even for high-powered transmitters. The quicker switching action of a small relay permits break-in operation at slow to moderate CW speeds.

fig. 2. Front-end protective circuit used in the SB-34.



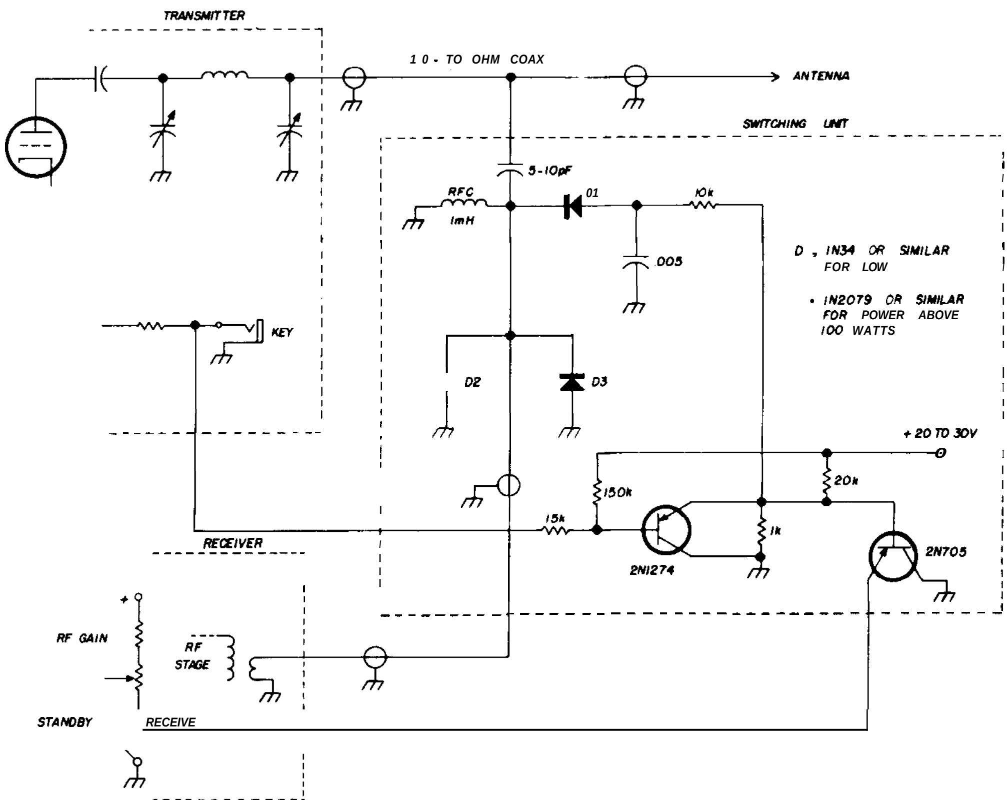
circuits

The basic antenna transfer circuit used in the Heath HW-16 transceiver is shown in fig. 1. It is typical of circuitry which can be adapted to other vacuum-tube equipment. Circuit operation is relatively simple; grid-block keying is used, so when the key is open, the power amplifier is cut off. The voltage appearing at the PA cathode is almost zero — not enough to appreciably forward bias the diode. Therefore, a high rf resistance to ground appears along the receiver line (the rfc presents a constant high-impedance path). The antenna signal flows from the pi-network through the 10- and 100-pF coupling capacitors to the receiver rf amplifier.

When the transmitter is keyed, the PA cathode voltage rises to about 11 volts. This forward biases the diode into the low-resistance region so the receiver is essentially grounded for rf through the .005-μF cathode by-pass capacitor. The 10-pF capacitor is now in parallel with the 250-pF plate-tuning capacitor but produces no noticeable detuning effect.

Fig. 2 shows how antenna switching is accomplished in the SB-34. The receiver input is loosely coupled through a 2-pF capacitor to the pi network. During transmit periods the diodes operate as clippers and reduce the rf voltage across the receiver input circuit to a low value that will not damage the transistor rf stage. The diodes are high-speed computer types manufactured by Hughes.

fig. 4. Switching unit for separate receiver/transmitter setups.



Regardless of the method used to protect the receiver input during transmit periods, it must work in conjunction with a muting circuit that silences the receiver. One circuit that will do this is shown in fig. 3. When the key or push-to-talk switch is open, the transistor switch is biased into conduction and the lower end of the 200-ohm gain control is effectively connected to ground.

When the switch is closed, the 220k and 120k resistors form a voltage divider which puts about 30 volts on the transistor base; then the emitter-collector path is essentially an open circuit. A high positive voltage is placed on the cathodes of several receiver stages and they are cut off.

You can easily adapt the basic concepts of these circuits to suit your own equipment. If an adequate receiver-muting circuit is available, you only have to add the simple diode circuit shown in **fig. 1.** If you want to develop a completely new circuit, you can combine some of the ideas.

An example of a combined circuit is shown in **fig. 4.** This circuit is for use with a separate transmitter and receiver. Some advantage might be gained by coupling the receiver to the transmitter before the pi-network to take advantage of the selectivity added by the extra tuned circuit. However, the constructional difficulties involved when the transmitter and receiver circuits are in separate boxes usually outweigh the gain.

Operation of the circuit is a combination of **fig. 1** and **2.** Diode D_1 presents a high or low impedance to ground depending upon how it is biased by the 2N1274. Diodes D_2 and D_3 are additional protective circuitry for the receiver; they are generally not required for transmitters with less than 100-watts output.

In the receive mode, D_1 presents a high rf impedance to ground. The negative blocking voltage on the key biases the 2N1274 into conduction; this returns the base of the 2N705 essentially to ground so it can conduct and complete the circuit for the receiver rf gain control.

When the key is closed, bias is removed from the 2N1274, and a positive bias is applied to D_1 thus presenting a low-impedance path to ground. The 2N705 is biased so that its emitter-collector circuit presents a high resistance path and the receiver is muted.

construction

Whatever circuit you devise, care in construction is necessary for proper performance. The rf leads should be kept short and shielded although there is no particular restriction on any low-impedance coaxial-cable leads.

The circuit shown in **fig. 4** can be conveniently built and shielded in a small mini-box. If you use a small chassis, it's a good idea to put the rf components at one end of the enclosure and the control transistors and dc connections at the other. Muting can be checked by grounding the receiver input and noting that when the transmitter is keyed the receiver is silenced.

The protective circuitry for the receiver can be checked by measuring the rf voltage across the coaxial line to the receiver (with the receiver disconnected) under keydown conditions. There will be some voltage, but it should be far below the level that would ever damage the receiver circuitry.

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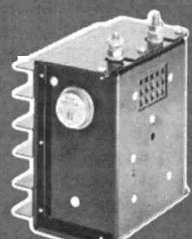

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pi and pi-L networks

for linear-amplifier service

Almost all rf power amplifiers that are used with coaxial feedline use a pi or pi-L network to match the antenna; here's a simple approach to designing these networks

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

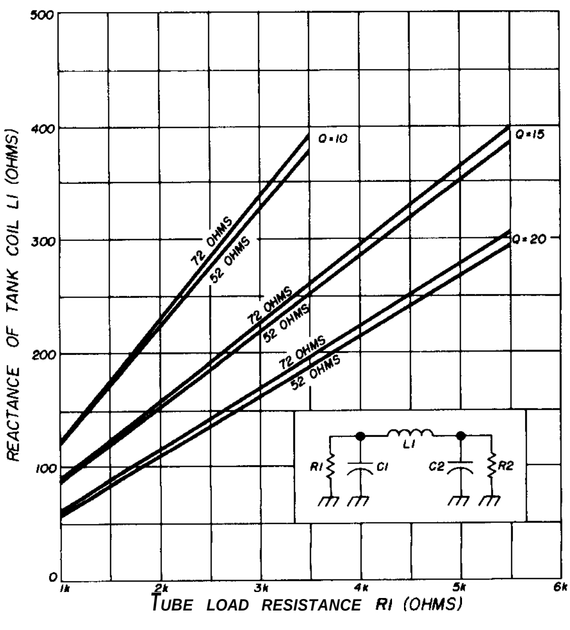
These graphs may be used to determine the component values used in pi and pi-L networks. The graphs cover the most generally used operating Q's, load resistances and antenna impedances. To use the charts, it's only necessary to know the plate voltage, peak plate current, desired operating Q and transmission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

fig. 1. Tank-coil reactance as a function of tube load resistance for pi networks. R1 is the tube load resistance and R2 is antenna resistance.



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:

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

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

2. Determine the approximate resonant load resistance from:

$$R_1 = \frac{\text{plate voltage}}{2 \times \text{plate current in amperes}} \qquad (2)$$

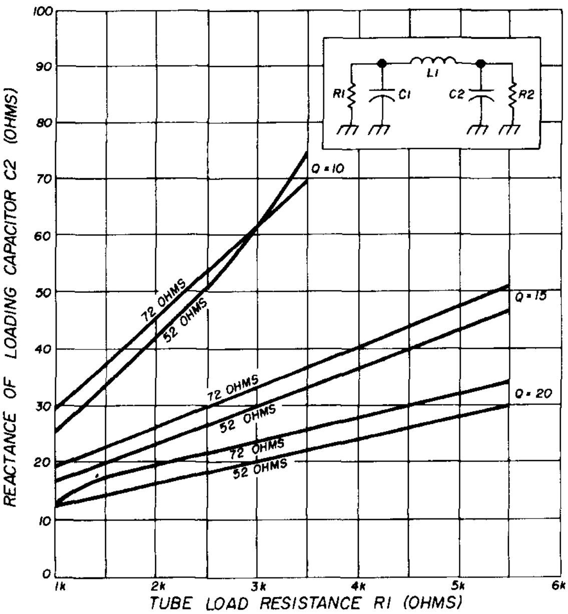
For the case of the 3-400Z's, the load resistance is: 2500/(2 x 0.8) = 1560 ohms.

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

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. 1. For the case of two 3-400Z's operating with a load resistance of 1560 ohms and a *Q* of 15, the reactance of the coil is approximately 120 ohms.

fig. 2. **Reactance of the loading capacitor C2 as a function of tube load resistance for pi networks.**



6. Find the reactance of the loading capacitor (C_2) from fig. 2. For the case of 3-4002's 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. 3. For the case of 3-400Z's 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. 7 and 8 show reactance values of inductors and capacitors commonly used in rf circuits in the h-f 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 μH.

using the pi-L network charts

Fig. 3, 4, 5 and 6 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.

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 out-

lined 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. 4.** For the case of the 3-10002 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. 5.** In this case, the reactance is about 47 ohms.

7. Find the reactance of the tuning capacitor (C_1) from **fig.** 3. In this case, the reactance is about 150 ohms.

8. Find the reactance of the loading coil (L_2) from **fig.** 6. In this case, the reactance is about 140 ohms.

For a single 3-10002 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; leading 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.** 7 and **8** 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 pH; L network coil (L_2) = 140 ohms = 6.5 pH.

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.

ham radio

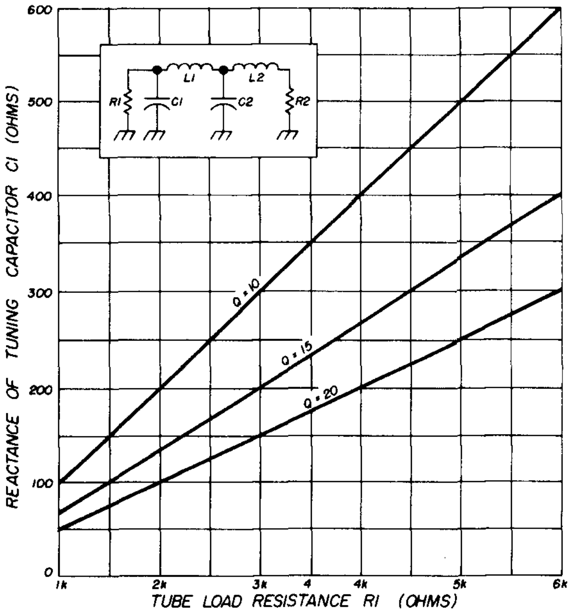


fig. 3. Reactance of the tuning capacitor C1 as a function of tube load resistance for pi and pi-L networks.

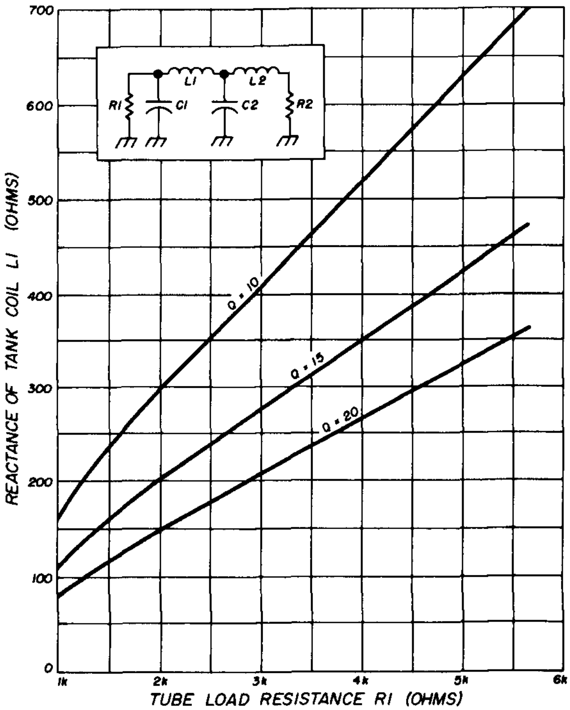


fig. 4. Tank-coil reactance (L1) as e function of tube load resistance for pi-L networks.

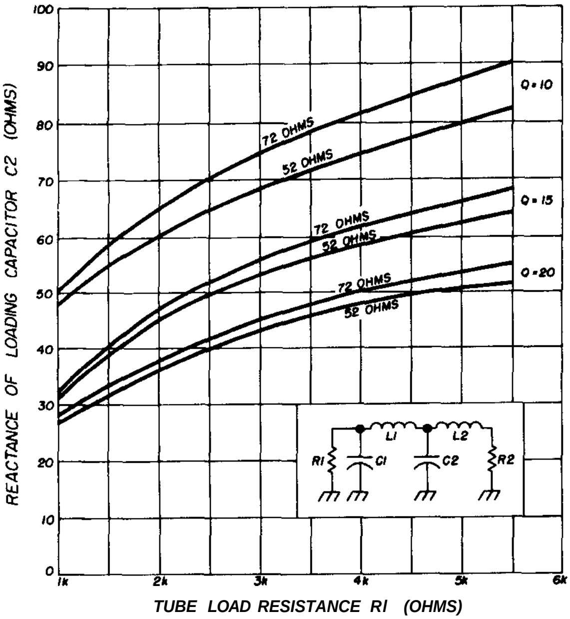


fig. 5. Reactance of the loading capacitor C2 as a function of tube load resistance for pi-L **networks.**

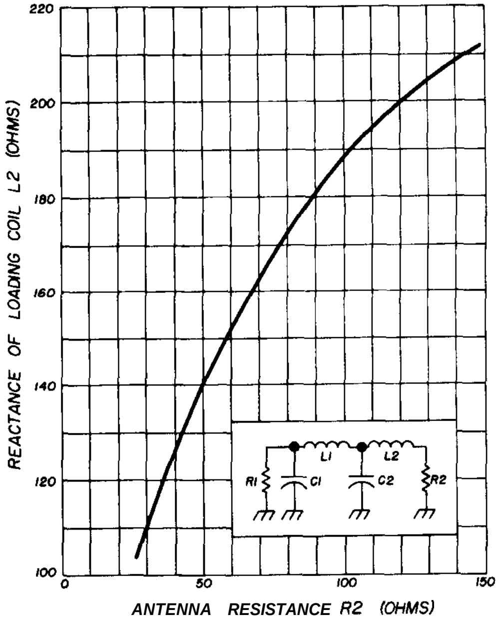


fig. 6. **Reactance** of the loading coil L2 as a function of antenna resistance (R2) for pi-L networks.

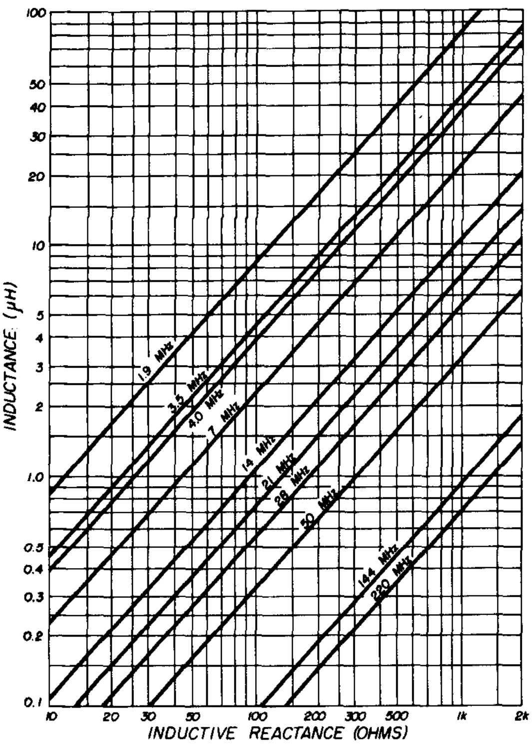


fig. 7. Reactance of inductors commonly used in the amateur bands from 1.9 to 220 MHz.

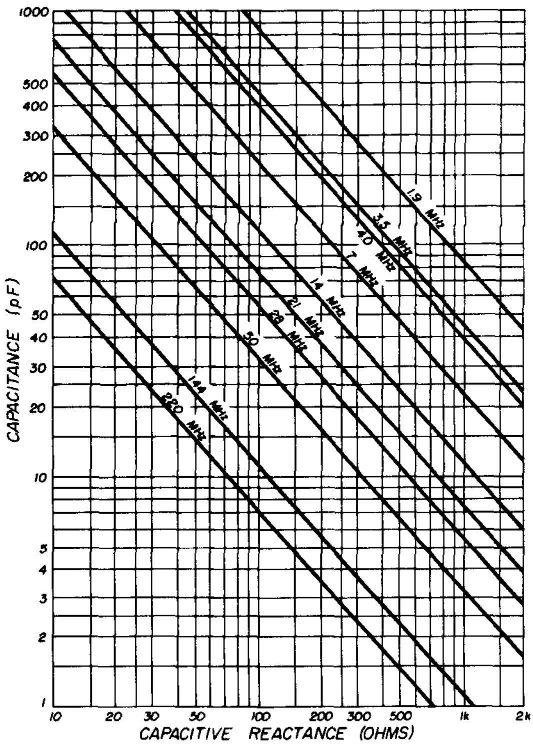


fig. 8. **Reactance** of capacitors commonly **used** in the amataur bands from 1,9 to 220 MHz.

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calibrators and counters

How to use inexpensive integrated circuits for accurate frequency-measuring equipment

Bill Conklin, K6KA, Box 1, La Canada, California 91011

I discussed amateur frequency measurements last month[1] and concluded that a temperature-controlled crystal in a calibrator is a pleasure to own, and calibrator harmonics every five or ten kHz are almost a necessity. I also suggested that a frequency counter is very useful for frequency measurements with accuracy better than one or two hundred Hz. These functions can be provided with convenience and simplicity by using the nonlinear integrated circuit (IC) known as a J-K flip-flop in single units, duals or decade dividers.

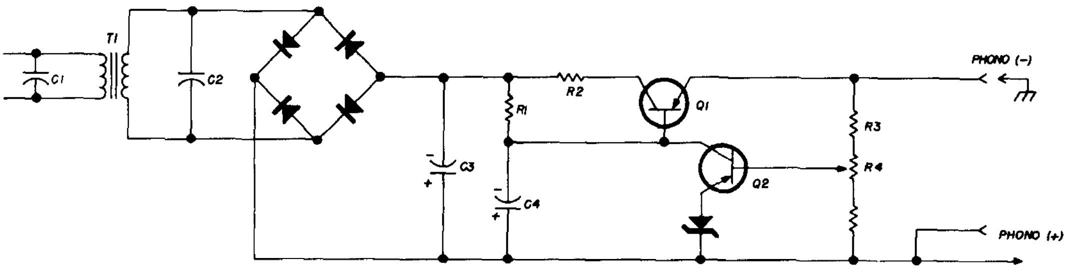
Building a calibrator and counter is so simple that it's completely dwarfed by one's inertia or stagnation in getting started. This is particularly true because we don't always know what we should about the IC's and must spend more time studying them than building. It is the purpose of this article to assist you in the planning job, and to demonstrate how easily the construction can be accomplished afterward.

taming regulated supplies

Regardless of the voltages required, there is usually a power supply to design. Nonlinear IC's are not very critical as to voltage and don't need closely regulated power. However, there should be good regulation of the power to the crystal oscillator and a way of protecting other units from overvoltage in case of loss of load. Unfortunately, most available transformers don't furnish the desired voltage; this leads to the need for a way of reducing the output without having it soar in the absence of part or all

Articles discussing transistor-regulated supplies, or practical circuits, are listed in references 2 through 19.

It is not unusual for surplus zener diodes to vary considerably in their performance, to operate in reverse or not



	5 V	12 V
T1	12 V	24 V
C1	.1	.1
C2	.01	.01
C3	large	large
C4	100	100
R1	350	220
R2	25 ohm, 25 W	200 ohm, 10 W

	5 V	12 V
R3	68	1000
R4	—	350
R5	120	600
D5	3.3 V	5.7 V
Q1	2N301	2N301
Q2	2N404	2N301

fig. 1 Regulated power supplies

of the load. The simple answer to this is some form of regulation which can be as little as a zener diode and resistor.

table 1. J-K next-state truth tables showing output after next clock pulse. A applies to MC853, 9093, LU321A and 8822; B applies to MC790/890, μL923 and μL926.

J	K	Q_{n+1}
0	0	Q_n
1	0	1
0	1	0
1	1	$\overline{Q}$*

A

J	K	Q_{n+1}
1	1	Q_n
1	0	1
0	1	0
0	0	$\overline{Q}_n$*

B

• Toggle condition to divide by two.

at all. More than one may have to be purchased for a particular job. On the other hand, it's not much more expensive to build a series-regulated supply which can be adjusted over a considerable voltage range. Taming one or two of these turned out to be an education and caused a delay in the much shorter job of wiring up the IC's. Some comment on **fig. 1** may prove helpful.

Many sources say that there should be a resistor in the transformer lead to protect the rectifier diodes. I found that the transformer resistance was sufficient—provided the current rating of the silicon rectifier diodes is around an ampere. These have always survived the short-circuit current through a very large surplus electrolytic filter capacitor. Good diodes are available as "glass amp" and other types at a few cents each from John Meshna, Poly Paks, Weinschenker and others. Be sure to test them; some operate in the reverse direction.

Unless your filter capacitor has a grounded mounting, either pnp or npn transistors can be used by putting the proper polarity on them and grounding either power supply output. The selection of transistors is covered in theory and arithmetic in the 12-page pamphlet, "Transistorized Voltage-Regulator Application Guide—ICE-254," available from RCA distributors. Practical circuits are found in the **CE Transistor Manual.**

table 2. Binary-coded decimal count-sequence truth table.

	output			
count	**D**	**C**	**B**	**A**
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1

Surplus 2N404-type transistors, Poly Paks number 14L492, are satisfactory for the amplifier job in the supplies that do not require a drop of more than ten volts or so. A good selection of low-voltage ten-watt zener diodes is available from Solid State Sales, so smaller ones may be avoided. The temperature-compensated GE type-RAI reference amplifiers are alsc attractive for output voltages of 8 or more with precision control.

A bag of Poly Paks number 14L404 transistors provides the large ones for handling the series current and as amplifiers where a large voltage drop is involved. These are similar to the RCA 2N2869/2N301 three-ampere, 30-watt audio device in a TO-3 case. They may be used in the Darlington-configuration to increase the effective h_{FE}; with emitter-equalizing resistors they can also be used in parallel to handle more current.

Nothing here has more than taken the chill off of them. Don't forget to take all the burrs off the chassis holes and check for case-to-chassis shorts if they are mounted with mica kits.

The rectifier usually puts out about twice the ultimate voltage in order to provide room for control. The greater this voltage difference, the warmer the transistors get.

The resistor between the series transistor collector and its base must not be too low or the amplifier transistor will get hot. A high-wattage resistor directly at the series transistor's collector will provide short-circuit protection, which is a necessity. Its size depends on the maximum current to be taken from the power supply. See GE's manual.

The zener diode should have its resistance checked in both directions. As a test, put a large resistor in series with it and feed it some dc so the regulating performance of the zener can be measured

table 3. Soma available J-K flip-flops and related devices.

	frequency limiting min	typ.	type	cost	V_{CC}	J-K to toggle
Motorola						
HEP558			single	$5.99	8	
MC1013P		85	single	4.55	–5.2	L
MC1027P		120	single	9.60	–5.2	L
MC1018P			translator	3.45	–5.2	
*MC838		30	decade	7.55	5	
MC853		12	dual	4.75	5	H
*MC790P	8		dual	2.00	3.6	L
MC890P	8		dual	2.30	3.6	L
Texas Instruments						
SN7490N	10	18	decade	11.10	5	
SN7473N			dual	5.90	5	
SN7476N			dual	6.20	5	
Fairchild						
9093		2	dual	7.45	5	H
9958		2	decade	11.20	5	
9960			decade	15.70	5	
μL923		2	single	1.50	3.6	L
μL926	8	20	single	4.50	3	L
Signetics						
*LU321A	10		dual	2.48	5	H
*N8280A		35	decade	8.90	5	
*N8822A	10	25	dual	5.90	5	H
N8826A	25		dual	5.90	5	H
N8829A	15		single	3.65	5	H
N8H22B	50	75	duel	15.30	5	
8T01B			decade	15.00	5	

* attractive

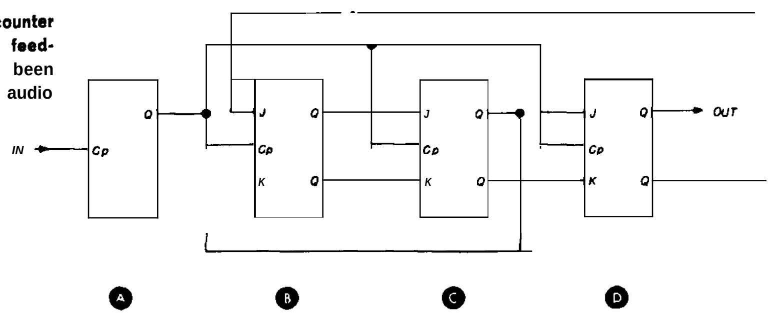
with both voltage polarities. The voltage drop across a reference zener may be around 40 to 80 percent of the desired power-supply output voltage.

The drvider across the output can be selected to provide a light bleeder load together with a tap that will determine how much higher the output voltage will be compared with the zener voltage. A potentiometer connected in series with

5-V transformer for 6-volt lamps. I used an expensive transformer (Lafayette 33H-3702, $1.39), although a new-condition Navy surplus one now furnishes all the power required for the equipment.

Some of the zeners may produce rf interference. A ceramic or mica capacitor directly across an offending zener diode will prevent this. Both polarities of each power supply are terminated in phono

fig. 3. This decade-counter connection without any feedback NAND gates has been tested successfully at audio frequencies.



this divider permits adjusting the output; this is desirable during tests but can be avoided in the final assembly unless variation is required later, such as to provide a fine-frequency adjustment of the crystal oscillator. It is a little difficult to pick the two resistors in the divider because of resistance inaccuracies.

The power supply for the 6-V lamp count-indicator can be simple, as shown in **fig.** 2. It has a resistor, and **5-** to 5.7-V ten-watt zener mounted on a bent sheet-aluminum angle as a heat sink. It is used only as a limitation on the maximum voltage that will reach the lamps. Their life depends on operating below their rating by at least a few per cent. They work fine with 4.5 to 5 volts on them. This arrangement permits using a 6.3- or

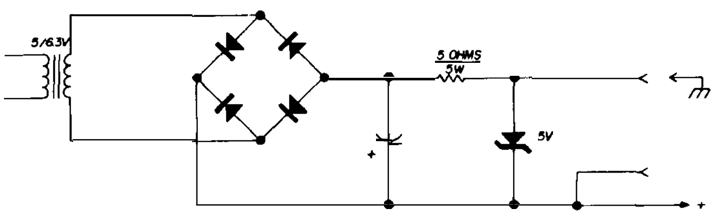
jacks for occasional use on experimental equipment. A shorted phono plug in the negative jack, captivated to a soldering lug under the jack, completes the connection to the crystal oscillator, the calibratorhime-base and the indicator lamps.

preliminary explanations

To help you select the IC's, some explanations are in order. Let's go through this. The arithmetical notatron 1•2•3 or A•B•C refers to AND gates which occur at the input of some flip-flops. All inputs must be activated before a pulse can pass through the unit.

When you see 1+2+3, it refers to OR gates, any input to which will pass to the output—but this may not concern us here. The AND gates are built into J-K

fig. 2. Voltage-limited indicator-lamp power supply. Use phono plugs for output connectors.



flip-flops to make them flip in alternate directions from successive input pulses on a single terminal, which is referred to as **toggling.** This is the condition that divides by two.

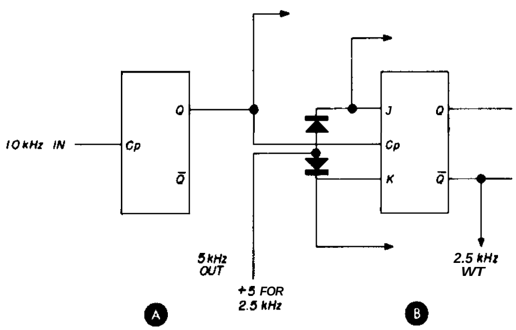


fig. 4. Proposed method for inhibiting the decade count with +5 volts through diodes to the J and K inputs on flip-flop B in fig. 3.

It is well to recognize some of the designations used on the terminals. C_p and T mean clock pulse or input which is fed to both sides of the flip-flop to cause it to toggle (usually on its falling or negative edge) and divide by two when other inputs permit it. J and K inputs determine whether it will obey the next clock pulse. S and C, R and S, R_c and S_c, all refer to somewhat similar inputs. R, S and C are also used to indicate reset, set, and clear terminals, frequently with a subscript letter D to indicate direct action on the flip-flop regardless of the clock pulses or the other determining inputs.

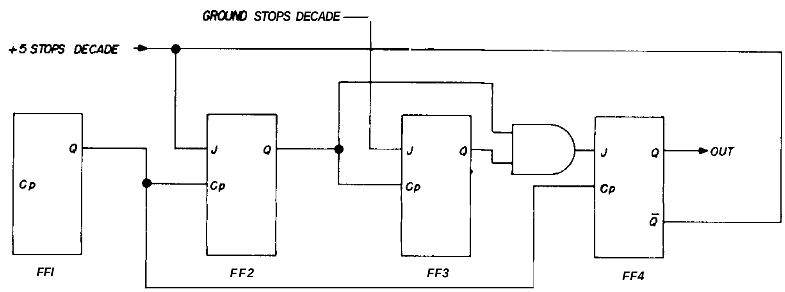
R_0 sometimes refers to reset to a binary zero, and R_9 to a binary count of nine, the latter being of no interest to us. Q or 1 is the normal output; $\overline{Q}$ (not Q) or 0 is the reverse of the normal Q or 1 output. V_{CC} is the power input to the collectors which is generally referred to as plus compared with ground; V_{EE} is also used in (M)ECL units other than the HEP558 where the power supply is minus on V_{EE} (the emitters) and V_{CC} is grounded.

Truth tables indicate what conditions will produce what results.3 The J-K next-state truth table in **table 1** shows whether the auxiliary J and K inputs should be at high or ground voltage to cause the FF to toggle with successive clock pulses. There are several different types of count-sequence truth tables; the one in **table 2** shows the indicator lamps that light in a simple additive 8, 4, 2 and 1 binary-coded decimal.

Another[25], has more lamps on and deviates slightly from the simple addition of binary numbers to convert from the lighted lamps on an additive basis to the digital numbers of zero through 9. Still another differs greatly[19], tending to limit its use to the provision of decoding gates for full 0-to-10 decimal indication.

For frequency division, several kinds of FF's will work. The inclusion of AND gates to obtain decade operation is another consideration that makes the J-K flip-flop desirable.

fig. 5. A decade counter circuit using Fairchild 9093 flip-flops that requires an AND gate. Two ways are shown for stopping decade action to obtain a 2.5-kHz output from the 10:1 decade.



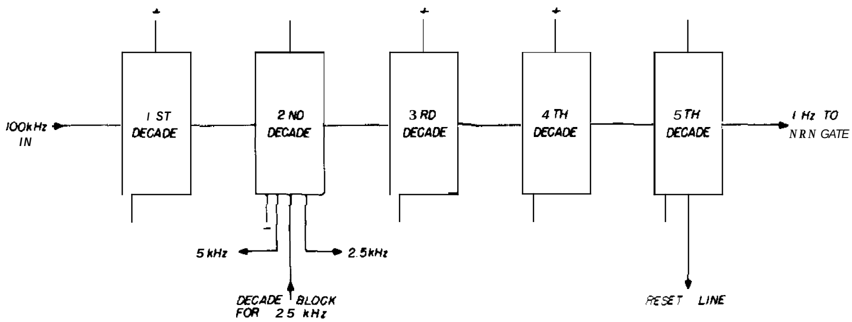
All counting flip-flops and the time-base gate must reset to zero just before a count starts. Thus, a reset or clear provision is required. However, if a flip-flop has only a **set** provision, this can be used to reset or clear simply by reversing the J and K inputs and the Q and $\overline{Q}$ outputs.

You should also consider the count-sequence truth table that will be produced by a decade divider using a particular type of IC and circuit, because of types of interpretations that may have to be applied to the indicator lamps. Two kinds can be mixed if you don't mind a little confusion. There will be some other points to consider in mixed types.

Table 3 lists many of the available J-K flip-flops, decade counters, translators (if (M)ECL high-speed FF's must feed into the more common saturated-logic units) and decode/lamp drivers for direct conversion to full decimal read-outs. Inasmuch as the prices frequently decline 20 percent for quantities of 25 or more, the single and the dual J-K flip-flops for an entire calibrator and counter may cost less than indicated.

The MC790P operate only down to 15° C; many may prefer the MC890P which operates to freezing. The 9093's that I used operate from sine, sawtooth and square waves down to several cycles per second. The 8822 made by Signetics Corporation operates independently of fall-time of the clock pulse. However, most of the other units have specifications on fall-time which may require some form of trigger or squaring amplifier ahead of the first calibrator and counter FF's for proper operation. See references 2, 4, 13, and 20 through 24. Try them first without it, and let me know what happens.

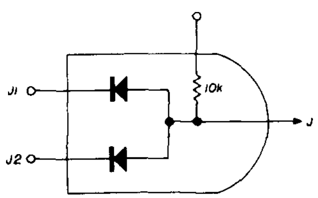
fig. 7. Block diagram of the calibrator time-base board. A shorter count delay results when the reset connection is moved one decade to the left.



available units

It is a convenience to have the IC's operate at the same voltage. The upper frequency limit desired for a counter may lead to selecting a more expensive first FF or decade, which could have a different requirement. If so, two or more series-regulator power supplies can be fed from the same rectifier.

fig. 6. Simple diode-resistor AND gate used with the circuit of fig. 5 to obtain decade operation.

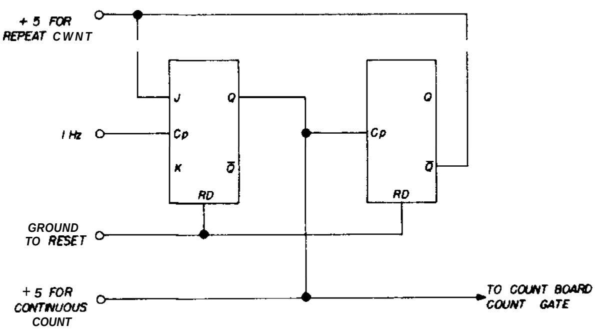


using the flip-flops

The most simple IC for calibrator/time-base and counter use is the decade. As indicated in table 3, these cost several dollars more than two dual J-K flip flops. They do not give access to the J-K terminals, so some provision must be made in the first counter decade to gate it. Other than putting a gate in front, an alternative is provided by the Texas Instruments SN7490N where the divide-by-two FF can be wired for use elsewhere; a separate high-speed FF with J-K inputs can be used as the initial divide-by-two part of the decade ahead of the remaining divide-by-five part. On the other hand, a quad-gate IC will do this job and do three other jobs elsewhere if desired.

You'll note that some units require the extra inputs to be grounded, and some require them to be brought to the V_{CC} or left open to toggle and divide the frequency. These two families require different frequency-division circuitry. Several articles have appeared giving the inter-

fig. 8. Time base and inhibiting connections.



connections for these types. See references 3, 4, 5, 13, 19 and 21.

In general, there must be a connection (internal or external) between the input and the output such as from J to $\overline{Q}$ and from K to Q to assure toggling. If no clock-pulse input is provided, the J and K terminals generally are connected; the input signal is fed to both of them.

The single or dual J-K flip-flop for decade operation requires some form of a feed-back connection so that four FFs will be fooled into counting to ten instead of sixteen. **Fig.** 3 shows a circuit with 9093 dual J-K flip-flops I tested successfully in the audio range. It's probably satisfactory for all decade applications. If the 100-to-1 kHz calibrator/time-base decade uses this circuit, and if 25-kHz output is desired, it can be obtained from the second FF by connecting the J and K inputs of this unit to V_{CC} except during counting.

If only one of each of these inputs is available, two separate switch contacts might be avoided by feeding the plus voltage through separate diodes to the J and K terminals so that no back voltage can feed across. This is shown in **fig. 4.**

The above decade circuit is complicated by many J and K input connections. I will explain another divide-by-two and divide-by-five circuit to accomplish division by ten with 9093 FFs that I use. Refer to fig. **5.**

This type operates by inhibiting an input to FF1 at the ninth and tenth counts, and by resetting FF4 on the tenth count or after every even-numbered count. FF2 and FF3 act normally through the 8th count. Then, the $\overline{Q}$ output of FF4 puts a 0 voltage on the J input of FF2, holding FF2 and FF3 at 0 during the tenth count when FF1 and FF4 also go to 0, ready to start at the beginning. Inasmuch as FF4 is fed from the output of FF1, but must not flip prior to the 8th count, an AND gate is placed at the J input of FF4.

This prevents FF4 from going to 1 until after both FF2 and FF3 are also at 1 at the sixth and seventh counts. **Fig.** 6 is the simple AND gate I used and consists of inexpensive surplus diodes and a resistor of about 10k ohms. Some FFs already have extra inputs to do this. The gate must be capable of operating up to about one-half the input frequency of the decade.

A ground on the J input of FF3 (or +5 V on the J input of FF2) in **fig. 5** will stop the decade action so that 5- and 2.5-kHz outputs can be obtained from the Q or $\overline{Q}$ outputs of FF1 and FF2 in the 10-to-1 kHz calibrator/time-base decade for frequency-measuring purposes.

The equipment I use has a Monitor Products oven-controlled crystal oscillator with its output fed directly to the first calibratorltime-base FF. The output of that is fed to the input of the next, as shown in **fig.** 7. Wiring was done with small, flexible tone-arm twisted pair until the supply ran out, then with Japanese earphone twisted pair, which is larger and retains its kinks.

When the frequency has been divided to one cycle per second (1 Hz), it's necessary to add one more flip-flop to produce a time gate that opens for just one second, putting "toggle" voltage on the J and K inputs of the first count FF, then closes for one second. It is also desirable to provide still another FF (the second half of a dual unit) to permit locking up the time gate after the one second that it is on, so that the indicator lamps can be read.

After reset, the first negative slope from the 1-Hz decade in the calibrator/time-base board turns on the time-gate FF, putting plus voltage on the J and K terminals of the first count FF, which had been held low after the gate was reset. The second negative slope turns it off one second later, but this flips the one-shot FF so that its $\overline{Q}$ output goes to 0. This is connected to the time gate's J input and thus stops any further gating until reset. Fig. *8* shows this circuit and the options provided by the count-selector switch.

I found that feeding the crystal oscillator output to a receiver caused a slight shift in frequency. As a result, the first output to the harmonic-selector switch is connected after at least one FF. The 10-kHz signal is available, and higher frequencies, but 5 kHz is more useful in my Collins receiver. A 2.5-kHz signal is also fed to the harmonic-selector switch, but produces 1 kHz during a counting condition of any type—repeat one-second counts, one-shot count, or continuous count.

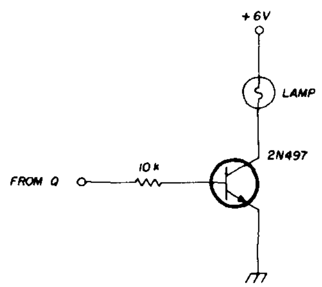
The harmonic-selector switch connects a ground to the J input of the third FF of the 10-to-1 kHz decade, stopping the divide-by-ten operation in the reset and indicator-lamp-off positions of the count-selector switch, thus making the decade divide-by-two twice, producing the 2.5-kHz output from the 10-kHz input. This is desirable with selective receivers to ensure that a suitable beat note is available at every frequency, including gaps left by lsb and usb selection in the receiver. The 1-kHz output during a counting period is mainly a novelty, but its 30,000th harmonic is strong and could be used for l-kHz receiver calibration.

The crystal-oscillator output lead should not be cabled with ac wires unless you use RG-174/U. It might be just as well for the other leads to the harmonic-selector switch and the receiver's antenna input to be coaxial cable or kept out of the cabling, although this has not caused any trouble so far.

electronic counter

The frequency to be counted, such as an audio oscillator, was originally brought directly to the first J-K input of the first FF to provide this function when being driven directly by the time-base gate. It was found

fig. 9. Lamp driver and indicator-lamp circuit.



that a subsequent FF can also be gated to prevent strong input signals from activating subsequent decades when the time gate is off.

Additional decades are added until the output reaches the maximum read-out frequency at which the counter will operate at if anything beyond audio frequencies is of interest. The first decade can be an especially high-frequency unit, and it can be selected so that its first FF is capable of toggling at the highest frequency you're interested in.

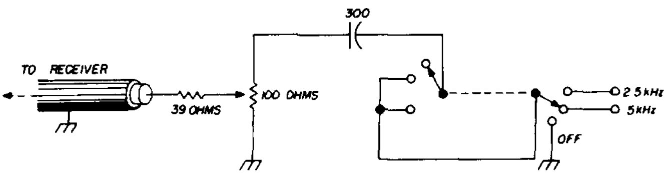
It's necessary to provide a reset or clear function for the counting decades so that the count will always start from zero. It is also convenient to connect this reset fine to the time-base gate and to the associated one-shot FF that locks it up when a single one-second count is desired. Furthermore, the line can go to some of the decades in the calibrator/time-base board to set or reset them to delay opening the gate until all switching transients (if any) have decayed.

count indicator

The FF outputs from each decade can be decoded with diodes and resistors to drive numerical neon read-out tubes costing $8 or more per decade. IC decoder-drivers are listed in table 3.

After much consideration, I used the simple, convenient and inexpensive indicator suggested by Phil Brassine, K7UDL. This has four lamps per decade, the lamps from left to right indicating 8, 4,

fig. 10. Harmonic selector switch with provision for "anti-leak".



2, and 1 in binary-coded decimal form as shown in table 2. These can be converted to a single digit by inspection, or by adding these digits. Then, more of these four-lamp sets are added, one for each decade to be indicated. Six sets will count to a million, or 1MHz. This is satisfactory for slightly higher frequencies, simply by watching the number of times that the count restarts during the one-second counting time. A seventh bank should suffice up to 30 MHz.

The lamps are low-current Sylvania 6ESB six-volt slide-base lamps at 40/ each. The unbased lamps at 31¢ would also work. Other voltages are available. Buss-bar wires are run across the board just below the holes for the lamps which are slightly larger than 1/4-inch. By **double** heat-sinking the slide base of the lamps, it's possible to scrape and solder the slide base to the cross buss-bar wire, holding the lamps neatly in the holes from the rear.

Holes are provided for the lamp-driving transistor and the base resistor leads connected as shown in fig. 9. Surplus 5 x 7-inch sheets of thin phenolic material are adequate to mount and hold seven decades of lamps (28 lamps). In the absence of any shielding, some flip-flops are triggered by rf entering the lamp-driving leads when transmitting at full power.

The lamp-driving transistors should not load the FF's heavily. With FFs that have a plus power requirement and output, npn transistors are selected as lamp drivers. New 2N1302 transistors vary a great deal; at least half had too low beta (collector current divided by base current) and required a low value base resistor. This resistance was around 3900 ohms. Other builders have selected the 2N1304.

After considerable checking, the 2N497 was chosen here, rejecting only about one-third of them after buying a bag of clean surplus devices from Solid State Sales at less than 20d each. About half of these work fine with a 10k base resistor, and almost all work with 4.7k or 3.9k. For convenience, 10k resistors are mounted in the indicator lamp board with the driving transistors; another resistor is soldered in behind the board where necessary to saturate the driving transistor and equalize lamp brilliancy. The fourth lamp of the first two decades does not light during a count but does when the count stops, because of the 20 per cent duty cycle.

interconnections

A double-pole double-throw toggle switch turns on crystal-oven heater power alone, or both this and all other power. The lamp power supply is also fed through a contact in the count-selector switch so that the lamps are extinguished except during counts and reading. The best sequence for a power switch is: off; oven heater power; add crystal-oscillator power; add IC power for calibrator and counter; and add indicator-lamp power; in that order. Of this, indicator-lamp power and possibly IC power for calibrator and counter can be in the count-selector switch.

The harmonic-selector switch has an off-position for no signal to the receiving antenna; and then a 5-kHz contact. The final position connects either 2.5 kHz or, while counting, the l-kHz signal. The output from this switch goes to a blocking capacitor and a potentiometer as shown in **fig.** 10. This is desirable to adjust calibrator output to obtain the strongest beat note with an incoming signal. A potentiometer in the regulator divider varies the voltage on the crystal oscillator from about 9.5 to 12.5 volts as a "fine" adjustment of crystal frequency.

There must be a "little red button" for grounding the reset line so that after a fractional second delay, a one-shot one-second count can be obtained without turning the count-selector switch if it's already in the right position. Since it's the only operation needed to make a new count, a phono plug and cable can be used (such as RG-174/U) with the little red button in a small minibox. This will start a new count from any convenient position, even out of reach of the counter itself.

construction

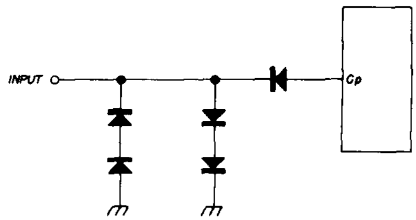
The chassis I used was overly large, since it was an available rack-mounted unit. It extends some inches behind the control panel which already has a 14-MHz front-end crystal filter,[26] audio switching for the loudspeaker and phone-patch, antenna switching and paralleling, and a small, easily-operated push-to-talk toggle for phone patching.

The chassis holds the surplus transformer and large electrolytic capacitors, the controls that are not brought out to the front panel, a Waber in-line grounded outlet box on the rear apron for 120 V to other equipment, and the count indicator, and leaves enough empty space for a good-sized cat.

Except for the heat-sunk transistors, the transformers and electrolytics, the three power supplies are mounted on Vectorboard with small holes on alternate 0.1-inch spacing inasmuch as there was to be some development work, and, later, the addition of the lamp supply. Quantities of Cambion solder terminals are mounted along two edges.

The oven-controlled crystal-oscillator plugs into an octal socket. It feeds the calibrator/time-base board which is made from phenolic material I had left over from "der Kleiner Keyer"[27]. It has a regular pattern of holes drilled to hold resistors and other parts along one edge, with holes here and there for transistors that may be needed in the future. The remainder of the board has sets of 14 holes for the plastic dual-in-line J-K flip-flop IC's. This board is 3½ x 6 inches.

fig, 11. Protective diodes will only permit a negative input pulse to get into the first count flip flop. One pair of silicon diodes may not be required because of the series diode. If higher input voltages are desired, the number of series diodes may be increased.



It is stacked with the count board and mounts on bolts and spacers from the rear apron of the chassis. A part of the count board is reserved for the many terminals needed for the leads to the count-indicator which extend up from the chassis. The new Vector microboards with holes spaced 0.1 and 0.05 inches will accept dual-in-line IC's and transistors directly without any drilling, but they were not on hand at the time.

The 9093 units have a direct-set terminal which, when grounded, makes all FF's go to 1. To provide a reset function, all J's and K's were interchanged, and all

Q and $\overline{Q}$ outputs were interchanged in the counter board and the gate FF's.

It is not easy to test the flip-flops prior to installation. One in each of several dual FF IC's turned out to be inoperative. Replacement was accomplished easily. Dc tests have contact-bounce problems although transistor keying or some of the proposals in reference 28 might help.

The best test method, in the absence of a low-frequency receiver to listen to the normal output frequencies of each FF when driven by the crystal oscillator, is to put an audio frequency in and observe the output at each FF and decade. This can be detected with headphones. The average current from the fourth FF in each decade is low because of the 20 percent duty cycle during a count.

The maximum frequency at which the FF will operate at can be determined by winding several turns of wire around the coil of a grid-dip oscillator and feeding it into the input—with a squaring circuit or trigger in between if required. The 9093 IC's that I used here reached the specified 2.2 MHz. I will probably add a faster decade ahead of the present counter so that the maximum frequency will be above 30 MHz to permit measuring all the oscillators in ssb gear. Then, by putting the counter on the vfo and making an addition (of any mixing crystals), fairly accurate frequency measurements will be available. However, for some purposes, the audio oscillator and oscilloscope will always be necessary.

In view of the direct application of inputs, there is some chance that overdrive might destroy the first IC. This doesn't happen with up to 6 volts from the audio oscillator, and counts are satisfactory down to about 0.06 volts. It is desirable to avoid a potentiometer input adjustment.

Just to be safe, two silicon diodes in series were placed across the input in one direction, and two more in the other direction, with a diode in series with the input to the IC to pass the negative swings of the input pulse. See fig. 11. Possibly the series diode makes one pair from the input line to ground unnecessary, but they haven't been removed yet. The diodes are inexpensive ones from Solid State Sales.

crystal adjustment

There is a little aging in the crystal oscillator until it operates for several weeks. The power-supply voltage adjustment is placed in midscale; the plunger-capacitor adjusting screw is then set for zero beat with a high harmonic of WWV; all subsequent adjustment is made with the potentiometer. Its position is noted for one cycle on each side of zero beat. At the midpoint of these two readings, only irregular fades in the 20-MHz WWV signal affect the receiver S-meter. With no evidence of a beat, it's helpful to be able to misadjust the power-supply potentiometer to determine that there is a beat note with WWV, and then to restore the pot to its previous setting.

There is another method of accurately adjusting the crystal oscillator to WWV without low-frequency transmissions.[29] It is described by Jay O'Brien, W6GDO. A 100-Hz output is fed from the calibrator/time-base board to a triggered external-sync oscilloscope input connection. When WWV transmits the 600-Hz tone for two out of every ten minutes, the scope sweep is set to show just one cycle from the receiver output. The crystal oscillator is adjusted so that the left-to-right movement of one cycle on the screen is less than one-tenth of a cycle during the two minutes. Of course, this must be done when the scope pattern is stable and does not shift left or right due to Doppler effect caused by propagation changes.

By using the first of these two systems to adjust the crystal oscillator, WWV has been measured on 20 MHz from the calibrator harmonic 2.5 kHz lower in frequency. This results in either a zero—or one—cycle error.

Better results probably would require going to a ten-second count period in order to avoid a possible phase error of one cycle in the count. The error does not occur when counting the crystal oscillator or any audio frequency fed into both boards—the result is always the same count as the nominal crystal oscillator output.

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In the ARRL frequency-measuring test, six measurements were submitted for the record. Five had a zero cycle difference from the official measurements made by the umpire. One had an error of two cycles. Propagation conditions can cause similar errors, especially when an elliptical pattern on the oscilloscope between the receiver output and the audio oscillator collapses repeatedly.

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ham radio

short circuit

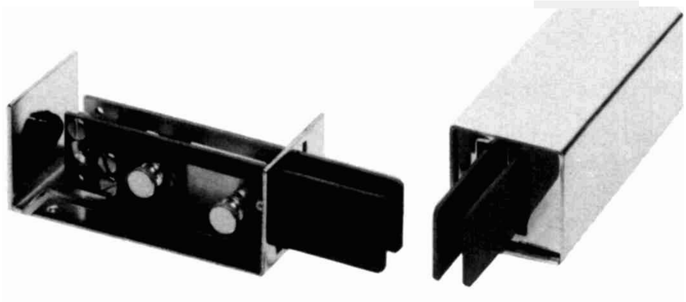
Some errors crept into the excellent article on the kilowatt amplifier for 1296 MHz which appeared in the August issue. First of all, the width of part E in fig. 2, page 12, should be 1-5/16 inch, not 1-15/16 inch as shown. The cavity may still work this way, but the adjustment range of the input probe is insufficient.

Part J of fig. 2 does not allow for the wall thickness of the cavity. It should be at least 1-1/8 inch wide initially and filed flat after soldering it to the top and bottom of the plate cavity. Its thickness depends upon the thickness of blade used to make the cuts in the cavity.

The perforation holes in part D of fig. 2 should be 1/8 inch in diameter over an area of 1-1/2 inch with at least 50% open area. The toggle switch in the schematic labeled "driver off should have two sections in series; a single toggle switch will not stand the voltage.

ham radio

We've re-invented the key



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more electronic units

More on the discussion of basic electronic units started last month

Jim Ashe, W1EZT, P. O. Box 343, Peterborough, New Hampshire 03458

What are **some** of the electronic units that trouble hams who have been working at radio for a while? I think there are four that could use some close attention; the cycle, the mho, the steradian, and the dBm. Each of these is a very useful idea, and any one of them can give you trouble way out of proportion to your real need for the term when you don't see what it's all about.

cycles

Cycle has a curiously indefinite, yet very precise meaning. Roughly, it means "through the sequence of events to where we came in, or to something like it." This tells us the term is applicable only to repetitive events, such as continuous movies, some electrical signals and some mechanical motions. Noise, while full of cycles, is not repetitive so we do not see a cycle of noise.

A definition in the 1968 Radio Amateur's Handbook is confusing, but will get by for some kinds of signals. However, it wouldn't do for the output of my thin-line generator, for example, which never goes negative. And modern communications and industrial electronic circuits can produce an infinite variety of repetitive electrical signals. We need a better definition: a cycle is the interval from any definite point in an electrical signal to the next point having the same amplitude, direction, and general shape. See examples in fig. **1.**

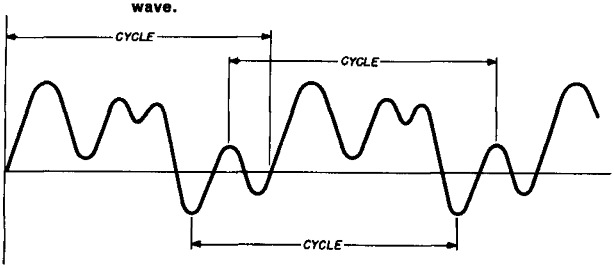
mhos and things

I hope you can stand an awful pun. I'm enjoying this. One of the most incomprehensible electronic units I've seen in amateur radio literature is the micromho.

Yet it is a very good term, once you know how to use it. If you are still working with vacuum tubes (and some of them are pretty good), micromhos are very handy thinking tools. But first we ought to turn them into something more sensible.

Reference to the books tells us a micromho is one-millionth of a mho, which is an amp per volt. So micrornhos are microamps out per volt in, and the familiar old 6AK5 is rated at 5200 microamps anode signal current per signal volt into the grid. But why microamps? The first vacuum tubes must have been very weak. I prefer to think in milliamps and thousands-of-ohms anode load resistances.

fig. 1. This waveform emphasizes the "cycle" as a distance rather than a definite zero-to-zero interval along the wave.



When I look at that 5200 micromhos I can read it as 5.2 milliamps per volt times kilohms plate load resistance, for a rough output figure.

For instance, imagine you have provided a 6EJ7 pentode with a 15k-ohm plate load resistor. For a schematic you can breadboard, see fig. 2. Now, the 6EJ7 is rated at 15,000 micromhos, or 15 milliamps per volt. Assuming a l-volt grid signal, 15-mA signal current times 15k anode resistance gives 225 volts signal out. You can do that in your head. In the real circuit there won't be enough anode voltage swing available for this large output, but we look for a voltage gain in the times-225 ballpark.

steradians

Sounds kind of hairy, doesn't it? If you're a vhf enthusiast you are sure to be concerned with steradians. What is a steradian? It is a way of describing how much sky your antenna illuminates.

Imagine a four-sided pyramid extending into the sky from your high-gain antenna as in fig. 3. How big is the end of that pyramid? Hard to say, not knowing how far away the end is.

But suppose we choose any convenient distance, call it R, and measure the lobe cross section area of the end in square R's. We'll get the same number whether R is measured in inches, feet or miles, and at any distance. Well, that pattern in space, measured in terms of the area in square R's at the end of it, is said to have a certain size or solid angle given in steradians.

A steradian is an area of one square R, whether it is a circle out there, an ellipse, or any other shape. Thinking is similar in style but more complicated in detail if the beam is multilobed in cross section. Next time you bump into 'steradians,' say 'square radiuses' and the shock will be diminished nearly to zero.

dbm

Here is a term that used to seem to me about like a magician's wand: wonderful, incomprehensible. But it's really a basic tool for you, if you're interested in long-distance radio communications. The dBm is not as hard as it sounds: as with other electronic terms, people started using dBm's in their thinking because it made things simpler. Once you see why, you have the whole thing. The dBm idea contains two other ideas tied together, which is the

table 1. Here are the power ratios for some common decibel values. The same figures, with appropriate signs, serve for power gain or loss.

power ratio	dB measure
100:1	20
50:1	17
20:1	13
10:1	10
5:1	7
2:1	3
1:1	0
1/2:1	−3
1/5:1	−7
1/10:1	−10
1/20:1	−13

cause of the comprehension problem.

The first idea is decibel measurement. What is a decibel? A decibel is a size of difference, rather than a size of a quantity or of a thing. A decibel change in signal strength is about the least your ear can detect by listening closely, regardless of the actual signal strength at the time. A 3-dB change will be quite perceptible, but if you are thinking about something else it could escape your attention.

A 10-dB change in signal strength will almost certainly be noticed whether you are listening or not. A l-dB change is about 25% increase or decrease in power; *a* 3-dB change is a doubling or a halving of power, and a 10-dB change is by a factor of ten. See **table 1.**

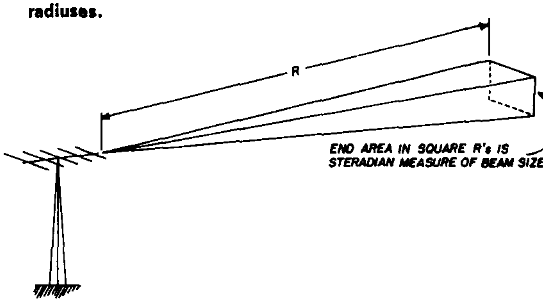
You ought to try this on the bench. It's easy to do and will do more for your perspective than any amount of math or reading. Maybe it will change your thinking about spending more money to double your rig power too. Set up a signal generator, an audio amplifier and a loudspeaker as shown in fig. **4.** Try changing audio levels in decibels and fractions of a decibel. It's surprising, even when you know what to expect and why you get this result. It is because of a built-in agc system in your ears, so that a "ten-times-as-much power" signal is judged to be, maybe, "twice-as-loud."

Now, all electronics signal devices respond to power input, not really to voltage input. It is an old custom, fortunately falling into disuse, to say a receiver responds to so-many microvolts input. The receiver actually accepts a certain definite signal **power** input, amplifies and detects it and amplifies it some more, and puts out a definite larger signal **power** output. How can we describe this performance?

We simply choose a unit of power and specify how much in terms of that unit must be fed to the input in order to get a satisfactory output. One commonly used reference power is the milliwatt applied to 50 ohms. Let's go ahead with a realizable bench test setup and see how it works. If we wanted to test a receiver we would set up things as in **fig.** *5.* Suppose we haven't had too much experience with high-gain circuits?

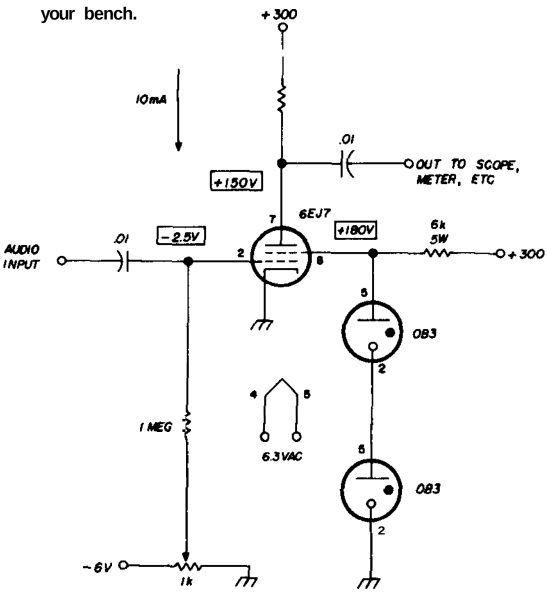
A milliwatt signal is easy to make but hard to measure, so we start with a larger signal and pare it down to milliwatt size

fig. 3. To describe the space filled by an antenna lobe, use radius as the unit of distance and specify the amount of sky filled in terms of square radiuses.



after verifying its power. Now, since the receiver surely has lots of gain, maybe even 1000 (30dB) or more, we apply our milliwatt to the receiver antenna terminals (variable attenuator switches all at zero dB) and get blasted as the receiver

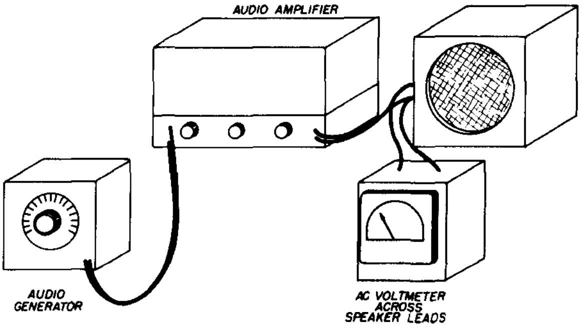
fig. 2. You can use micromhos to quickly estimate stage gain of a pantode amplifier; try this circuit on your bench.



overloads badly. But let's assume for a moment that the receiver did put out the watt specified in **fig. 5.** How would we describe this?

We'd say the input was zero dB different from 1 milliwatt, or zero dBm (into 50 ohms). The receiver would be showing a power gain from input to output of 1 milliwatt rf to 1 watt of audio, or 30-dB gain. This emphasizes that receiver sensitivity and receiver gain are different aspects of receiver performance. And it shows us that we cannot specify receiver gain **or** sensitivity until we have chosen an output power to use in the input-output gain figure.

fig. 4. Simple setup for demonstrating apparent loudness variation with dB change in signal strength.



Typically, receiver sensitivity is the smallest signal that will produce a satisfactory output, but what if you have the gain control turned down? The receiver has *a* sensitivity that is smaller than its best sensitivity, but the same thinking applies.

Now we continue our experiment. At one microwatt input (this is minus 30 dBm, now) we discover we are still overloading the receiver. One millionth of a watt! No wonder some circuits are very sensitive to noise and feedback. Input-output gain would be 60 dB if the receiver were working properly, but a weaker signal is still required.

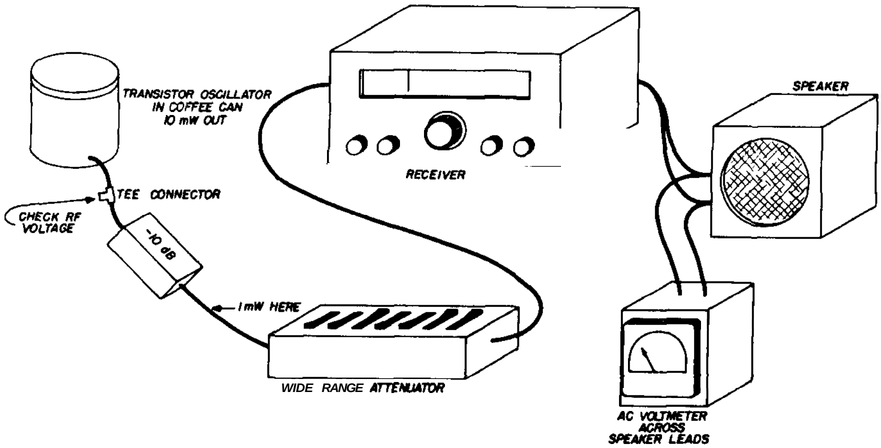
Finally we switch in enough attenuation for a normal I-watt receiver output. Our generator is still feeding its milliwatt into the variable attenuator which now reads 110-dB attenuation. These are realistic figures for a rather average receiver. At last we know our receiver has a sensitivity, **at** certain control settings, of minus 110 dBm, and its over-all gain is 140 dB.

A lab visitor says, "Please translate that into microvolts." Converting from dBm to watts we have 10^{-14} watts input, and since $E^2 \approx WR$, we get about 0.7 microvolts across 50 ohms receiver input resistance.

I expect a number of you will see how well this procedure settles questions about receiver performance, and will want to start doing tests like this. Let me pass on a few thoughts.

Start with a signal large enough to mea-

fig. 5. Test setup for evaluating receiver performance



sure. A milliwatt signal is about 223 millivolts across 50 ohms: pretty small. A ten milliwatt signal would be just over three times greater in voltage. Calibrate your diode probe with audio, remembering to use capacitors whose time constants are not too small.

Base your attenuator design on Daughters' and Alexander's excellent article on attenuators.[2] This article is one of the best I have seen, because it is written from the viewpoint of a ham who is working with such things.

Set up for the test to minimize rf leakage from the oscillator-attenuator assembly **and** to minimize receiver sensitivity to rf from anywhere but the input terminals. You probably can't get perfection in either department, but a good job in both will produce the same practical results.

summary

That's enough on electronics units. The rest is up to you. When an electronic unit seems hard for you to use, spend some time reading about it. If you look for the way in which the term is built up from simpler terms or from common-sense ideas, it will gradually become more real to you.

Then try to put it to work in some simple breadboard circuit that tests or illustrates your thinking about the unit. Finding work for a new idea is the best way to make it stick.

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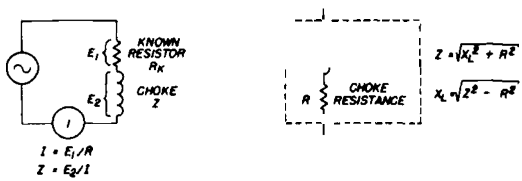
ham radio

unmarked filter chokes

Have you ever picked an old power choke out of your junk box and wondered how you could find out its inductance? 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 vtvm with reasonable ac accuracy and a composition resistor of known resistance (± 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.

fig. 1. Here's the circuit used for determining choke inductance. The principles are explained in the text.



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 vtvm 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 $Z = E/I$.

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

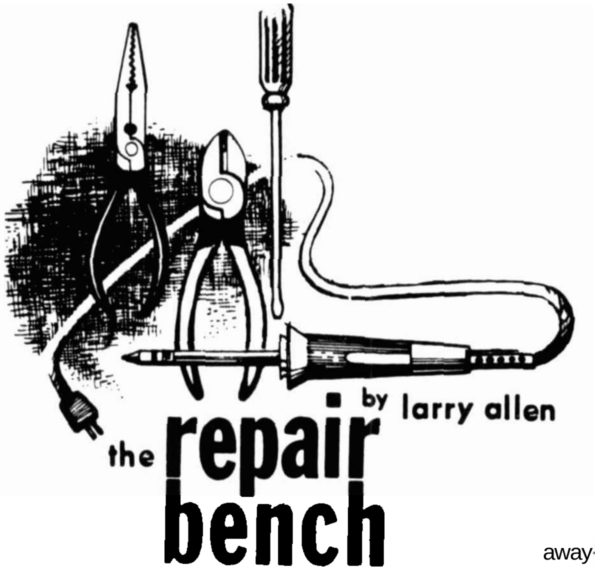
Assume a choke with 320 ohms dc resistance. You measure 4.7 Vac across a 1000-ohm resistor; this gives an ac current of .0047 amperes. You measure 7.5 Vac across the choke. Dividing 7.5 by .0047, you find the choke impedance at 60 hertz is 1600 ohms. From $X_L = \sqrt{Z^2 - R^2}$, where R is the choke 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.

Now that you see how it's done, there shouldn't be a single unknown choke in your junkbox.

table 1. summary of process

1. **Measure the ac voltage across the series resistor.**
2. **Compute I from E/R1 where R1 is the series resistor.**
3. **Measure the ac voltage across the inductor.**
4. **Compute Z from E/I.**
5. **Compute X_L from $\sqrt{Z^2 - R2^2}$ where R2 is the internal resistance of the choke.**
6. **Compute L from $Z/2\pi f$ where $2\pi f = 377$ for 60 hertz.**

Bob Tellefsen, WØMKF



trouble shooting by resistance measurement

In a recent repair bench column* I said you should only work on transmitters with power off. I suggested checking resistances to find the trouble. Since then, readers without repair experience have asked how to go at this kind of trouble shooting. No one has argued the advice, but many don't see the approach.

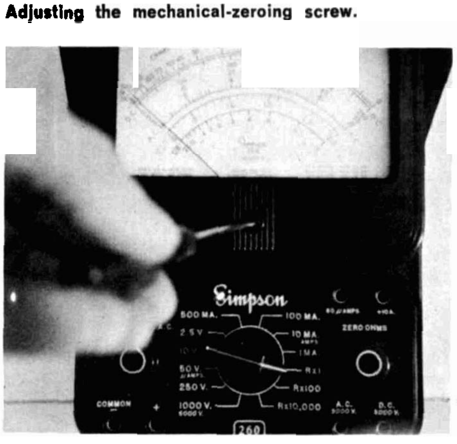
The best excuse for using resistance measurement as a fault-hunting method is safety. You can check out most of a transmitter without any lethal voltages applied. Sure, for a few tests the transmitter or receiver has to be "live." But so many tests can be handled more safely with power off, there's just no excuse to ignore these advantages.

A few doubters may contend that the method isn't complete. Certain parts in high-voltage circuits break down under stress, yet test okay by resistance measurements, they say. That's true. But you soon learn to spot those troubles right away—usually by their smoke. Also, certain shorts and arcs in transmitters occur only after high voltage and drive are applied. The way to track them—safely—is by a process of elimination, part by part. . . but that's a story I'll save for a later column.

Resistance trouble shooting can be done with a simple volt-ohm-milliameter, or vom. That's a major advantage; other techniques demand fancier instruments. A vom is simple to operate, portable, and usually inexpensive. The main criterion is that the ohmmeter readings be accurate.

A vom has an advantage over a vacuum-tube voltmeter, even when the vtvm is more accurate. Since a vtvm must be plugged into the power line, it

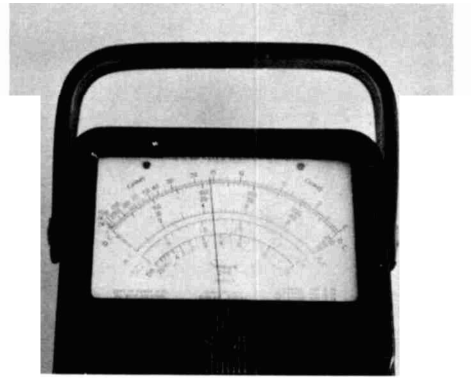
Adjusting the mechanical-zeroing screw.



* *repair bench,* **August 1968, page 52.**

can introduce problems in certain equipment. With the vtvm common lead connected to some non-ground point in a transmitter or receiver, you may get some faulty readings. A vom is usually battery-powered; there's no power-line connection. You can get better readings without worrying about "unseen" ground paths caused by a test connection.

The needle should rest somewhere near midscale for best accuracy during tests.



reading the ohms scales

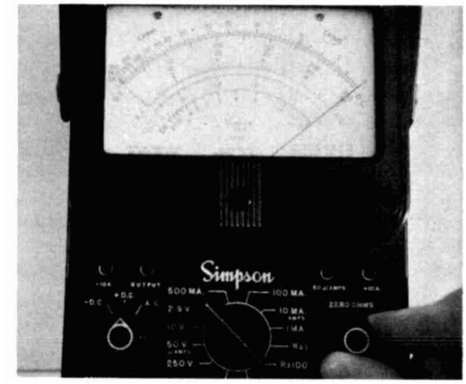
Dependable ohmmeter measurements are useless if you don't read them accurately. You can't even go on to the next step—interpreting their meaning—unless you get the numbers right. Three aids will help you read your ohmmeter correctly.

First of all, calibrate the instrument. Place the vom upright or lying down, in whatever position you'll use it. (Position will affect vom readings.) With the test leads plugged in but not shorted together, make sure the meter pointer rests exactly over the left .edge of the scale. You may have to adjust the mechanical-zero screw that is just below the meter face. Tap the meter glass to jar loose any pointer friction.

Next, clip the two test leads together, making zero ohms between the tips. The pointer should move up to full scale. Zero ohms is at full scale since the ohms-scale numbers go right-to-left. Again, tap the meter glass to be sure the pointer rests normally. That takes care of calibration.

A second aid to accurate resistance measurements concerns the range switch and the multipliers you use with the scale readings. After you connect the test leads across the resistance you're measuring, rotate the range switch to a position that makes the pointer rest somewhere in the middle section of the scale. That's where

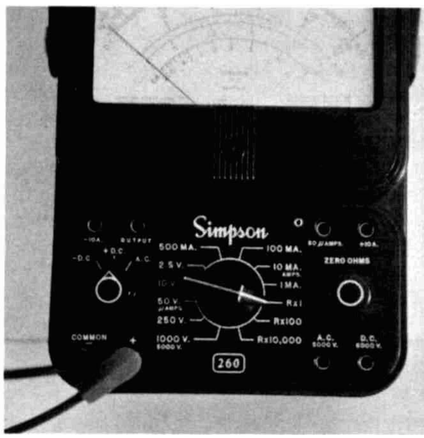
The needle should move up scale when the t a t leads are clipped together.



readings are the most accurate because the numbers there are easy to read. You can see that they are, if you examine the ohms scale in the photos; it's the top row of numbers.

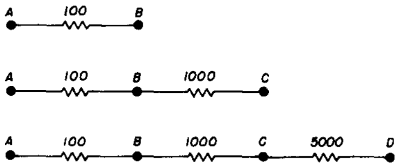
For a third aid to accurate readings,

Pay close attention to the multipliers on the range switch.



make sure you multiply the scale reading properly. Look at the multipliers on the range switch. Pay close attention to which one you set the switch at. Then multiply whatever number the pointer stops at by the multiplier from the switch. Do it carefully. It's awfully easy to drop

fig. 1. When resistances are in series their values add directly even if they're not arranged in a straight line as the ones shown here.



a zero or add one and end up with a wrong reading. Use scratch paper to be sure. Whenever the pointer stops toward the left end of the scale, you'll have zeros in the scale reading as well as in the multiplier; be sure you get those right, too.

simple series paths

What do you do with those accurate resistance readings once you have them? Analyze them, of course! But that brings up the next problem: How?

What you do first is learn to recognize the different kinds of resistive paths. There are plenty of them. Every ham receiver and transmitter has dozens of resistors. Besides them, many, many other components show readings on an ohmmeter. For example, the heaters of tubes; the windings of transformers; the leakage of electrolytic filter capacitors; the forward and backward effects of diodes; and, of course, transistors, which produce some readings that can really fool you. Resistive paths exist all around the chassis.

The trick of resistance trouble shooting is finding paths that aren't where they should be or paths that have too much or too little resistance. To do it, you'll have to know how to spot the paths on diagrams and figure out where and what they should be in the chassis.

The simplest, both to spot and to analyze, are series paths. **Fig. 1** shows a bunch of different kinds. The A-B path shown first can't very well be mistaken; it's 100 ohms, between points A and B.

In the second series path shown in **fig. 1,** the resistance between A and B again is 100 ohms. In series with the A-B path, however, is a path from B to C. Again, the over-all circuit is easy to figure out; the A-C path is the sum of the A-B path plus the B-C path. Resistance in the A-C path totals 1100 ohms.

fig. 2. Here are several series resistance paths whose valuer add directly like those in fig. 1.

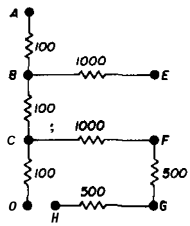
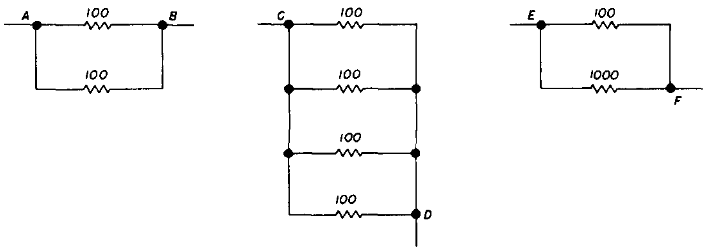


fig. 3. Parallel paths are more difficult to calculate; use the formula in the text to find path resistance.



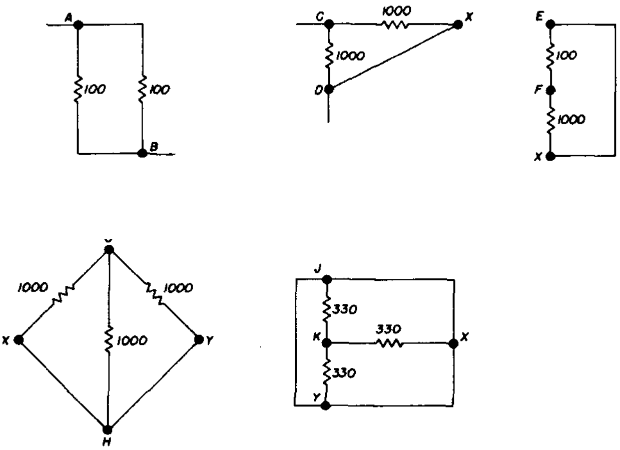
The third series path is A-D. This one, too, is pretty easy, even if it does have three series resistances. Add all three of them, and you have a total A-D resistance of 6100 ohms.

It's important that you recognize another relationship from fig. 1. Connecting the B-C and C-D paths had no effect on the A-B path; in all three cases, an ohmmeter connected to A and B measures 100 ohms. You can add a dozen more paths in all directions, but A-8 remains 100 ohms—as long as all the paths are added in series.

Another thing. You can measure path B-C independently, and path C-D independently. The other series paths don't interfere with measuring either one. Also, you can measure path B-D, if you want to, getting a direct 6000-ohm reading between points B and D.

Take a look at fig. 2. Paths don't have to be drawn in a straight line to be in series. Nor do they have to be connected directly to each other. An ohmmeter connected to A and D measures the A-D path directly; the resistance is 300 ohms. None of the other resistances is of any consequence to that particular path, because they are not in series with it.

fig. 4. Different ways branches may be arranged and connected in parallel; don't let them fool you.



What about path A-H? Only the resistance between A and H have a bearing on the resistance of that path. An ohmmeter connected to A and H measures only the resistances between them. The resistances of path B-E and path C-D are ignored because they're not in series with path A-H. Path A-H measures 2200 ohms. Trace its path—from A to B, to C, to F, to G, to H.

Trace the path from E to G. It goes from E to B, to C, to F, to G. An ohmmeter connected between E and G measures only that path, and indicates 2600 ohms.

The D-H path is 2100 ohms, the B-F path is 1100 ohms, and the F-A path is 1200 ohms. Because these resistances are in series, they're easy to measure and easy to analyze.

Suppose you're trouble shooting a circuit like fig. 2 and measure 2700 ohms between B and G. The schematic shows series resistances totaling only 1600 ohms in the B-G path. At least one resistance has changed value, and you have to figure out which one.

One way is to measure each resistance individually. There may be several in an actual circuit, and that can take a lot of time. A better way is to work your way toward one end or the other. Leave one ohmmeter lead on B and move the other one from G to F. If the reading drops to 1100 ohms, that means the B-F path is okay, so the F-G path must be at fault. Suppose, though, the reading only drops to 2200 ohms. That means F-G is okay; the 500-ohm change tells you so. The trouble is therefore in the B-F path.

Move the test lead from F to C. If the reading drops to 100 ohms, B-C is okay. You can then move both leads to measure path C-F, just to be sure. If it measures 2100 ohms, you've found the trouble; it should measure only 1000 ohms.

Those are the principles. As long as paths are in series, tracking down trouble is a cinch. Plenty of resistive (and continuity) paths in ham equipment are series types. With them, you'll have an easy time. However, the other kind can be a real pain.

resistive paths in parallel

The simplest kind of parallel path is shown in fig. 3. Parallel paths are hard to describe. As far as an ohmmeter is concerned, the path between points A and B is merely path A-B. As you can see from the diagram, however, it isn't that simple. There are two paths. One is from A to B . . . and, yet, so is the other! You'll find several paths like this in electronic equipment, and very often they aren't at all obvious on the schematic diagram.

Resistances in parallel add inversely (another word for upside-down). Put two equal-value resistors in parallel and the resistance of their path is only half the value of either one alone. Take path A-8 in **fig.** 3, for instance. The two 100-ohm ressistors in parallel make a path measuring only 50 ohms; that's what an ohmmeter reads. The C-D path, with four 100-ohm resistances in parallel, measures only 25 ohms with an ohmmeter.

From this you can figure that adding a parallel path always lowers the ohmmeter reading between two points. Parallel path resistance is, in fact, less than the lowest value of resistance in any one branch of the parallel path. The E-F path in **fig. 3** shows a situation where one branch has a much lower resistance than the other. Calculated by the formula for parallel resistances,* path E-F has a resistance of 91 ohms. That's what an ohmmeter measures between points E and F.

Consider what it means, then, if you measure path E-F and find a reading of 100 ohms instead of 91. You can assume the 1000-ohm resistance has been removed, is defective (open), or has changed value to some extremely high resistance. On the other hand, if the reading is 1000 ohms, the 100-ohm resistance is at fault. Get the idea?

odd parallel connections

The trouble with parallel paths in typical ham gear is that they are seldom as obvious as those in fig. 3. For example, they could be arranged in all sorts of odd shapes, like some of those in **fig. 4.** Strange as they may appear, these diagrams are all of parallel resistance paths.

A-B is easy to recognize; it's like the one from **fig.** 3, only turned up on its side. Path C-D looks at first glance like a 1000-ohm path. Follow path C-X, though, and you'll find that X and D are the same point insofar as the circuit is concerned. The C-D path is therefore a parallel one involving both resistances. The true resistance is 500 ohms between C and D; that's what an ohmmeter reads.

Consider the E-F-X path. The two resistors appear in series; but they're not, because the jumper connection makes E and X electrically the same point. So, any current from E flows through both resistances to reach F. The true path is E-F (or F-E if you prefer). An ohmmeter at E and F measures the combined resistance of the two — combined in parallel, not in series. The path from E to F measures 91 ohms.

Path G-H is fairly easy to see, now that you're looking. The path is through three parallel resistances. Points X and Y are mere connections, and are the same (elec-

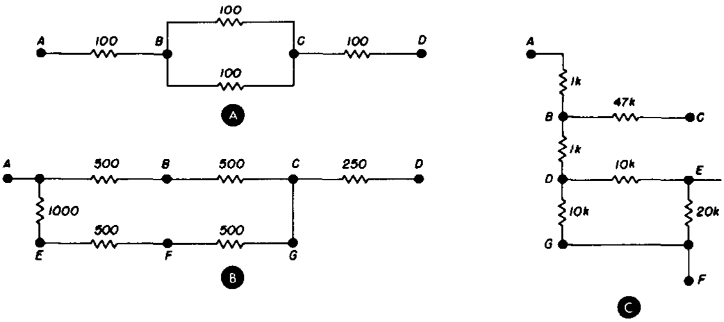


fig. 5. Series-parallel resistance paths are more like what you'll find in real equipment. Still . . . the method of figuring out proper path resistance is the same: calculate parallel paths and add series ones.

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \cdots \frac{1}{R_n}$$

trically) as point H. Combining the three resistances by the parallel-resistance formula, you can figure out the net resistance of path G-H: 333 ohms. An ohmmeter shows it to be around 330 ohms.

Whether it looks that way or not, path J-K is a parallel path through three branches. J, X, and Y are tied together by several jumpers, and can all be considered point J. The net J-K resistance is 110 ohms.

The point of studying odd resistance hookups like those in **fig. 4** is that you won't overlook a parallel connection just because it isn't obvious. In fact, many you'll encounter in actual equipment are even less obvious. Some are through components not drawn with resistor or wire symbols. In every case, keep in mind that a parallel path, whether visible on the schematic or not, always lowers the ohmmeter reading to some value less **than** the lowest value in any branch of the parallel path. If you forget that fact, resistance trouble shooting can be confusing; remember it, and you can use the technique to pin down some very elusive faults.

series-parallel combinations

A combination of paths is what you'll find most in equipment circuits. Some parts form series resistance paths, and some form parallel ones. You'll find them in all kinds of arrangements. A few possibilities are sketched in **fig.** 5, just to give you an idea.

In diagram **A,** paths A-B and C-D are in series with two-branch parallel path B-C. To know what resistance to expect between A and D, you first calculate the resistance of path B-C; that's 50 ohms. With that settled, it's easy to figure the rest: all paths are in series. A-B is 100 ohms, B-C is 50 ohms, and C-D is 100 ohms; the total is 250 ohms.

Fig. 5B is a little different. The paths are complicated. The path from A to C is made of two parallel branches, both of which have several series resistances. To find out what each branch should measure, solve one at a time. Path A-G totals 2000 ohms. Path A-C totals 1000 ohms. Points G and C are jumpered together, so are the same point. Combining the resistances of both parallel branches, you'll find the resistance of path A-C should be 667 ohms; an ohmmeter connected to A and C measures 660 or 670 ohms. The whole path, of course, extends from A to D. Combining path A-C with path C-D is simple addition, since they're in series. Path A-D should measure slightly over 900 ohms.

Fig. 5C shows what you might find in a real circuit. You can figure out what your ohmmeter should measure from any convenient point to any other point. Suppose F (and G) is ground; that's usually a good place to clip one test lead. Path F-E should measure about 10,000 ohms (10k), because the G-D-E path is in parallel with it. Path F-D should measure 7500 (30k in parallel with 10k). The path from F to B includes F-D in series with B-D and should measure 8500 ohms. The F-C path includes F-B as well as B-C and thus totals 55.5k (47k plus 8.5k). From F to A totals 9500 ohms or 9.5k; path B-C doesn't enter into the F-A path at all.

next month

You can see from all these diagrams that trouble shooting by resistance measurement is based entirely on figuring out from the schematic diagram just what a resistance path should be and then measuring it with the ohmmeter. If the resistance isn't what you expect, you isolate the one that's wrong. There, you'll find the trouble.

Principles alone aren't enough to make you feel at home with the resistance-measurement technique. In the next column, I'll show you some real trouble-chasing with this method. I'll show you how to track down parallel-series resistive paths, and how to use resistance tests to pinpoint specific parts troubles. By the time you're through with this subject next month, you'll know how to find faults in any kind of ham equipment with no power applied at all—the safest way you can troubleshoot.

ham radio

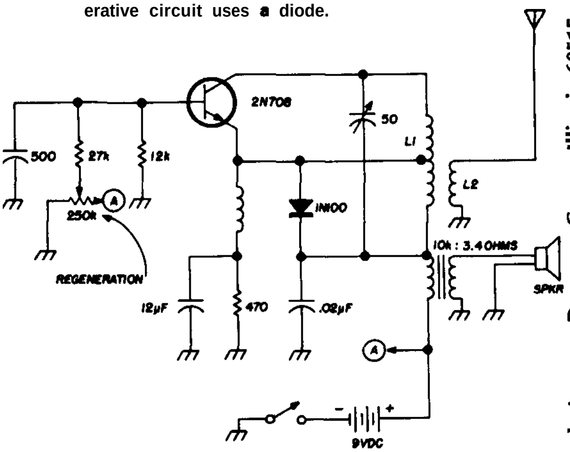
no-radiation no-hangover 28-mhz superregen receiver

Robert M. Brown, K2ZSQ, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

Any amateur familiar with superregenerative receiver design is pretty well aware of their inherent disadvantages. Experienced hams tend to bypass this design altogether because of previous problems. Others simply ignore superregens because they tend to emit spurious radiations on other frequencies.

True, the circuit does have disadvantages—rather non-selective tuning, hangover and radiation—but it still represents a simple approach if you're looking for extremely good sensitivity at minimum expense. A look at the Heath Company's contributions on 28, 50 and 144 MHz in years past can attest to the sensitivity aspect. Now to cure some of the less desirable features.

fig. 1. New type superregenerative circuit uses a diode.



diode circuit eliminates radiation

After years of tinkering with superregenerative receivers, Bell Laboratories recently announced a major breakthrough: the addition of a diode shown in fig. 1. This diode eliminates the problems with hangover and radiation. While the selectivity factor remains, it has been shown that this simple addition promises better performance than any previous designs.

Residual oscillations persisting in the tank circuit of a superregen after active oscillation ends produce the well-known "hangover" effect—resulting in spurious responses and blocking action. Blocking inhibits detection because the receiver is swamped by its own residual signal.

By simply shunting a germanium diode across the tank coil, undesired energy is immediately dissipated after the oscillation burst. This eliminates hangover effects during the remaining sensitivity period of the decay slope. Note that the diode is installed so it doesn't affect the necessary positive feedback.

damping action

The problem of overcoming superregen radiation is also achieved by this diode-insertion technique. The diode damping action lowers the amplitude and significantly shortens the duration of the radiated pulse.

What happens to sensitivity? Since barrier potential—about 0.2 volt with germanium diodes—prevents conduction, tank loading with respect to incoming signals is also eliminated. Therefore, there is no change in sensitivity.

The circuit shown in fig. 1 affords a practical opportunity to experiment with this technique. Any superregenerative receiver can be used, or you can start from scratch. The receiver shown here covers the 28-MHz band and is designed to match 50-ohm lines. The regeneration control should be adjusted just past the point a "hiss" is detected in the speaker output.

ham radio

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Output impedance – **50K** (nominal)
Power **supply** – **115** volts AC

Dimensions – **2-3/4" x 4-11/16" x 6-3/8"**
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propagation predictions for november

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

As I write this column in August, some uncertainty exists about the trend of the present sunspot cycle. If, as predicted by the Zurich Observatory, the present cycle reached its peak this summer, then ionospheric conditions this November will very likely be similar to those experienced during November 1967. On the other hand, there is a 10% probability that the peak may yet occur during 1969, and that the muf's may be significantly higher this fall than last.

However, the smoothed sunspot number may not be the best predictor of quiet ionospheric conditions. The sunspot number must commonly used is the Wolf number, R, given by

$$R = k\ (10G + S)$$

where G is the number of sunspot groups, S is the number of observable individual spots, and k is a correction factor to account for different observers' characteristics. Daily sunspot numbers, obtained from the Swiss Federal Observatory at Zurich, are averaged over each month to obtain a monthly average sunspot number. The daily sunspot numbers vary widely because of the non-uniform distribution of sunspots across the sun's disk and its rotation (at an average 27-day period).

The monthly sunspot numbers are then smoothed by forming a running average over 12 months to form the sunspot numbers used in ionospheric prediction studies. Since the smoothed sunspot number is dependent on future observations, it can only be determined six months after the fact.

Predicted sunspot numbers are formed by extrapolating observed smoothed sunspot numbers and comparisons with past sunspot cycles. The prediction process is not particularly accurate. The National Bureau of Standards underestimated the peak of the last cycle by 50, and the 90% confidence level for last month's predicted smoothed sunspot number was ±29.

What do sunspots have to do with the ionosphere and why should present ionospheric conditions depend on anything that is going to happen in the future? There is no direct causal relation between sunspots and the solar radiations responsible for the ionosphere—both are different manifestations of solar activity. However, during the last two sunspot cycles a case has been built for statistical relations between the two.

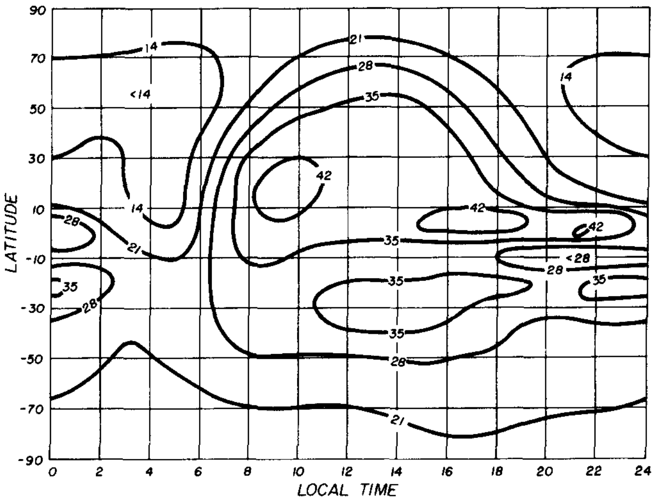
A correlation between solar activity and ionospheric behavior was found by the National Bureau of Standards as early as 1936. In the same year, the first automatic full-range (500 kHz to 16 MHz) ionospheric sounder was used by the Carnegie Institution of Washington.

By the outbreak of World War II, ionospheric observations from only seven locations in the world were available to Allied powers; by the end of the war, 43 stations had been added, and ionospheric data was being published monthly.[1] Thus world-wide synoptic studies of ionospheric behavior have only taken place since about 1943.

By 1946, enough data had been gathered from soundings made at Washington, D.C., Huancayo, Peru, and Watheroo, West Australia to indicate linear statistical relationships between the ionospheric-layer critical frequencies and the smoothed sunspot number.2 In general, the correlation was not as good if monthly average sunspot numbers were used.

By 1961, data collected over cycle 19 indicated a departure from the linear relationships between critical frequencies and smoothed sunspot numbers for sunspot numbers in excess of 150.3 Some differences (less than 1 MHz of critical frequency for a given smoothed sunspot number) are evident between cycles 18 and 19, and between the ascending and descending parts of the cycle.

fig. 1. Maximum usable frequency curves for November 1968 based on 90° W longitude.



Solar cycle 17 peaked at a sunspot number of 113 in 1937; solar cycle 18 peaked at a sunspot number of 150 in 1947 and solar cycle 19 peaked at a sunspot number of 201 in 1957. Solar cycle 20 may have peaked near 101 this year. The sunspot numbers during these past four cycles were greater than during any other cycles since 1870. The point is, past correlation of world-wide ionospheric behavior with solar activity has taken place only during the past two solar cycles, which were of unusually high activity. The same correlation may not necessarily hold for future cycles.

propagation during the month

No matter what the actual smoothed sunspot number is for November, wintertime conditions will soon be here. Twenty meters will probably be closed most evenings before midnight, and fifteen meters will probably be closed within a couple of

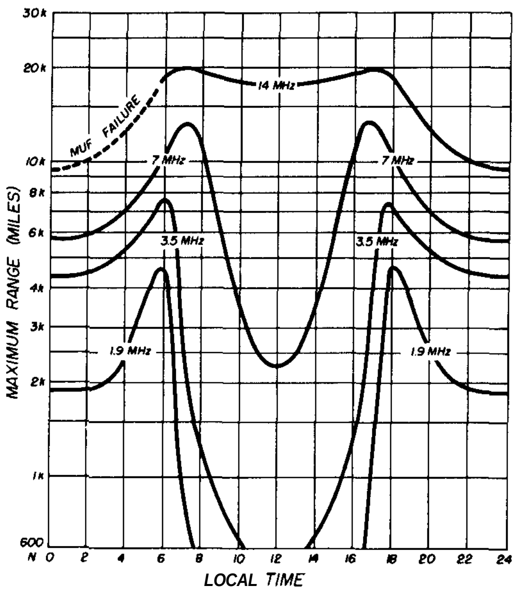


fig. 2. Maximum range to the north due to absorption.

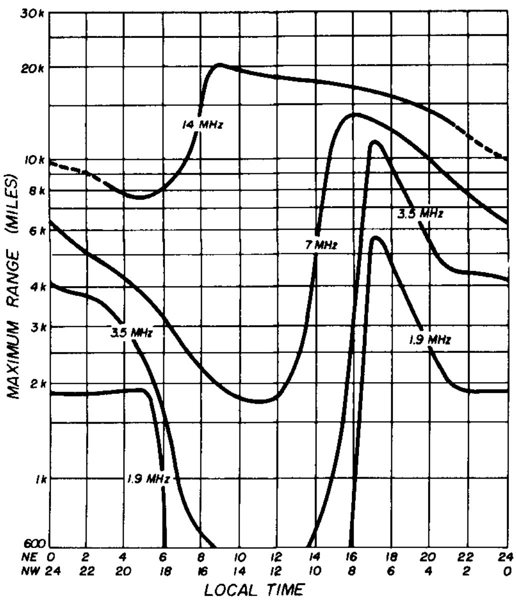


fig. 3. Maximum range to the northeast (top time scale) and to the northwest (lower time scale) due to absorption.

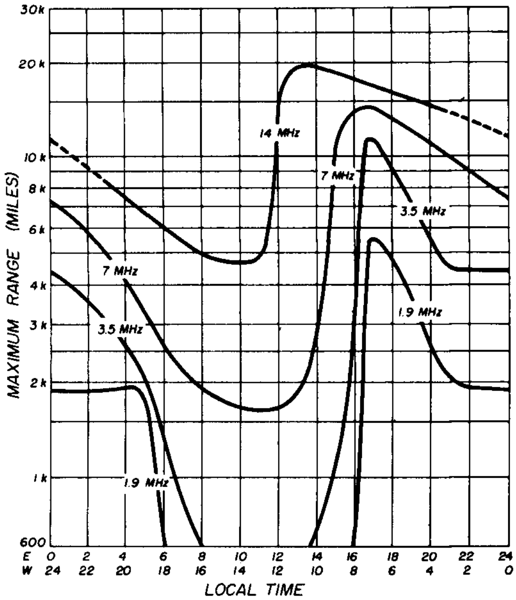


fig. 4. Maximum range to the east (top time scale) and to the west (lower time scale) due to absorption.

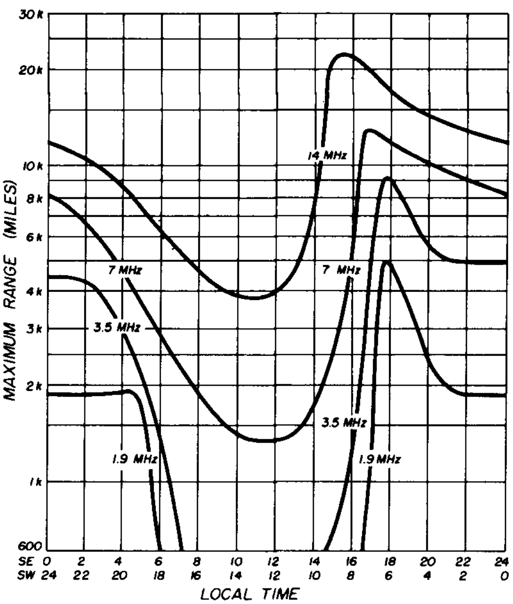


fig. 5. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) due to absorption.

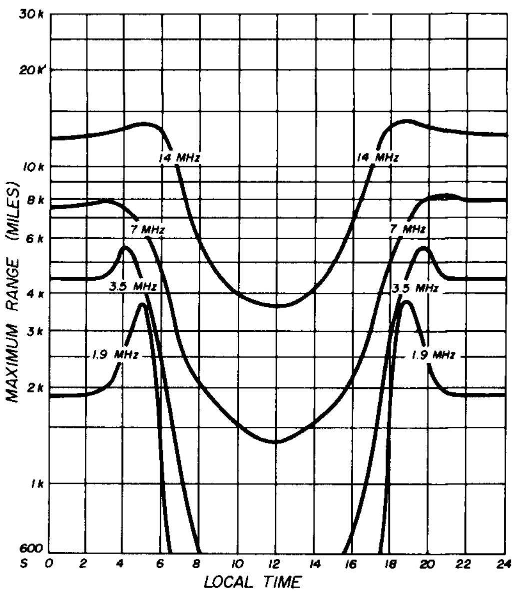


fig. 6. Maximum range to the south due to absorption.

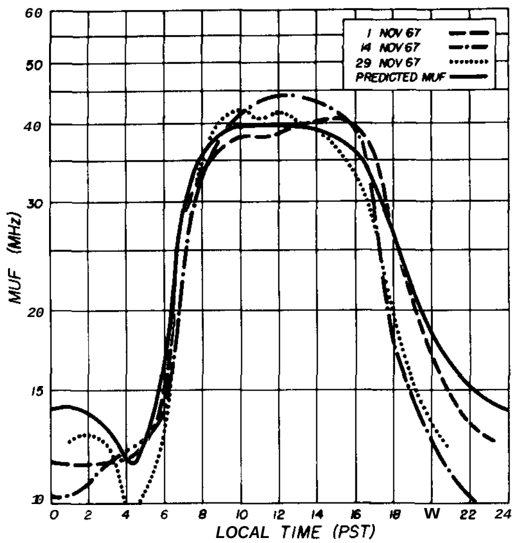


fig. 7. Graph of scaled muf's vs time of day for 1, 14 and 29 November 1967.

hours of sunset. The greatest uncertainties in this month's predictions are in the daytime muf's. Daytime muf's between 20 and 65 degrees North latitude are expected to be significantly higher during November than during October.

During the early afternoon (time of maximum muf's), the muf should increase by about 1/2% per degree of latitude between 30 and 50 degrees North latitude. Nightime muf's during November are expected to be significantly lower than during October. Twenty, fifteen and ten will open later and close earlier as the month progresses. Peak muf's will occur closer to local noon.

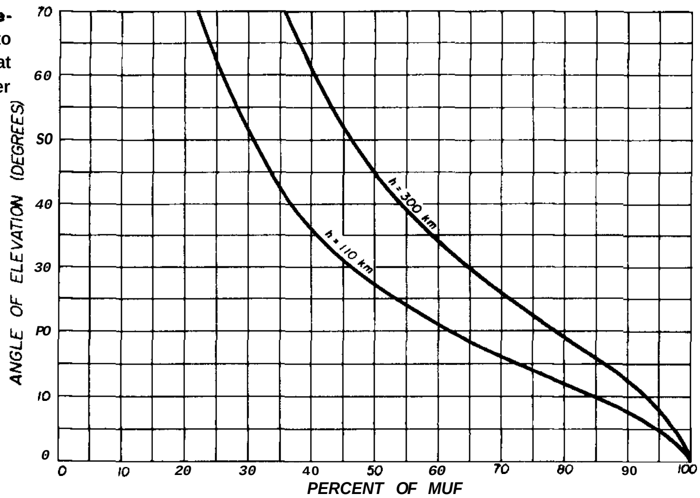
If the six-meter band has been open for single- or multi-hop F2-layer east-west paths during October, six-meters may open during November for two-hop paths between the West Coast and Japan and between the East Coast and Southern Europe peaking at 0000Z and 1600Z respectively, for about the same number of days. The predicted path muf for the two paths is only about 38 MHz. In any case, six-meter openings from the southern half of the U.S. to the south should be possible at least 20% of the days of the month.

Fig. 1 is a time chart of the predicted 4000-km F2-layer muf for November 1968. Maximum range vs time of day charts for 160 to 20 meters are shown in fig. **2** to **6.** The box on page 76 tells you how to use these diagrams.

A graph of scaled muf's vs time of day for November 1, 14 and 29, 1967 along with muf (4000) F2 predictions for the location of the ionospheric sounder (35.5° N) is shown in fig. **7.** The scaling procedure was discussed in the September column.

Fig. 8 is a graph of the maximum angle of elevation that is returned to the earth

fig. 8. Maximum angle of elevation that will be returned to the earth for a frequency that is e given percent of the layer muf.



how to use these propagation charts

1. To find the maximum usable frequency for **F2-layer** propagation in any direction, read the frequency at your control point on the muf time chart. Your control point is 1200 miles away from your station in the direction of propagation, about 18 degrees difference in latitude for a north-south path, or **1½** hours difference in time for an east-west path. To help you find the control point, a control-point circle has been drawn on the muf chart, centered on noon and 38 degrees N latitude. An overlay of this circle may be made on plastic or transparent paper and shifted so that the center of the circle is at your latitude and local time. Greenwich Mean Time may be used **instead** if you make an arrow along the bottom of the overlay at a time (your longitude in degrees W divided by 15) in hours to the right of center.

2. Over any particular path longer than 2500 miles, the path muf is the lower of your and the other station's control-point muf. Curved lines may be drawn on the **overlay** representing the great circle path. The great circle path may be found **from** a globe, a great circle chart or curves in NBS publications.?. [3] However, this muf time chart (for a longitude of **90°** W) will be somewhat in error outside the range of longitudes between **45°** W and **135° W,** and is least representative of actual conditions between **0°** and **180°** E longitudes.

3. **To find the maximum propagation distance** for 160, 80, 40 and 20 meters as limited **by** ionospheric absorption and atmospheric noise, refer to fig. 2, 3, 4, 5 or 6, depending on the direction you wish to work. Note that the time scales are reversed for westward propagation in fig. 3, 4, and 5. These curves are based on unity signal-to-noise ratio in a **6-kHz** bandwidth with 100 watts output power (equivalent to **90%** copy with 100 watts CW or 800 watts **ssb)** on 80 to 20 meters and 1/8 that power on 160 meters. The combined receiver and transmitter antenna gains over an isotropic radiator are −12 dB for 160 and 80 meters, **0** dB for 40 meters and +12 dB for 20 meters. On 10 and 15 meters, the communications range should not be limited by absorption to less than one transit around the earth, although marginal round-the-world propagation will occur at times of minimum range for opposite direction on 14 MHz.

The maximum distance curves were derived from consideration of atmospheric noise levels (from CCIR report 322) and calculated path losses at fixed distances in each direction from **38°** N latitude. Only minor differences in maximum range would be noted due to change in absorption for stations located between **26°** N and **50° N** latitude. Somewhat greater daytime ranges could be expected at more northerly latitudes, and somewhat lesser daytime ranges could be expected at more southerly latitudes. The muf time charts were prepared from basic propagation predictions published monthly by the Institute for Telecommunications Sciences (ITS), Boulder, Colorado **and available** through the U.S. Government Printing Office.

for a frequency that is a given percent of the layer muf. There are two curves, for **true** heights of reflection of 110 km (E-region) and 300 km (F2-region). These curves indicate the importance of good low-angle coverage at frequencies near the muf. They also indicate that if your horizon is blocked, your muf will be lower than the layer muf.

Unfortunately a time chart of muf (2000) E is not available. It would indicate noontime E-layer muf's of about 16 MHz and night-time E-layer muf's of about 4 MHz. For frequencies below the E-layer muf there is a screening angle given by the 110-km curve. Below this angle the F2-layer is not illuminated, and paths longer than 2500 miles will have excessive attenuation. Thus, for long paths and frequencies below the E-layer muf, there may be both an upper and a lower limit to the angles of elevation that are propagated. If the screening angle is too high (depending on the path length) there may be no propagation at any elevation angle with low enough path loss.

references

1. S. Gladden, "A History of Vertical-Incidence Ionosphere Sounding at the National Bureau of Standards," National Bureau of Standards Technical Note *28,* Office of Technical Services, U.S. Department of Commerce, Washington, D.C.
2. NBS Circular 462, "Ionospheric Radio Propagation," U.S. Government Printing Office, Washington, D.C., 1948, p. 60-61, 70.
3. K. Davies, "Ionospheric Radio Propagation," National Bureal of Standards Monograph 80, U.S. Government Printing Office, Washington, D.C., 1965, p. 144.

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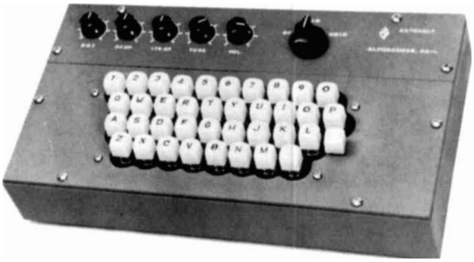

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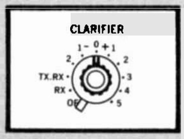

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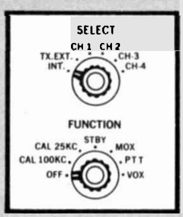

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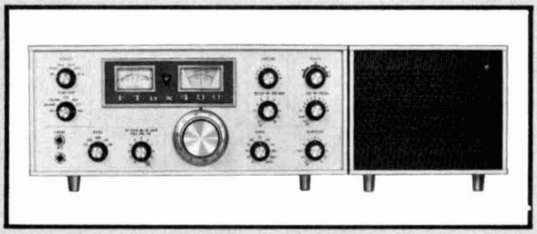

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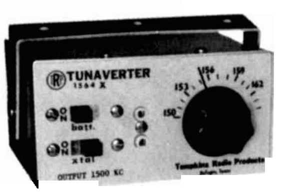

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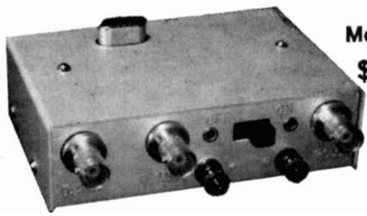

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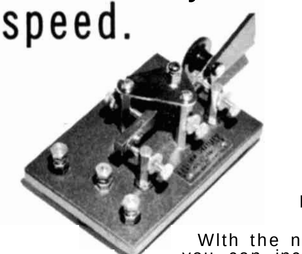

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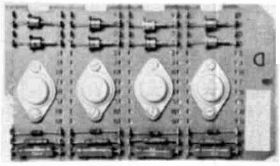

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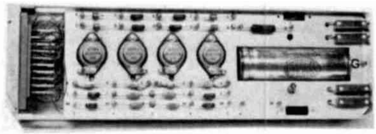

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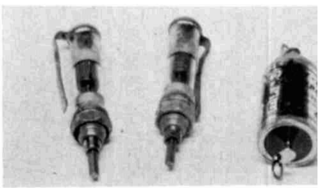

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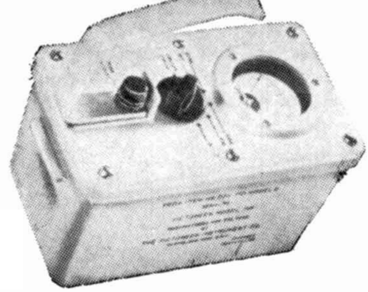

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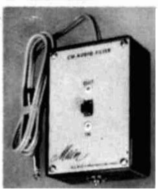

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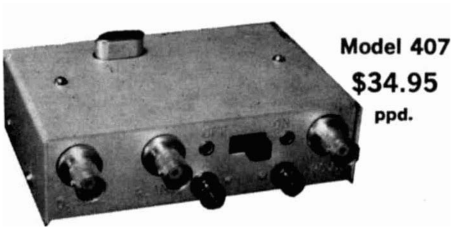

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Some Antenna Notes

Because an antenna is the most important part of your station

To start with, neither QST nor Ham Radio Magazine accepts antenna manufacturer's gain figures in their present advertising. This is because there is not any industry standard for measuring gain, a problem that is presently being worked upon by a committee of the EIA. Some manufacturers in their published literature compare the performance of their product with a somewhat theoretical device called an isotropic source, while others compare against a conventional half-wave dipole. Unless You know what the reference is you will likely not obtain accurate comparisons and, therefore, have false ideas of the product's performance.

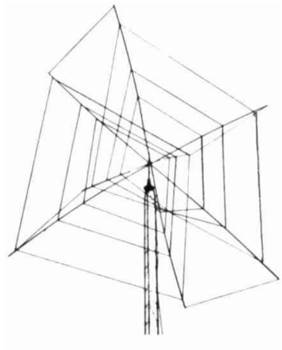
All other factors being equal, low radiation angle and maximum capture area are the most important antenna considerations. Long wire antennas, drooping verticals, and verticals all have their place — for certain kinds of work. But it should be realized that a vertical will radiate a donut shaped pattern in all horizontal directions while a horizontal antenna will radiate a half-donut sideways and upward. Upward doesn't count. A vertical is therefore very advantageous. but must usually have a good ground or counterpoise system.

When an antenna is excited at its resonant or fundamental frequency, it can radiate energy very much better than a random length of wire turned through a matchbox.

The useful frequency band of a particular type of an antenna is inversely proportional to its design frequency and is directly related to the design of the antenna and its installation. Typically, a long wire antenna of No. 12 wire cut at 3900 kHz will range plus or minus 12½ kHz for good results. At 40 meters this will double, at 20 meters the width will be a good hundred kHz while at 10 meters the response will typically be 300 or more kHz. The greater the cross section of the antenna the broader the response.

Multi-band antennas are not normally optimized for each band of operation. They will exhibit excellent results on one band, good on a second and fair response on a third. Typical tri-band beams or quads have an ideal electrical spacing on one band only and offer compromise spacing and feed matching on the others.

A theoretical antenna at resonance is purely resistive and, therefore, can offer the best match at this frequency. Since the maximum radiation occurs at the resonant frequency of the antenna, amateurs



need to know the actual response curves of their antennas. Several devices exist by which antennas may be evaluated. Rf impedance bridges, antenna scopes, antenna bridges or vswr meters may be used. But it must be realized that the typical vswr meter is inserted at the transmitter instead of where the most meaningful results can be obtained, right at the antenna feed. When inserted in the feed line at the transmitter the vswr meter reflects the terminating impedance of the coax, not the performance of the antenna although a resonant and properly matched antenna will exhibit a low vswr.

Transmission line loss invariably causes an artificially lower vswr than in reality exists at the feed point of the antenna. The loss in the feed line lowers the reflected power measured at the transmitter, but in no way improves the antenna itself. The reflected power as shown by the vswr meter is simply an indication of your complete antenna system's efficiency at the frequency you are working on.

It will help, therefore, to recognize that the practical standard of most manufacturers is a maximum 2 to 1 vswr. For maximum performance and reliability you should try to operate within this limit. One practical result of these notes is for the reader to prepare for each of his operating bands a graph showing vswr versus frequency and then to learn to operate within the limits of his antennas.

A possible second approach is to buy a Regina f321A three band quad for operation on 10, 15 or 20 meters. This exclusively distributed product is the only commercial device of its kind whose vswr is guaranteed not to exceed 1½ to 1 over each band. The gain of the Regina f321A is optimum and the new all aluminum version is mechanically superb. It is still priced at only $90.00 f.o.b. Harvard. Previous ads describe our product more fully or you may write for this information today.

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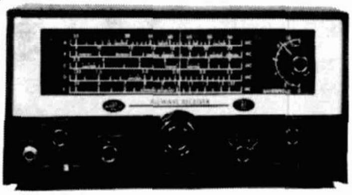

MODEL R-5 ALLWAVE RECEIVER

An exceptionally fine receiver for the short wave listener and beginning amateur operator. Fully transistorized—solid state. Covers 54 through 54.0 Mc in five continuous bands. Includes standard broadcast band, all foreign broadcast bands, all amateur bands from 160 through 6 meters, all 71 Mc CB Channels, all 2-way radio frequencies from 30 to 50 Mc including many police and fire departments. Controls include Variable Beat Frequency Oscillator, Noise Limiter, Bandspread. Compare with tube-type units costing as much!

Kit **$64.95**
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TX-62 TRANSMITTER

In response to the demand for an inexpensive, compact VHF transmitter, Ameco has brought out its 2 and 6 meter transmitter. There is no other transmitter like it on the market. 75 watts input phone and CW. Built-in solid state power supply. Broadbanded circuits make tuning easy. Uses inexpensive crystals or external VFO.
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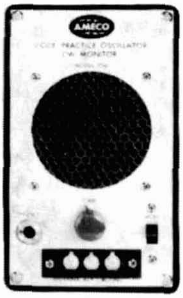

MODEL OM

A combination code practice oscillator and CW Monitor. (No connection to transmitter required.) Transistorized. Has built-in speaker, tone control and headphone jack.

Model OMK, Kit (less batteries) .... $ 9.95
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Similar unit, but without CW RF Monitor Feature.

Model OCPK, Kit (less batteries) .... $ 7.95
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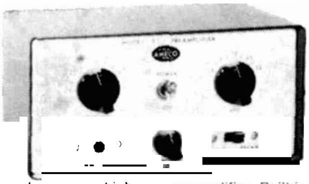

MODEL PT

Low noise, high gain preamplifier. Built-in power supply. Inter-connecting cables included. Station power control center—one switch controls everything. Automatic changeover, transmit to receive. Feeds and mutes second receiver.
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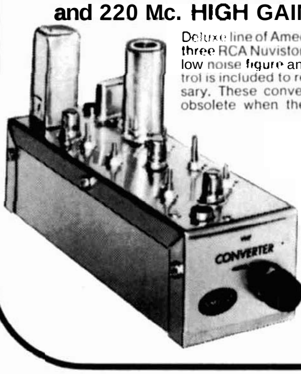

Deluxe line of Ameco VHF Converters uses three RCA Nuvistors to obtain an extremely low noise figure and high gain. A gain control is included to reduce the gain if necessary. These converters do NOT become obsolete when the receiver is changed.

Model **CN-50W, CN-144W** or **CN220W wired, (specify IF)** . 149.95

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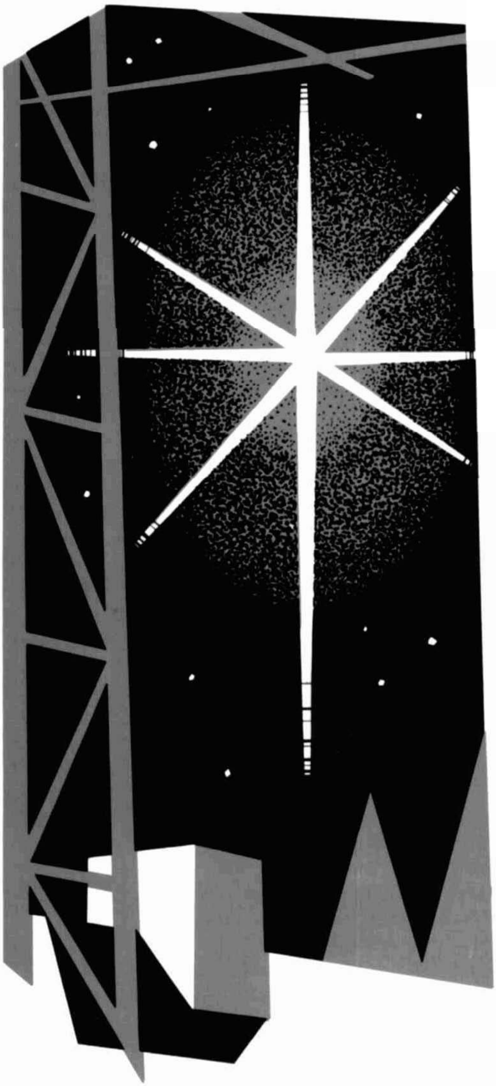

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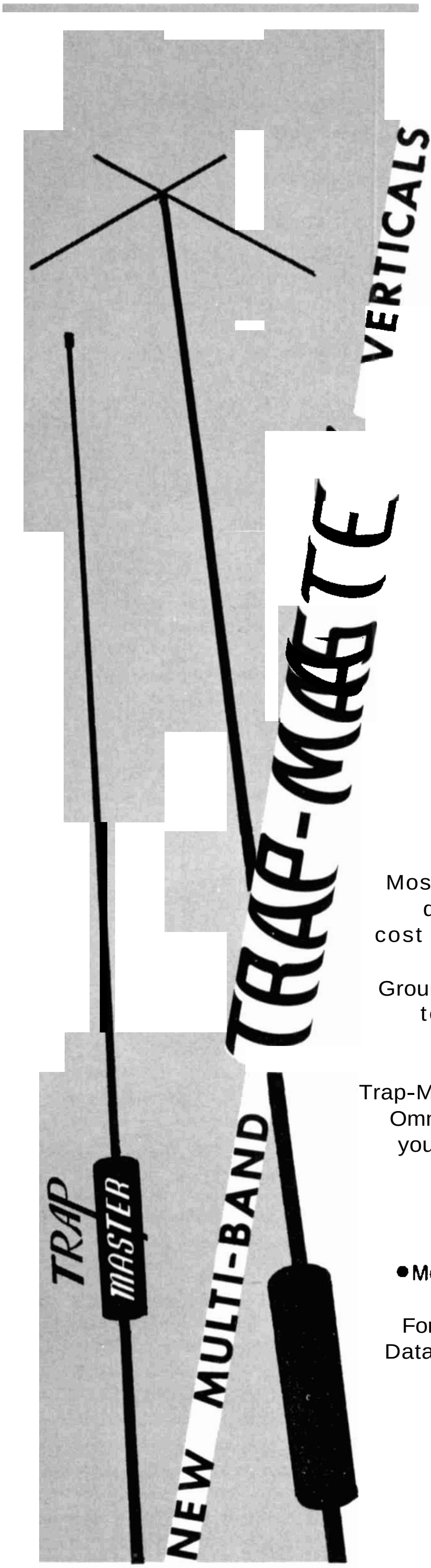
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SIZE. 5⅝" high. 10⅝" wide 16⅝" deep (plus feet and knobs). WEIGHT: 15⅝ lbs.
FREQUENCY COVERAGE: 49.4 to 54.0 MHz (crystals supplied for 49.9 to 51.1 only).
VFO DIAL CALIBRATION: 1 kHz divisions; dial accuracy is within ±1 kHz.
CALIBRATOR: 100 kHz calibrator built in.
FREQUENCY STABILITY: Less than 100 Hz over. all drift per hour after 15 minutes warm-up; less than 100 Hz for 10% supply voltage change.
SPLIT FREQUENCY OPERATION: Xmt and Rcv frequencies may be separated by up to 600 kHz by use of the RV-6 or FF-1 accessories.
MODES: SSB, AM, and CW.
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TUBES AND SEMICONDUCTORS: 19 tubes, 7 bipolar and 3 field effect transistors, 12 diodes.

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SELECTIVITY: 6 dB bandwidth 2.4 kHz with USB filter provided. Accessory filters available for LSB, AM (6 kHz) and CW (.3 kHz).
AUDIO RESPONSE: 400 to 2800 Hz at 6 dB.
INPUT: 50 ohms unbalanced.
OUTPUT: 4 ohms to speaker or headphones.
AUDIO OUTPUT POWER: 2 watts at 10% HD.
AVC: Output variation less than 3 dB for 60 dB input change. Fast attack. Release time selectable.
MANUAL GAIN CONTROLS: RF ain control sets threshold for AVC, AF gain confrol.
DETECTORS: Switch on front panel. Product detector for SSB and CW Envelope detector for AM.
NOISE BLANKER: On-off switch for accessory noise blanker on front panel.
INPUT: 13.9 to 14.5 MHz receiving input/output jack for converters and/or outboard IF receivers.

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MODES: SSB (USB provided. LSB with accessory filter). AM (controlled carrier system), CW (semi-break in. Sidetone).
AMPLIFIED AGC: Prevents flat-topping.
CARRIER INSERTION AND SHIFT: Automatic on AM and CW, shifted carrier CW system.
VOX AND PTT: VOX and Anti-VOX built-in.
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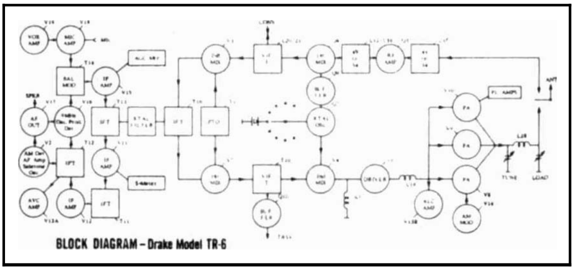
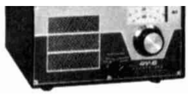
- Extra input and output jacks for converters and/or outboard receivers. Permits monitoring of more than one frequency simultaneously.



BLOCK DIAGRAM – Drake Model TR-6

TR-6 ACCESSORIES



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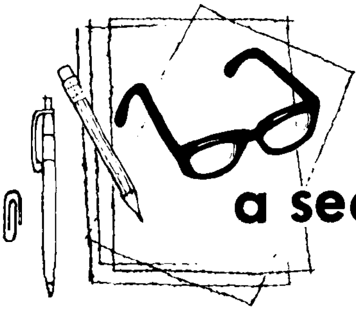

a second look

by jim fisk

Although integrated circuits have only recently started making a real dent in the consumer market, they are actually celebrating their tenth birthday this year. It was during the summer and fall of 1958 that Jack S. Kilby of Texas Instruments built the first integrated circuit. Other firms had been working on ways to miniaturize electronic circuits, but most of these approaches used miniature components of one kind or another. Kilby was the first to use semiconductor material for both the active (transistors) and passive elements (resistors and capacitors) to build a complete circuit on a single piece of germanium.

His first circuits, a phase-shift oscillator and multivibrator, demonstrated the feasibility of this approach. Since germanium was well established and silicon was not, Kilby used germanium. On top of the germanium substrate were the contacts of the diffused transistors, junction capacitors and resistors. A gold-plated metal frame protruded from the lower surface of the substrate and thermally-bonded gold wires were used for connections between those elements not linked by the wafer itself.

The first circuits were large and irregular—a lot different from the precision units that are available today. The photo masks and resists necessary to ic manufacturing were yet to be developed, so the patterns were hand painted on the semiconductor chip with black wax. Needless to say, they were rather crude looking.

About the same time that ic's were first being built, Fairchild Semiconductor developed the planar process—an innovation that is generally conceded to be the foremost semiconductor discovery of the decade. The planar process made semiconductors more reliable and cheaper to produce, as well as accelerating ic progress and acceptance.

Since these early discoveries, the number of circuits per unit area has increased and prices have gone down. In 1962, a typical ic flip-flop chip was 0.1-inch square; a similar circuit today is ten times smaller. You can buy a dual flip-flop for a couple of dollars or a complete decade counter for about seven. Linear ic's are available for all types of applications at reasonable prices: audio amplifiers, i-f amplifiers, audio output stages (with powers up to 5 watts) and voltage regulators, just to name a few. If you haven't tried them yet, you should give them a try—most manufacturers will be glad to furnish data sheets and most large distributors carry inexpensive integrated circuits in stock.

With the low costs involved, it is possible for the amateur to attempt construction projects that he would never even consider in the past. This is also true of commercial amateur equipment. Look for some interesting uses of integrated circuits in new amateur equipment that will reach the market early next year. In the meantime, why not get your feet wet with some of the simple ic projects that have appeared in the various magazines? I wouldn't recommend the digital frequency counter featured in this issue, but it will give you an idea of the complex equipment that you can economically build in your basement. It wasn't too many years ago that only large laboratories with big budgets could afford such an instrument.

Jim Fisk, W1DTY
Editor

EIMAC

3-400Z's or 3-500Z's interchangeable in new Swan Mark II amplifier.

Swan's new Mark II linear amplifier uses EIMAC high-mu power triodes to achieve 2 kW PEP SSB input and 1000 watt input on CW, AM, or RTTY.

Unique circuitry permits the choice of a pair of either 3-4002 or 3-500Z zero-bias triodes in this all purpose high-frequency linear amplifier.

Swan chose the heavy duty EIMAC power triodes because these rugged, low-cost power grid tubes are ideal for cathode-driven grounded-grid service, providing a power gain of up to 20 in linear service.

For more information on either the 3-4002 or the 3-500Z, or any other tube in EIMAC's line for advanced transmitters, write Amateur Services Department or contact your nearest EIMAC distributor.

TYPICAL OPERATION

	3-500Z	3-400Z
DC plate voltage (Vdc)	2500	2500
Zero-sig DC plate current (mA)*	130	73
Single-tone DC plate current (mA)	400	400
Single-tone DC grid current (mA)	120	142
Two-tone DC plate current (mA)	280	274
Two-tone DC grid current (mA)	70	82
Peek envelope useful output power (W)	600	560
Reaonsnt load impedance (ohms)	3450	3450
Intermodulation distortion products (dB)	–33	–35
Plate dissipation rating (W)	500	400

'Approximate

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San Carlos, California 94070

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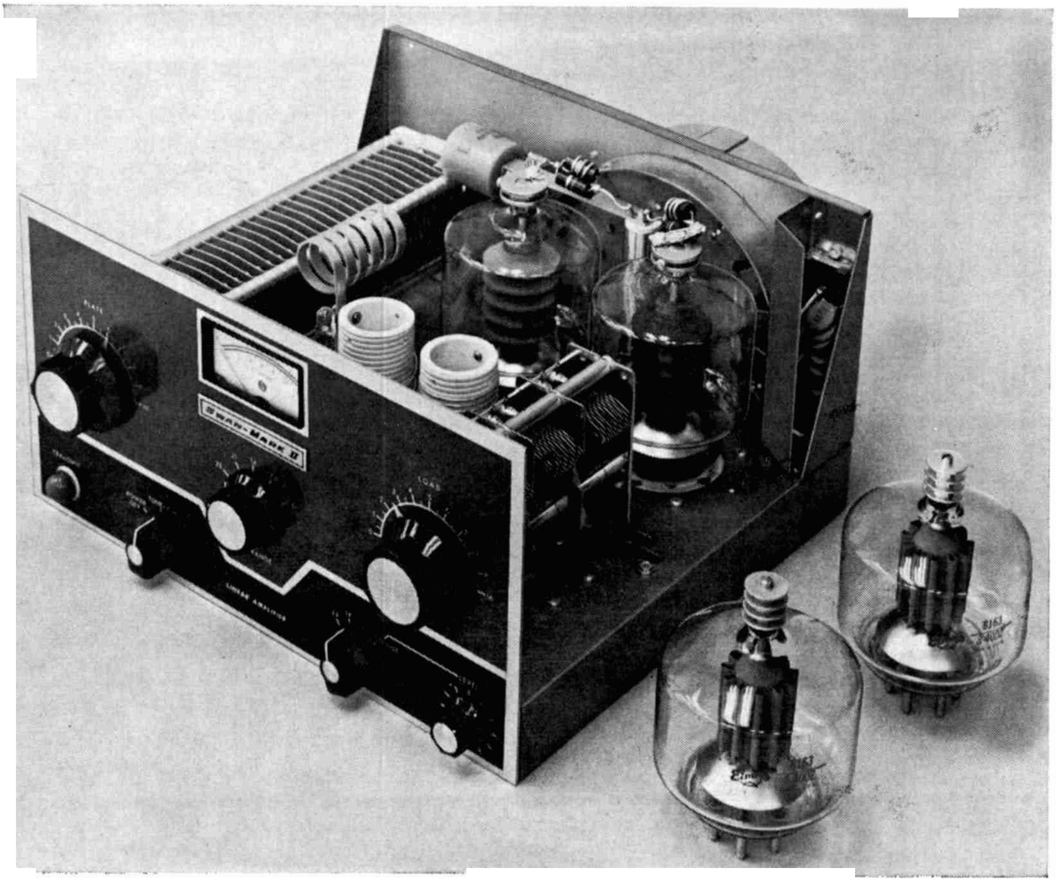


a word of thanks

As our first year of Ham Radio draws to a close, all of use can look back on what has certainly been a most exciting year. A year ago this magazine seemed like little more than a dream. We had no subscribers, no advertisers and a nearly empty room for an office equipped with one small table, two chairs, a telephone and a cardboard box for a filing cabinet.

Next came an ad in Gus Browning's DXer's Magazine introducing Ham Radio and offering $1.00 trial subscriptions. Soon the mail started to come in and we found ourselves with over 100 subscribers. About the same time we approached the many advertisers in the amateur field and enough of them said, "OK," to start our magazine rolling.

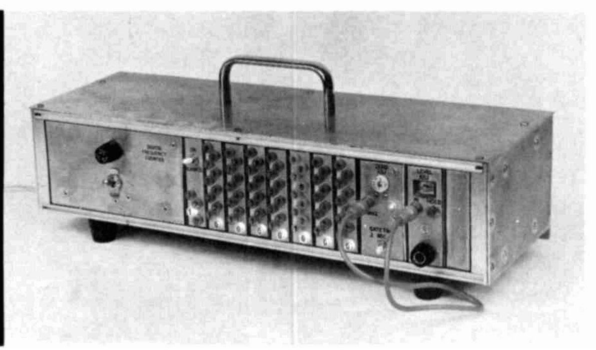
The following months flew by as we worked too many hours a day to line up a printer, to make arrangements with the many authors, artists and others who are so necessary to the successful magazine which you see today. At the same time we had to spread the word about Ham Radio to as much of the amateur community as possible to build up our all-important subscriber list. Hamfests, radio club meetings, visits to radio stores helped. All these activities and more became part of our new routine.

Our pace today is perhaps a bit more orgbnized, but is certainly no slower moving as we work to further improve the magazine, to add new subscribers and to better serve what has now grown to a list of thousands of subscribers.

One person has been very important throughout all of this past year; that is *you*—the reader of Ham Radio. Without you there would be nothing. Your hundreds of letters of encouragement have helped us over the rough spots. Your criticism has also been a big help. You have brought us advertisers and other subscribers. For all of this we want to say thanks.

We wish all of you—readers, advertisers and distributors—a Merry Christmas and a very Happy New Year loaded with lots of cooperative sunspots and many pleasant hours from our hobby.

Skip Tenney, W1NLB
Publisher



digital frequency counter

Application of low-cost IC's in a highly accurate frequency standard featuring automatic operation up to 15 MHz

rt Kelley, K4EEU, 2307 S. Clark Avenue, Tampa, Florida 33609

You won't put WWV out of business with this extremely versatile and accurate instrument, but you certainly won't have to take a hack scat to anybody in a frequency-measuring contest. It's hard to beat the combination of a crystal-controlled time base and digital frequency counter for accuracv and stability. It puts the old heterodyne frequency meter in the high-button shoes class, and if you're mathematically inclined you can use statistical methods to determine the mean and standard deviation of your measurement samples. The accuracy of the final result will be limited only by the number of samples and your patience.

This digital frequency counter, which employs inexpensive RTL (resistor-transistor logic) IC's and surplus transistors, features digital readout and counting capability up to 20 MHz.* It can be used to:

1. Calibrate audio oscillators.

2. Check the 19-kHz suhcarrier of stereo fm stations.

3. Check 3.579545-MHz television color subcarriers.

4. Determine the exact value of compensating capacitors for receiver local oscillators.

5. Monitor vfo drift.

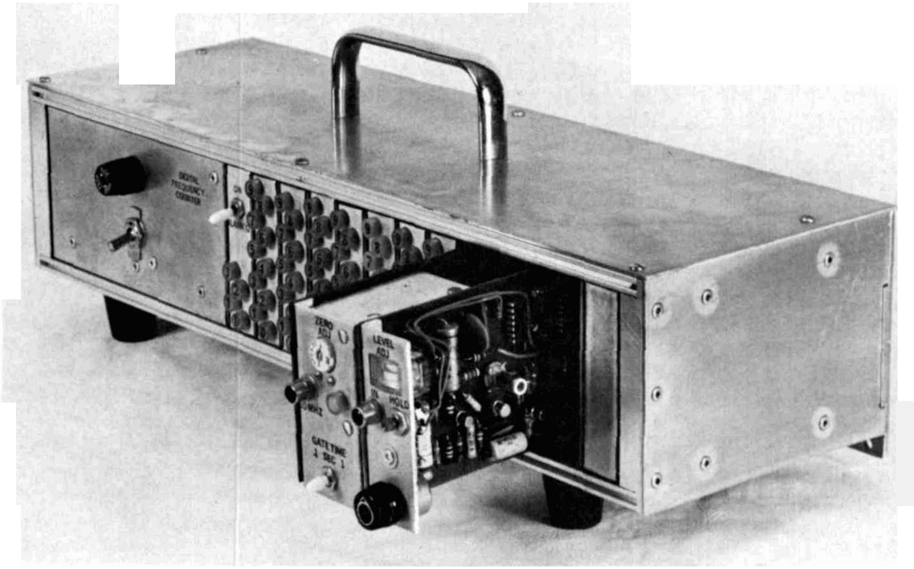
6. Measure toroids for RTTY applications.

The home construction project that resulted in this counter was started several months ago as an educational study of the digital and integrated circuits that were beginning to appear in equipment with which I worked at WFLA-AM-FM-TV. It turned out to be a most interesting project, and the completed counter met and even exceeded my expectations.

in ten million should be obtainable. An idea of the magnitude of this measurement error may be had when you consider that one ten-millionth of the distance from San Francisco to New York is about 16 inches.

The input impedance is in the vicinity of 50,000 ohms, and the sensitivity varies from 0.05 V p-p at 1000 Hz to 1 V p-p at 15 MHz. Operation is automatic once the input level

Frequency counter with the control end time-base modules pulled out.



The accuracy of the unit depends directly on how accurately the counter crystal is zeroed with WWV. To this error is added the plus-minus one-digit error inherent in any electronic counter.[1] You should be able to obtain long-period accuracy of the order of one part in one million. With calibration checks against WWV, accuracies of one part

* The upper frequency limit of the counter depends on the particular batch of IC S you use and the ambient temperature The author built two models of this counter, although the RTL IC's he used arc only rated to 8 MHz. the first worked to *22* MHz, the other to *19 MHz* In *tests here,* I *found* that *counting to* 15 MHz was reliable—above this, it was necessary to adjust the level pot carefully to get meaningful counts Editor.

control is adjusted.

The published information on Motorola MC790P RTL integrated circuits* rated the units to "toggle" to 8 MHz. However, my experiments indicated that it was possible to push this uppcr limit to the vicinity of 25 MHz. The counter decades were built so they could be interchanged, and I found that all decades would toggle above 16 MHz. This upper limit seemed to depend on the module's ambient tcmpcrature. The first completed model of the counter consistently counted to 22 MHz, while the second model,

* Data sheets are available. Write to Motorola Semiconductor Products, Box 955, Phoenix, Arizona 85001.

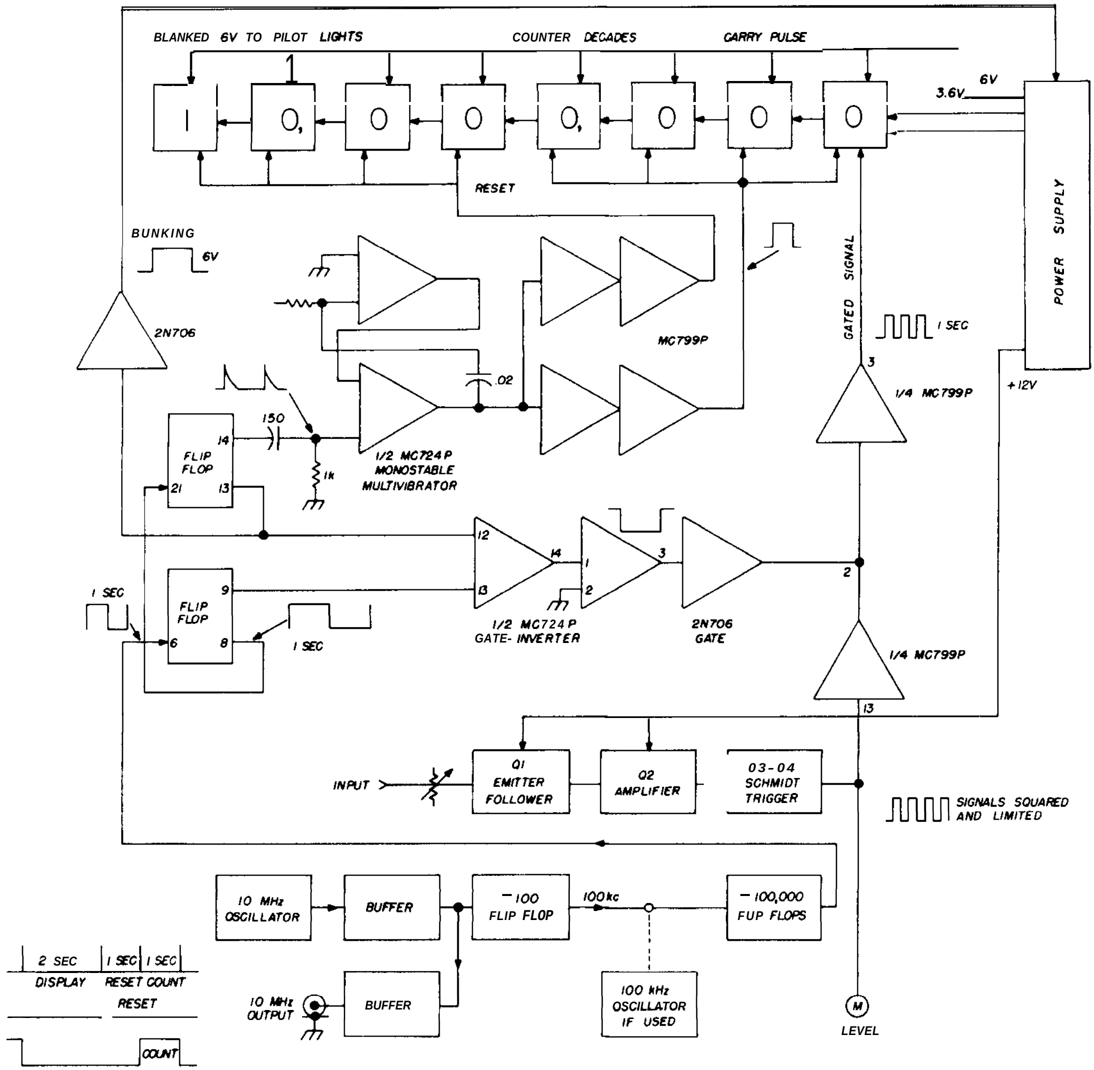


fig. 1. Block diagram of the digital frequency counter.

with MC790P's from a different production run, counted to 19 MHz. In any event, the performance well exceeds Motorola's ratings for these inexpensive integrated circuits.

circuit elements

The complete unit is moderately complex. It uses 48 IC's, 67 transistors, and 75 pilot lamps for indicating the measured frequency. The instrument consists of a time base (clock) module containing a 10-MHz crystal oscillator and associated dividers, a control module, the RTL counter decades, and a power supply. It is compact but not crowded. Dimensions are four inches high by six inches deep by seventeen inches wide. The entire instrument weighs only nine pounds.

Plug-in, "building block" construction is used. The counter can be built in sections, and construction can be simplified by omitting some of the decades. If maximum accuracy is not required, a 100-kHz crystal can be used, thus eliminating four MC790P JK flip-flops. If an input impedance of 1000 ohms is satisfactory, the higher impedance input stage can be omitted. The 12-V sec-

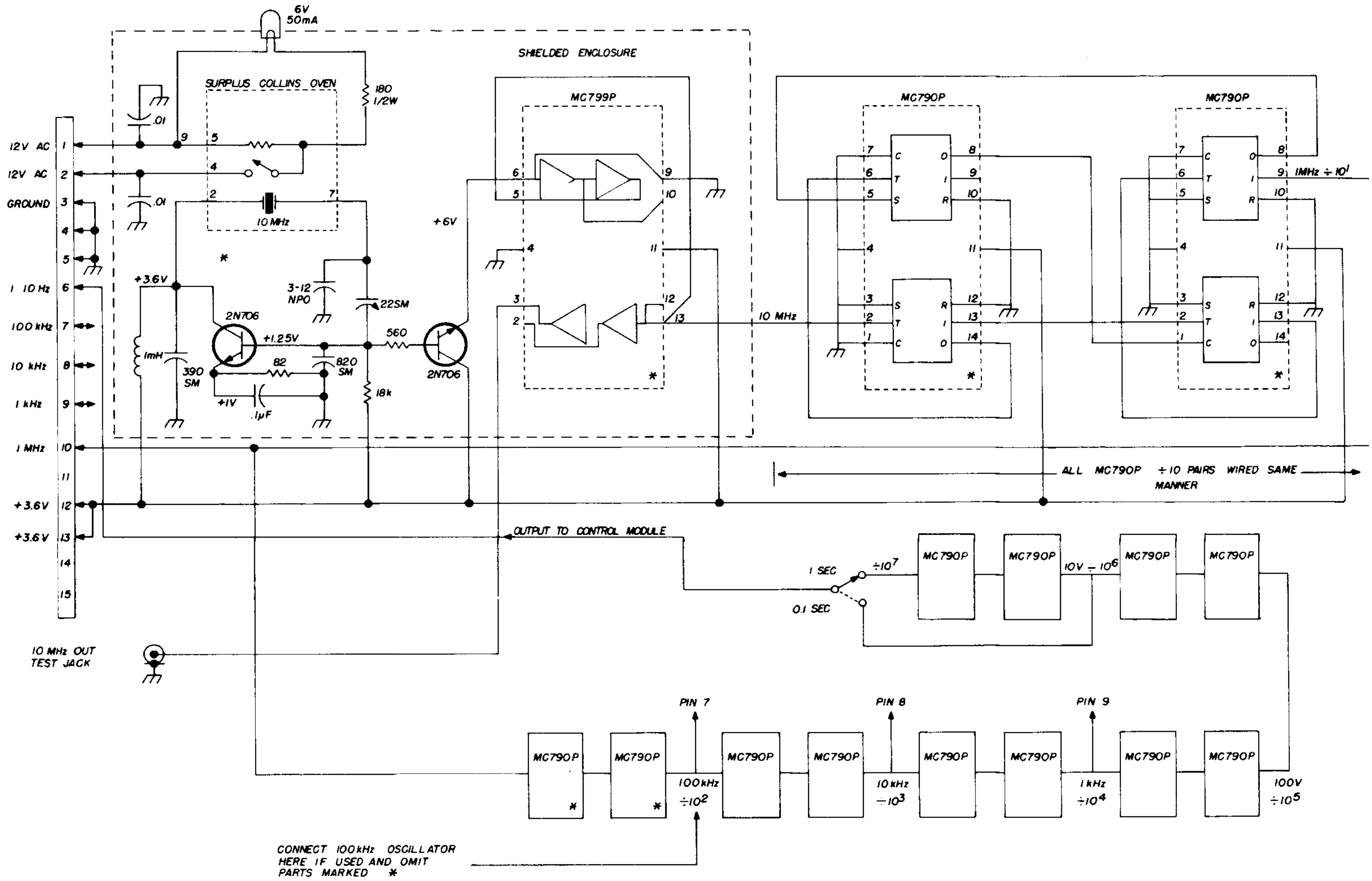


fig. 2. Frequency-standard module provides precise 1 and 0.1 second pulses for gating. The IC pin layouts are shown viewed from the bottom.

tion of the power supply may then be omitted, since it won't be needed in the control module; also you can eliminate the blanking circuit for the indicator lights. Like most things, you get what you pay for; and the fewer the number of decades, the lower the upper count frequency.

The transistors are available in surplus at seven for a dollar. The IC's can be obtained from Allied for about two dollars each.

Three different readout methods are in general use in digital frequency counters; nixies, neon bulbs, and incandescent lamps. Of these, the lamps are the least expensive. The other two require a high-voltage supply, and there is also a strong possibility of cross-talk.

This counter, which features digital instead of binary readout, uses a design published by Lancaster,[2] and those interested are urged to refer to his article for additional information as well as for a source of decade parts. Other parts used in the counter are in the Allied industrial catalog and Meshna surplus listings.

Before attempting to build and test the equipment described in this article, you should have some experience with transistorized circuits. If your experience has been only with tube equipment, it is a good idea to take a course such as the RCA home study series. You will learn everything needed, and the subject matter is not too technical.

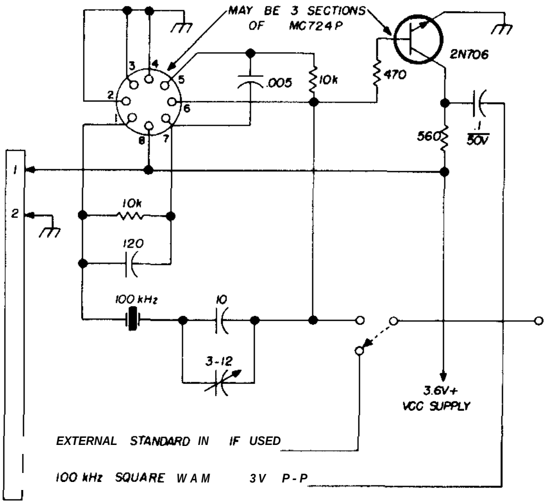
test equipment

Test equipment should include a multimeter such as the Simpson 260. If the meter has a polarity reversing switch, the meter can be used to check transistors and diode junctions. The current ranges can be used to check power consumption of the JK flip-flops. Each MC790P should draw from 40 to 50 mA. Transistors can be zapped very easily when checking them with a multimeter. Make certain there is no excessive voltage on the test leads.

It's convenient, but not absolutely necessary, to have a bench power supply than can be adjusted from zero to about 10 volts, at a rating of about 1 A.

As a minimum requirement a scope is necessary with a frequency response to 4 or 5 MHz. Many circuit malfunctions can be isolated with a low-priced kit scope. Much of the work in the later stages of construction consists of adjusting circuits to extend the high-frequency counting range and tailoring bias circuits to fit the transistors. Furthermore, some transistors don't have the high-frequency response to give clean square waves in the signal processing circuit at 10 MHz. You will therefore need access to a wide-band (15-MHz or higher) oscilloscope.

fig. 3. Time-base module using a 100-kHz crystal. This circuit may be substituted into fig. 2 if so desired.



Two signal sources, an audio oscillator and an rf signal generator, are needed. You should also have a receiver that will tune WWV, of course.

circuit boards

IC's and printed circuit wiring permit compact construction without crowding. Kepro etching materials were used, but a possible alternate would be perforated Vector boards and hand wiring. The etched boards are preferred because of the close pin spacing on the IC's (0.1 inch center-to-center).

Using etched circuit boards is not at all difficult. Circuit layout is drawn full-scale on 10-line-per-inch paper. Hole centers are transferred with a sharp awl punch to the copper side of the board, using light taps. Holes are drilled with a number-60 drill, and burrs are removed by burnishing with crocus cloth or emery paper. Circuit trails are then drawn in to mach the paper layout. These connect the component mounting holes. The finished marking job must be carefully checked against the circuit diagram. After etching, the board is burnished again and is then ready for parts. It's a good idea to make a circuit extender board so modules can be tested under actual operating conditions.

Parts should be soldered using a 22-watt iron, with particular care given to the solder connections. Poor solder connections and hairline voids on printed circuit trails are the greatest cause of trouble. Proceed slowly, making sure there is a connection to the pins, and, when finished, examine the work closely for solder shorts to adjacent pins, etc. Etching instructions are packed with the Kepro circuit boards, and there is further useful information in the current Radio Amateurs Handbook, page 523.

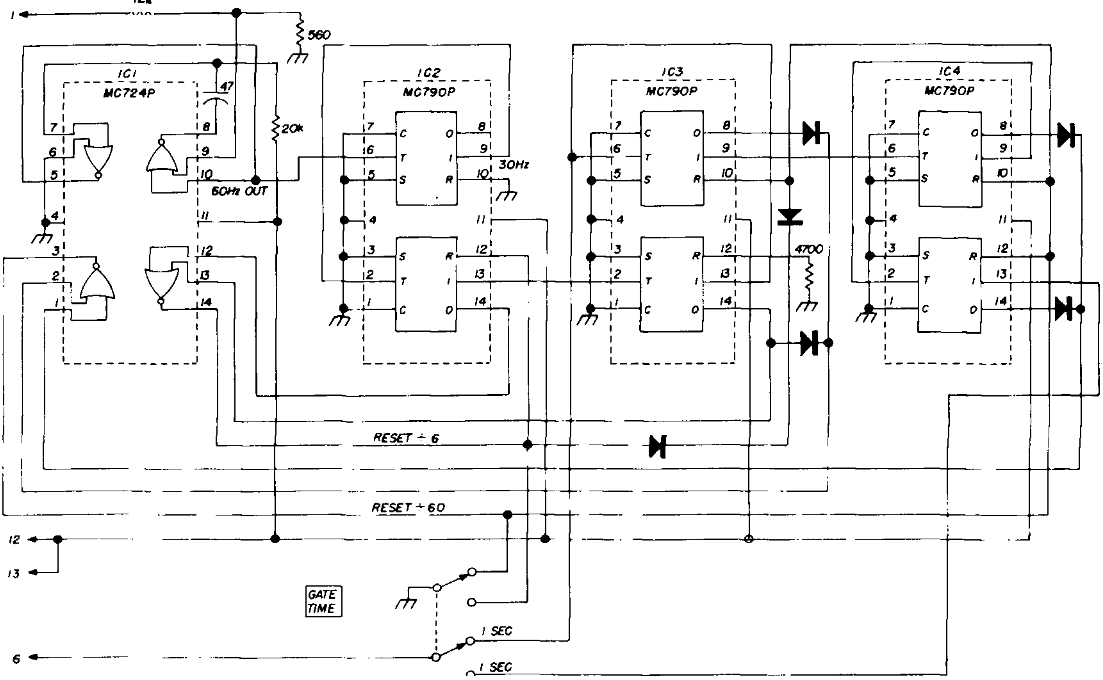
construction

Two aluminum straps support the printed circuit sockets. Though not used in the counter described, Amphenol number 143-012-01 sockets were used in the first model and are recommended. If unetched circuit board is used, Amphenol number 133-012-21 mating connectors may be secured to the board with small machine screws. Pop rivets were used for much of the aluminum fastening. The etched board can be obtained from Meshna or Allied.

For ease in reading, the counter modules should be arranged in line and as close together as practical. The indicator lights are special 50-mA bulbs obtained from the source given in reference 2. These come with small plastic holders and a supply of number decals. If desired, Allied number 60F7413 (Sylvania type 6ES) lamps are a satisfactory substitute. In any case it is necessary to get low current lamps to keep the total power requirements down and stay within the ratings of the lamp driver transistors.

The case is a military surplus version of an Elco Varipack printed circuit enclosure, fitted with extra Elco 63-9016-1204 slides. A local metal shop sheared a supply of 0.064-inch

fig. 4. Circuit that derives a time-base from the 60-Hz ac line. Stability compares favorably with a 100-kHz crystal-controlled time base.



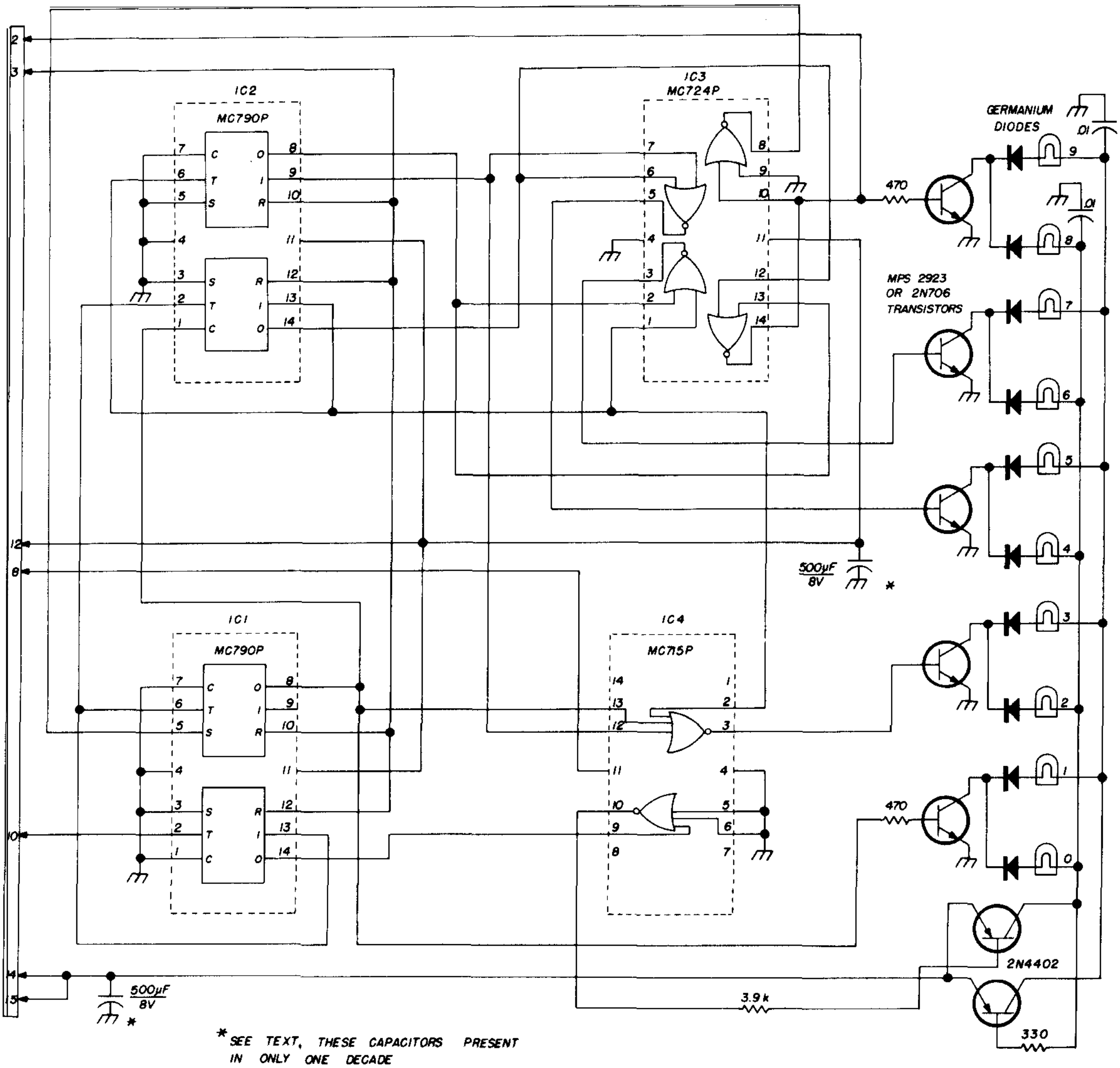


fig. 5. Decade counter module. Printed circuit board layout for this decade is shown in fig. 16.

aluminum to 3-inch widths for later use in constructing module fronts, and three other pieces were obtained for the sides and top. I left the bottom open, and mounted feet from an old TV set to raise the unit above the table to ensure air circulation through the modules.

power supply

The power supply should be constructed first. An aluminum tray approximately 6-inches wide is formed to fit the case. Parts are mounted inside. Two power transformers are necessary; one is a miniature 12-V transformer for the crystal oven, and the other is an ordinary 6.3-V filament transformer. Parts should be arranged for most efficient utilization of space, and the fit is a little snug, especially if the 12-V supply is built **(fig. 13).** The power regulator transistors and power connection sockets are mounted on the back wall. A little silicon grease should be used on the transistor mica washers to aid in heat transfer.

After construction, the 3.6- and 6-V outputs should be loaded to about 1 A each and voltage and ripple checked. The supply must deliver between 3.6 and 4.0 V, and, if necessary, the output voltage can be adjusted by substituting slightly different zener diodes from an assortment available from Allied (stock number 24C9340) or by adding or removing the silicon diode in series with the regulator zener.

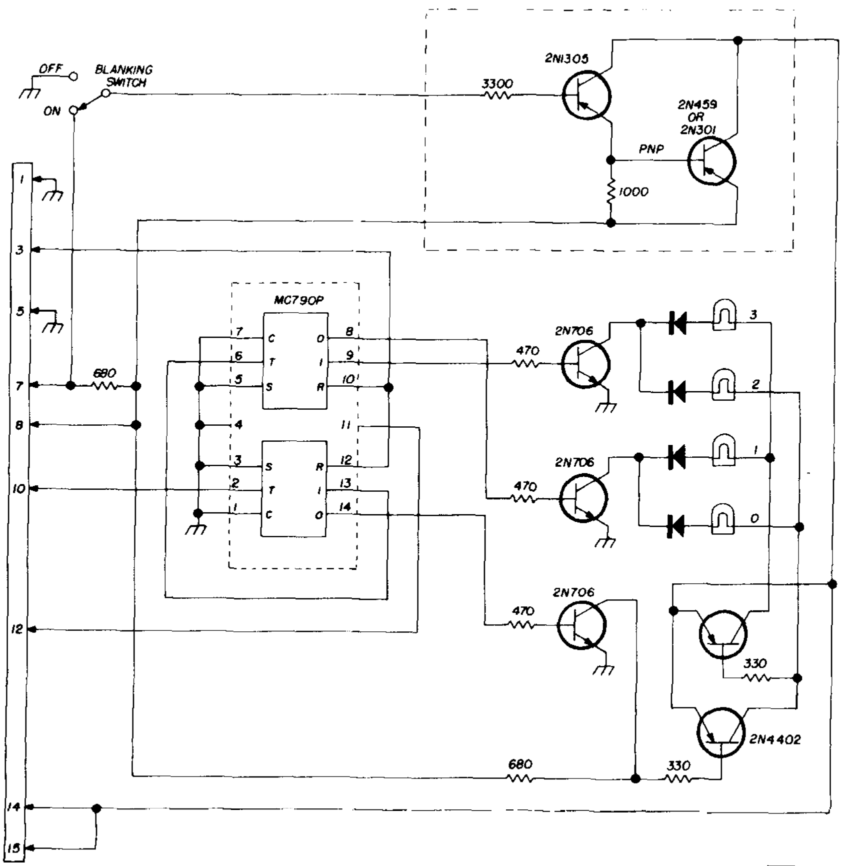
The hum content on the 3.6-V supply should be less than 0.1 volt p-p, but the 6-V supply is not as critical, and the surplus capacitor specified is sufficient. RTL IC's are sensitive to spikes and hum in the power supply, so it may be necessary to add 500-μF/8-V electrolytic capacitors across the 3.6- or 6-V supply in other modules if it is suspected that the flip-flops are toggling on noise. Additional 0.01-μF disc capacitors would be advisable.

The 12-V supply is necessary for the high-impedance input circuits in the control module, so if this is constructed, it should be checked for approximately 12-V dc with low hum content. There is no heavy drain on this supply, and it was built larger than necessary to permit future modifications in other modules.

time-base module

The time-base module (fig. 2) provides the precise signal that controls the time interval gate in the control module. It contains a crystal referenced to WWV and a divider network to divide the crystal frequency down to 10 or 1 Hz, depending on the setting of the gate time switch. The 1-Hz pulse is ideal for the initial checking of counter modules, therefore the time-base module should be constructed next.

fig. 6. Left-hand counter module. Because of the frequency limitation of the counter, this module only counts up to 3. Printed circuit board layout is shown in fig. 17.



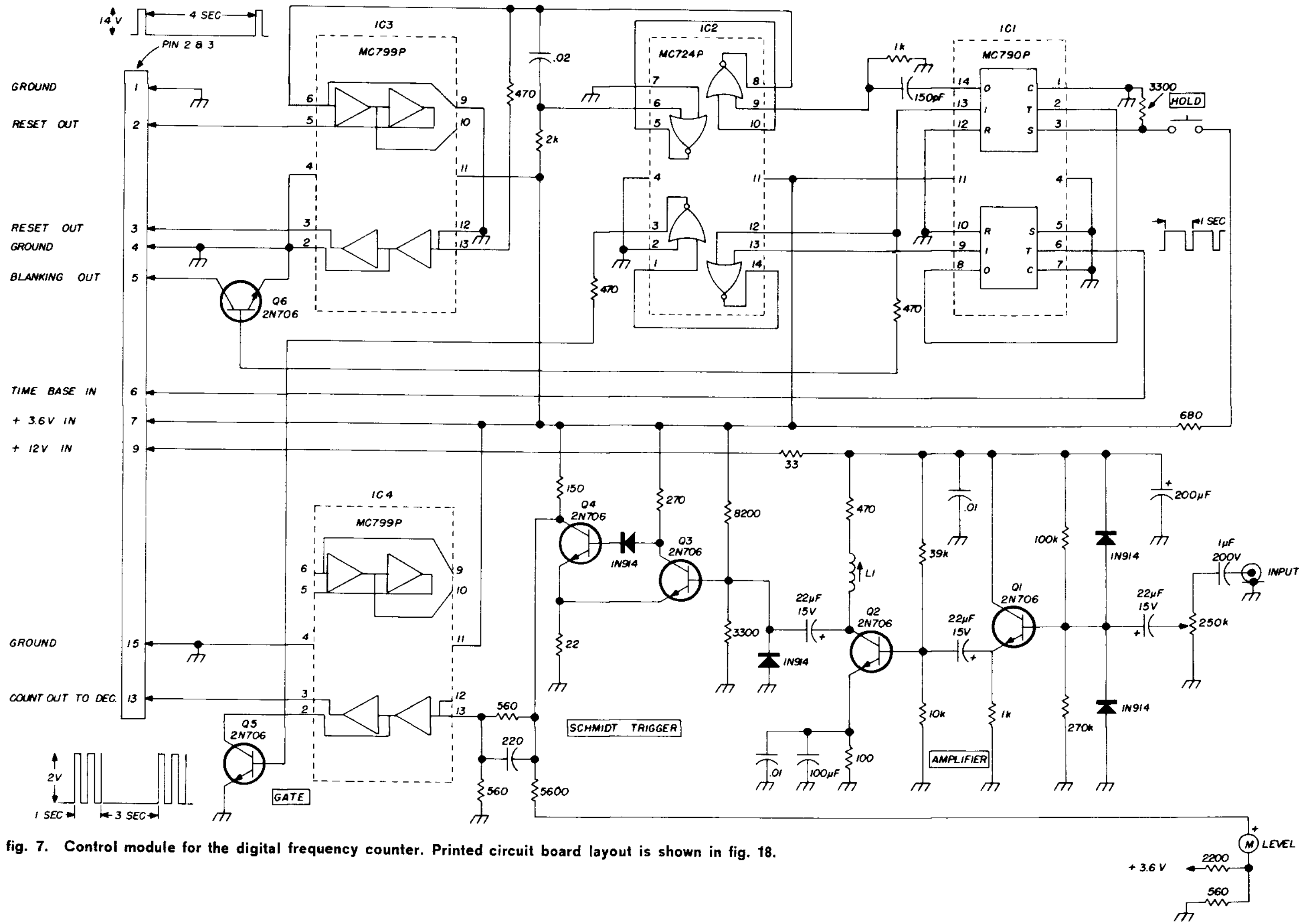


fig. 7. Control module for the digital frequency counter. Printed circuit board layout is shown in fig. 18.

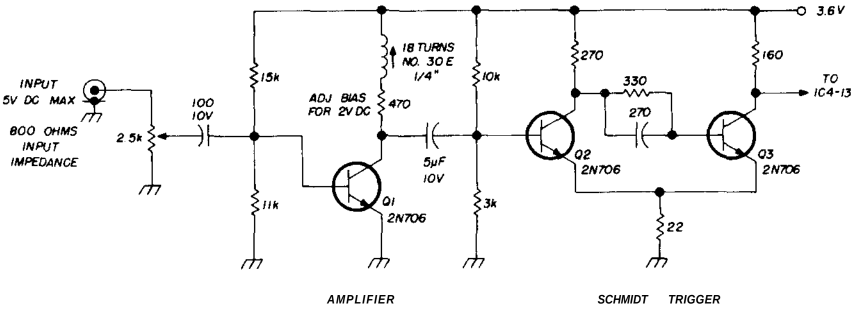


fig. 8. Alternative input circuit if high-impedance input is not required.

The time-base circuits were divided into two modules in the original model. The first contained a 100-kHz crystal and the dividers necessary to derive a 1-Hz pulse. This counter was operated for a time in this configuration, but the 100-kHz reference was later replaced with a 10-MHz reference oven. This oven oscillator and four MC790P dividers were constructed on another module, and the resulting 100-kHz signal was jumpered into the original time base module in place of the low-frequency crystal, which was removed.

The second version has all the components mounted in one module. This requires careful circuit layout and double deck construction. All oscillator components were mounted on a 1¾-inch × 1'12-inch circuit board, which is mounted by a single tapped standoff pillar inside a small shield enclosure. The crystal oven plugs into a socket at the back of the enclosure, and as much unetched copper as possible was left beneath this enclosure to provide shielding, and small circuit trails were etched to provide an exit for 3.6 V dc, 12 V ac, and 10 MHz to the first divider pair. The board also contains a dual driver IC that is used for isolation and to provide a strong signal to the test jack.

After the oscillator assembly is checked out and zeroed with WWV, you can wire the first two MC790P dividers and check them for 1-MHz output. If this checks, wire MC790P JK flip-flops in pairs, checking each pair as you go for division by 10. Use care not to bend back the pins, and be sure the IC is oriented so the number 11 pin goes to 3.6 V before soldering as the IC's are difficult to remove, once mounted. If you have trouble getting the first divider pair working, check the 10-MHz signal at pin 2 with a wide-band scope. The rf signal should rise from near zero to about 1.4 V p-p. If there is a large dc component, the JK will not toggle, so it will be necessary to make some adjustment to the bias on pin 6 of the MC799P. Remedies might include replacing the 2N706 emitter follower, adjusting the base bias on the oscillator, or modifying the value of the internal 1000-ohm resistor in the MC799P by substituting an external resistor. Temporary use of a 10k-ohm pot will quickly determine the optimum resistor size.

In this unit the crystal is a Collins oven from an AN/ARC-27 spectrum oscillator assembly operated with 12 V ac on the oven heater. If a small oven cannot be obtained, it would be worthwhile to mount an ordinary 10-MHz crystal inside the enclosure with the other oscillator components. The oscillator circuit is the most stable and non-critical of several tried, and any active crystal will oscillate, within a frequency range of at least 400 kHz to 10 MHz. Therefore it could be used with a 1-MHz oven with no changes except minor adjustment in the 22-pF series capacitor to permit WWV zeroing.

The divider outputs also can provide a variety of precise signals that can be used to check counter operation or for marker purposes. For example, with a 10-MHz crystal, 5 and 1-MHz; and 500, 100, 10, and l-kHz square waves of approximately 1 V p-p are available for external use, provided that the output is not loaded too heavily. These waves have extremely fast rise times and high harmonic content. An alternative circuit for a 100-kHz crystal is given. This circuit puts out a nearly square wave that is ideal for direct connection to the JK flip-flops.

low-cost time base

After building the crystal-controlled 10-MHz time-base generator, I built another time base (fig. 4) to check the accuracy of the local power-line frequency. From comments in other articles[4,6] it appeared that the true frequency was very close to 60 Hz. I found this to be true.

The 60-Hz time base was operated along with the 10-MHz time base. The I-Hz pulse outputs were selected by a spdt switch; the counter was connected to a stable 10-MHz crystal oscillator for comparisons between the two time bases. With the 60-Hz line-frequency derived time base, readings were always within 1500 Hz of 10 MHz; usual

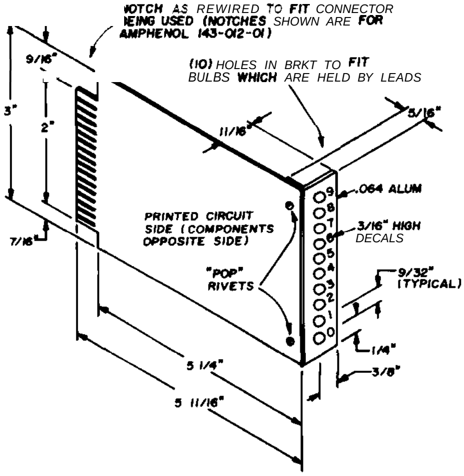
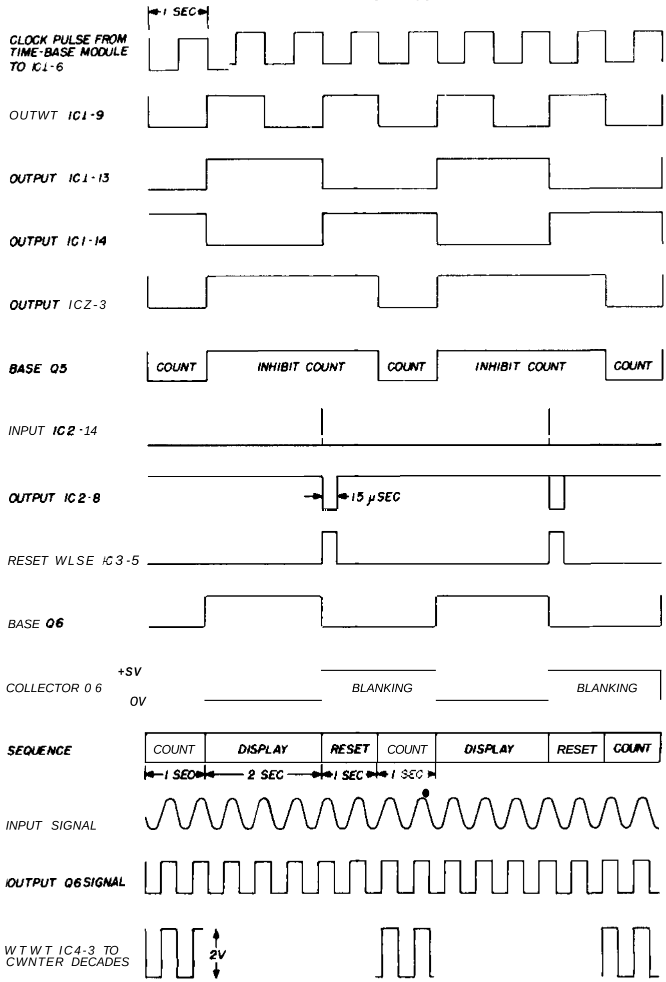


fig. 10. Typical module size and layout designed for Amphrnol 143-012-01 receptacle. Printed circuit boards are Kepro type PI-365.

fig. 9. Control-module waveforms.



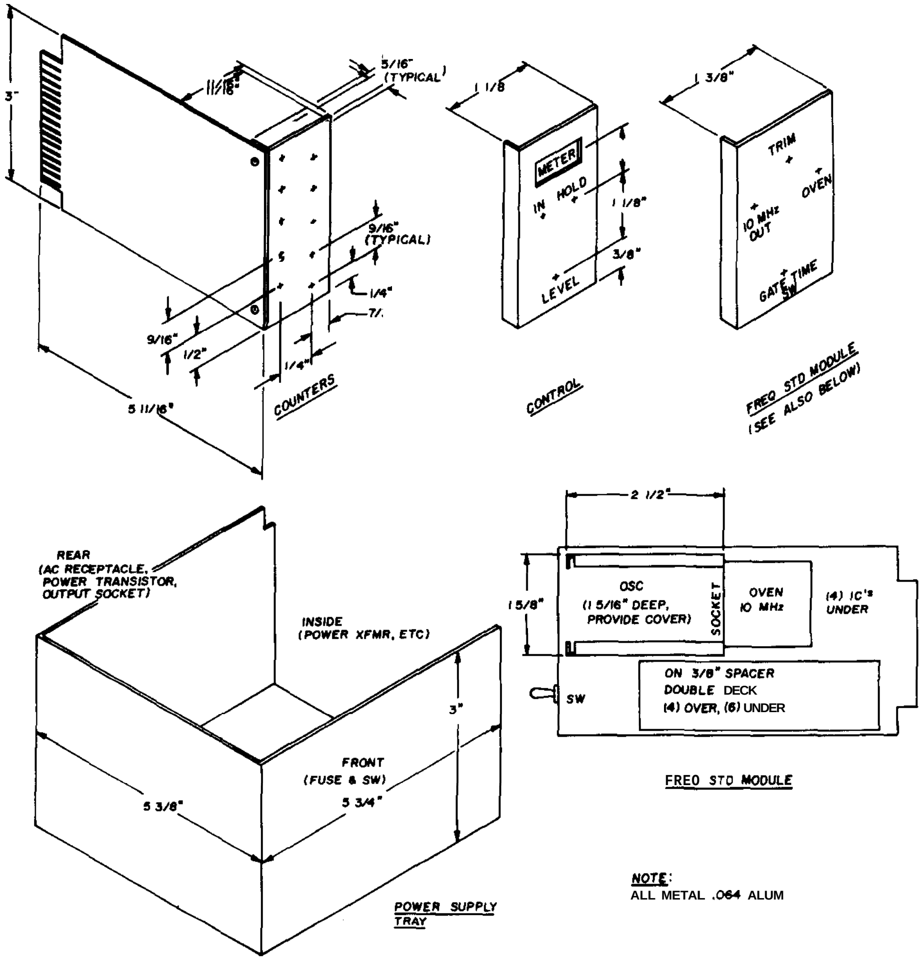


fig. 11. Module sizes.

readings were within 700 Hz of the true frequency. This works out to be a maximum error of 15 Hz per 100 kHz or 1.5 Hz per 10 kHz and compares favorably with an earlier counter that used a 100-kHz crystal-controlled time base. Performance with the 60-Hz time base was so good it doesn't appear worthwhile to use an ordinary 100-kHz standard since the 60-Hz unit can be built for one-third the cost.

On the negative side, short-term stability does not match the crystal. Readings changed continuously within the limits noted above. For example, on one count the reading might be 200 Hz low and the next, 500 Hz high. Readings with the crystal-controlled time base never drifted any more than 1 Hz.

These variations are a result of the line voltage characteristics—there is no precise point on the ac waveform that stays constant—the waveshape and slope change from cycle to cycle. In addition, there are small voltage spikes on the line which can cause the monostable multivibrator to misfire. Once the one-shot has fired, the frequency division is very accurate, but the variation in trigger point is an unavoidable drawback of the power-line time base.

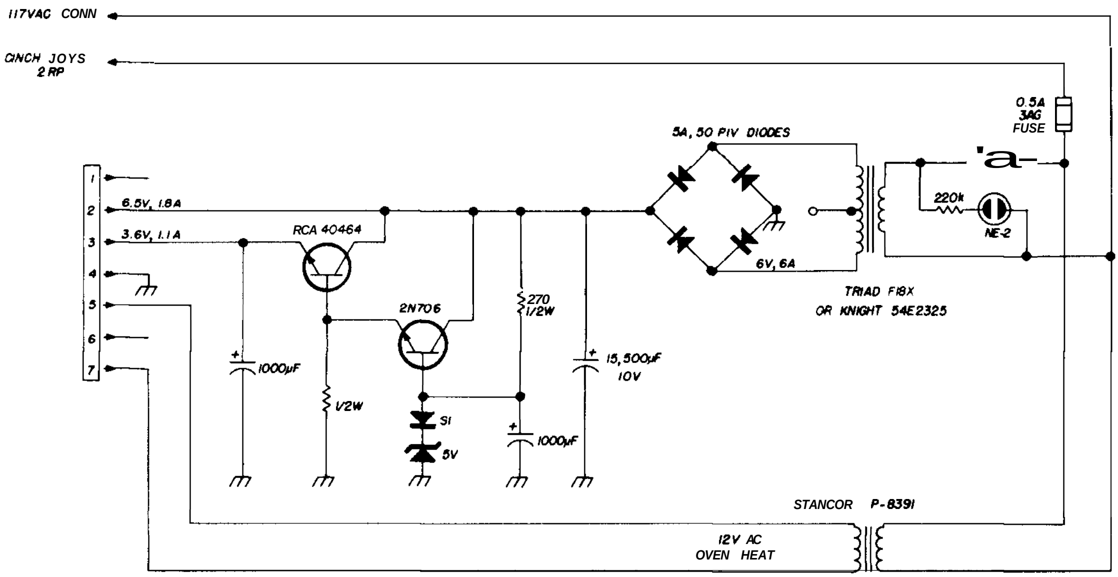


fig. 12. Power supply for the digital frequency counter.

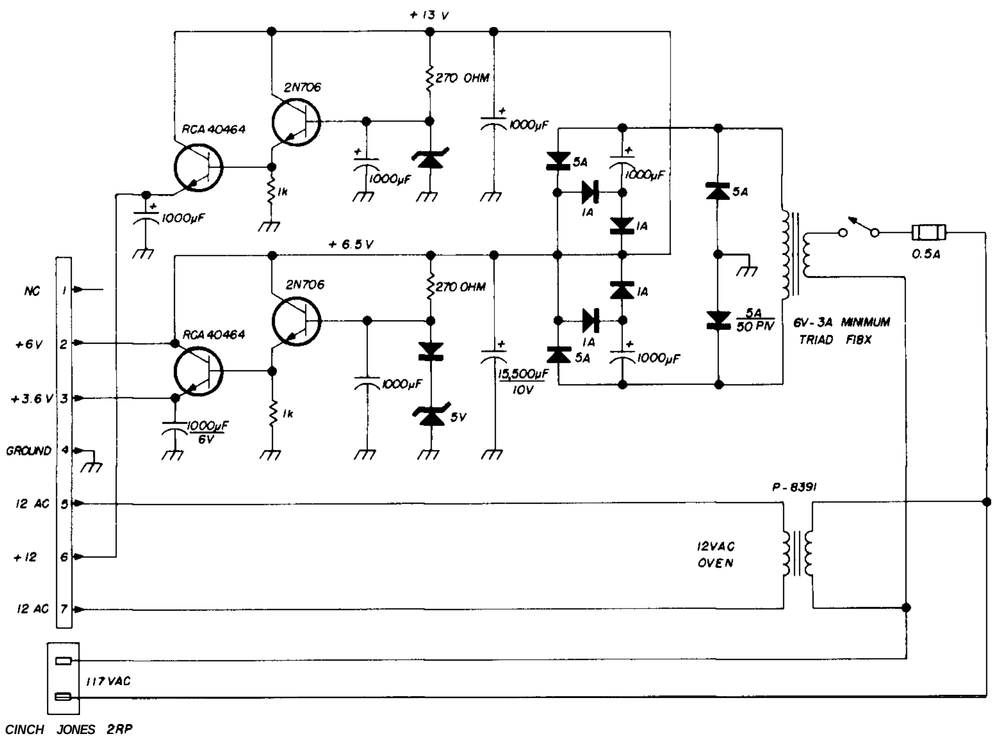


fig. 13. Power supply circuit if the high-impedance input circuit is used.

counter decades

If the counter decade kit* is used, construction can be relatively simple. The boards can be assembled and bolted to a bracket assembly that contains the indicator lights and the printed-circuit plug. There are some minor differences between the commercial kit and the counters used in this version.

In the commercial unit, 33-ohm swamping resistors were used to equalize indicator lamp drain, and all lights glow dimly all the time. This greatly increases the total current drain on the power supply. The first version did use this swamping resistor circuit, however, and it worked very well. The pilot light glow is so dim that it is not a problem.

In the second version of the counter, I decided to include a blanking circuit to shut off lights during the time they were not indicating a count. Diodes were placed in series with each light to remove sneak paths. This eliminated two 33-ohm equalizing resistors. The current drain on the 6-V supply for each module dropped to 50 mA. The 6-V power for the lamps and the MC715P's was separated, and the lamp voltage was then controlled by a blanking circuit housed in the left decade. This decade is a simplified version of the others, and there is plenty of spare room for it.

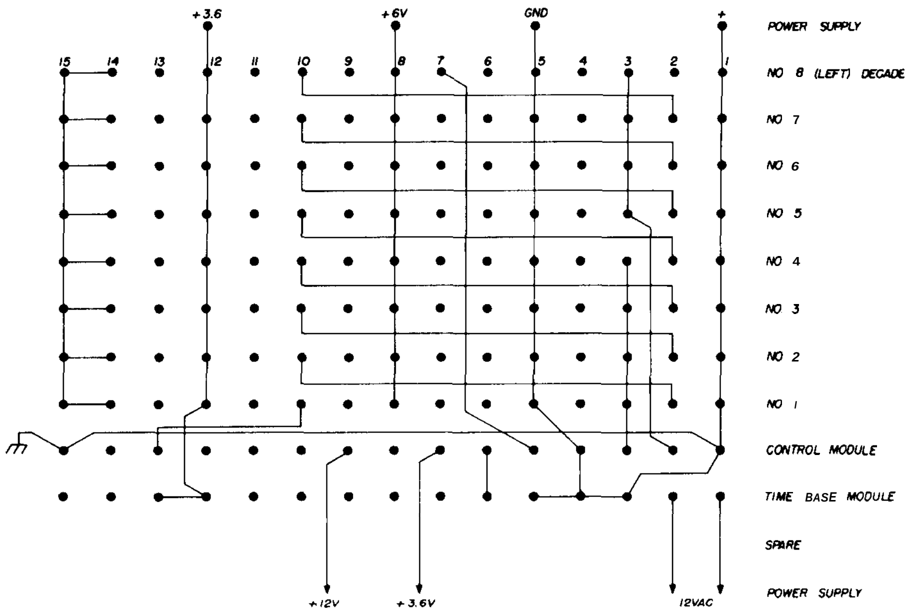


fig. 14. Module interconnections. This shows the wiring that must be run between the various circuit-board receptacles.

Construction of the decades **(fig. *5*)** is well covered elsewhere.[2] It's unnecessary to add more except to note that it's a good idea to check each light with an ohmmeter and to check the completed module for about 40 ohms between ground and the 3.6-V pin, observing ohmmeter polarity. The 6-V lamp circuit should also be checked for a short.

The completed decades can be checked by connecting the I-Hz output from the time base module to the "count" input of the decade. When power is applied, the lights should advance at the rate of 1 each second, to the top of the decade, and start over again at zero. In the event of trouble, read

* Decade kits include etched and drilled printed-circuit board, set of 10 lamps with plastic covers and spare bulb, and complete set of electronic parts. Order from Southwest Technical Products Corporation, 219 W. Rhapsody, San Antonio, Texas 78216. $12.00 postpaid in the U.S.A. One kit required for each decade.

10-MHz oscillator board

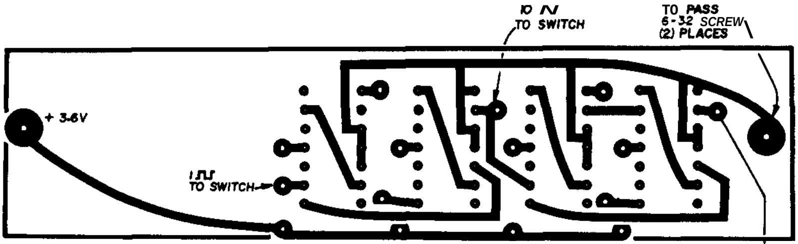
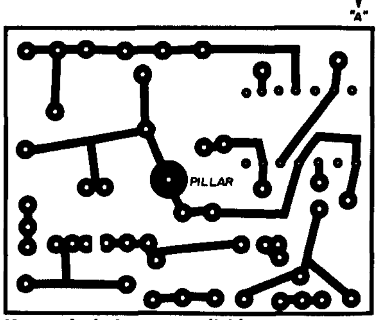
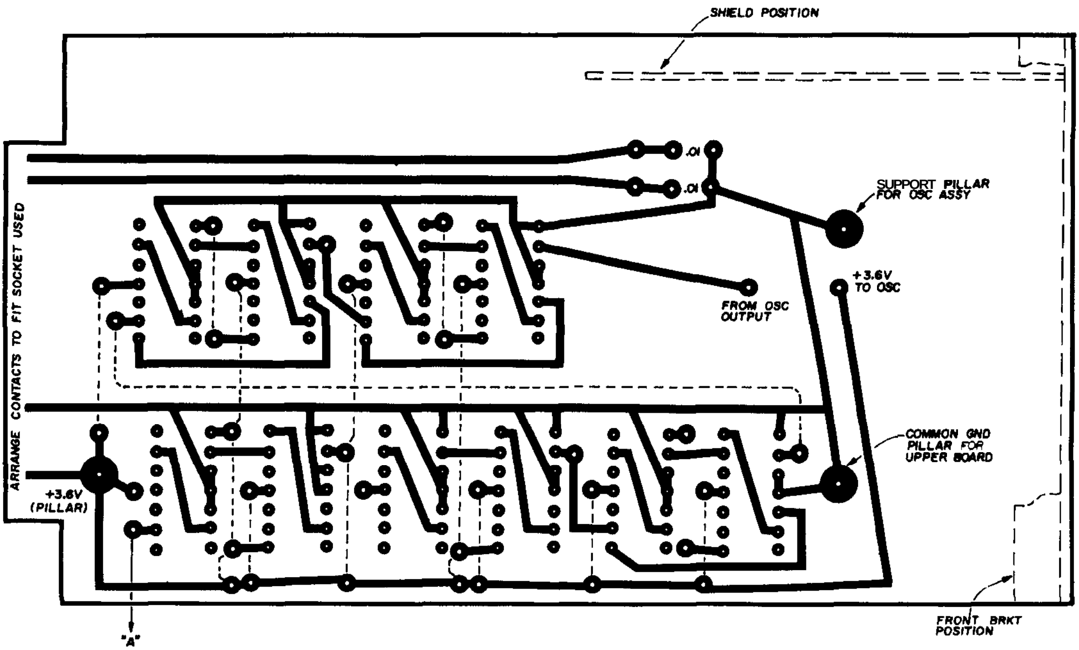


fig. 15. Printed circuit boards for the frequency-standard module. Boards are positioned as shown in fig. 11. Pillars are used for both support and as electrical paths.



Upper deck frequency divider



Basic control-module board. Small dotted lines indicate wiring underneath the board.

the circuit theory and check with a multimeter or scope. It is worth keeping in mind that the IC's are the least likely to cause trouble.

control module

After three or four decades are finished, the counter can read out audio frequencies, and the control module construction can begin. The control module (**fig.** 7) is the heart of the unit, so its circuits will be explained.

In order to count frequency in the electronic counter, several operations must take place in the proper sequence. The input sig-

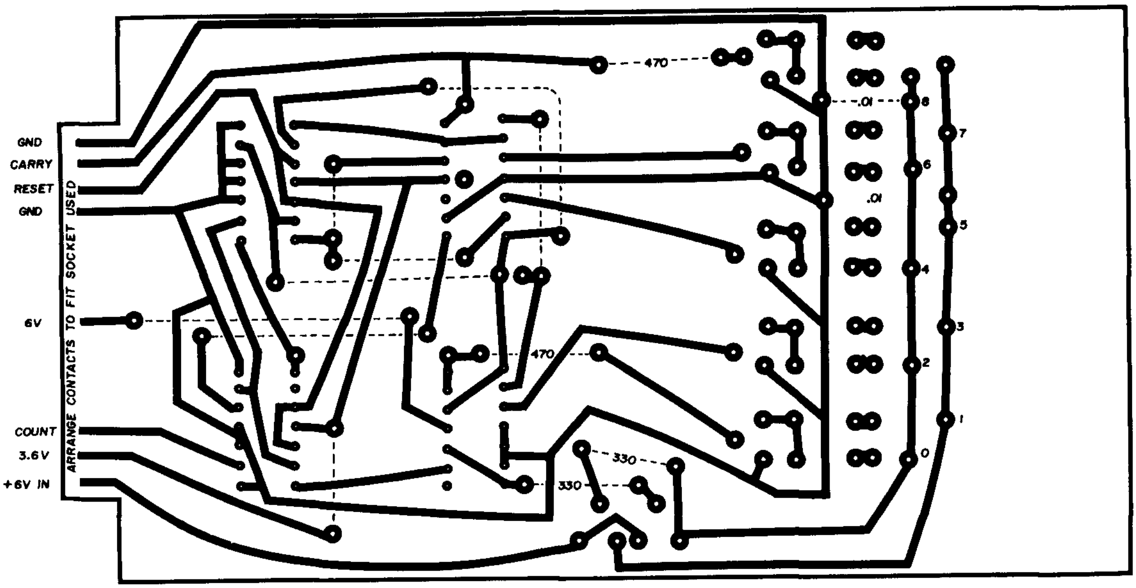


fig. 16. Printed circuit board for the decade counter.

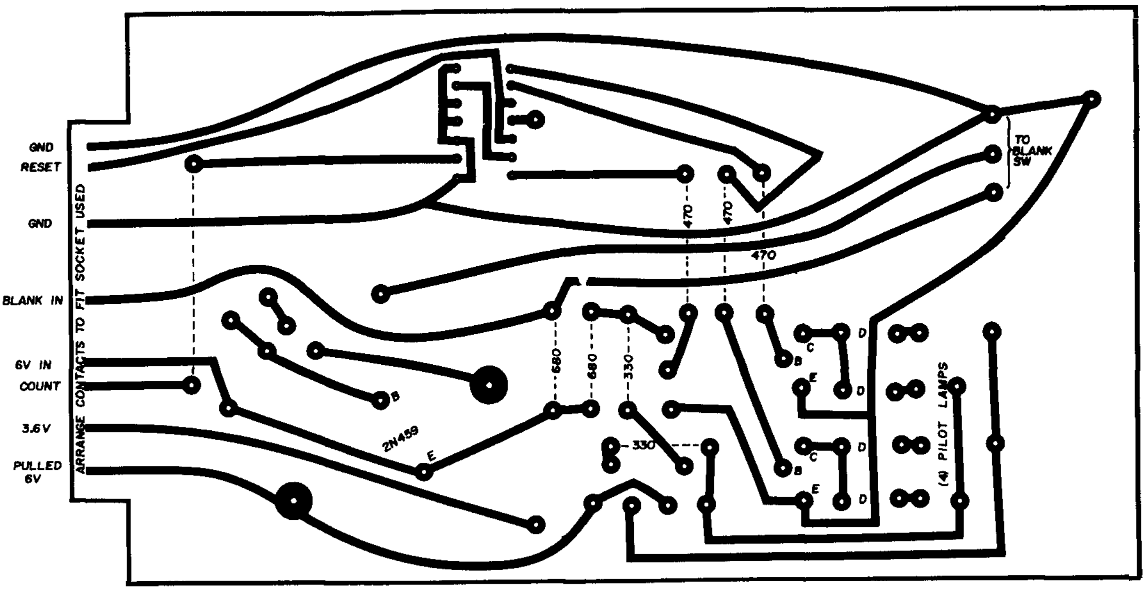


fig. 17. Printed circuit board for the left-hand counter and lamp blanking circuit.

nal must be shaped and limited so it has uniform amplitude and a fast fall time. The optimum fall time is in the order of 100 nanoseconds. The processed signal must then be gated on and off for exactly one second while the counting process takes place. A two-second interval then follows during which the pilot lights are read. A reset pulse follows to reset all decades back to zero for the next count. The process then repeats.

input circuit

The signal enters the control module via the phono jack on the front panel and appears across the input level pot. The input impedance of the circut is set by the high impedance of the emitter follower stage and the resistance of the pot. The limitations as to maximum dc input voltage is set by the rating of the 1-μF coupling capacitor, and high voltage surges and excessive ac input levels are clamped by the two reverse biased diodes across the input line. These diodes aren't expected to stand extreme surges, so it is good practice to connect the counter with the input pot turned down. Note the 33-ohm resistor in series with the 12-V power supply that is intended to protect the power supply if these diodes short.

amplifier

The next stage is an amplifier with a passband from 20 Hz to 25 MHz. The inductor (L1) is a peaking coil to extend the frequency range and is composed of a 318-inch long winding of number-30 enamel wire on a Meshna number-479 coil form. The best way to adjust this circuit is to use a TV sweep generator, disconnect the 22-μF capacitor at the junction of the 3300- and 8200-ohm resistors and substitute a 1000-ohm load to ground.

Connect the scope demodulator probe across this I k resistor and vary the inductance until the passband extends to 25 MHz and falls off immediately after. Check the dc voltage from the 2N706 collector to ground to see if it is from 6 to 7 V. If not, the value of the 39k bias resistor must be changed. The idea is to drop half the supply voltage across the transistor.

schmidt trigger

After the signal leaves the amplifier stage it enters the Schmidt trigger circuit. This circuit converts the input signal to a square wave. Due to the wide tolerances in transistors it is necessary to adjust the 8200-ohm bias resistor to the input transistor used, and it may be convenient to temporarily connect a pot to the circuit until the circuit is optimized. It is worthwhile to spend the time necessary to select the best transistors for the Schmidt trigger circuit.

Assuming the wide-band amplifier has been checked out and is working properly, the procedure is to connect an unrnodulated signal generator to the input jack and a wideband oscilloscope to the 150-ohm collector load resistor. Temporarily disconnect the 220-pF mica coupler to the MC799P. With no input signal the voltage on the collector of Q3 is high. Adjust the 8200-ohm resistor bias until Q3 is on the stable edge of conduction.

Now a signal applied to the base of Q3 causes it to conduct, the collector voltage drops sharply, dropping the forward bias on Q4 base, and the greatly reduced current through Q3 reduces the emitter bias on the 22-ohm resistor. This reduced bias accelerates the action in Q3 until Q3 reaches saturation. Q3 in saturation cuts off Q4 until the reverse polarity of the incoming signal again exerts control on Q3 base and starts a reduction in Q3 collector current. The reverse action then takes place with Q4 in saturation.

With an input signal of from 6 to 10 MHz, the Schmidt trigger should provide a good square wave of approximately 50-percent duty cycle at a noncritical adjustment of the input level pot. The square wave will be rounded somewhat at 10 MHz (possibly due to the scope used here), but should have excellent waveform at 4 or 5 MHz. If it does not, substitute new transistors for Q3 or 4 and try different diodes. The purpose of the diode from Q3 base to ground is to clamp Q3's bias with input level changes and aid in obtaining a 50-percent duty cycle square wave. The diode coupler between Q3 and Q4 further sharpens signal transitions.

Once the Schmidt trigger is working properly, the 220-pF capacitor can be reconnected and the 470-ohm resistor in the Q5 gate base circuit lifted. Check the 10-MHz at pin 3 of the MC799P (IC4). It should have a good waveform. The MC799P extends the frequency range of the counter, due probably to the fact that it drives the counter decade with a low impedance driving source.

the level meter

Since the input signal must drive Schmidt trigger Q4 properly if the unit is to count, use is made of this to operate a level meter connected to Q4's collector. This meter is not absolutely essential to the operation of the counter, but it is very useful and is recommended. The best meter located so far is a Calrad EW75L level meter with a 140-μA movement. I was unable to locate a second Calrad meter for this version of the counter and therefore used a Lafayette meter from a Citizen's Band transceiver of unknown type number. It is adequate, and the main thing to keep in mind when searching for a substitute is to try to locate a 100- to 150-μA movement. Modify the 5600-ohm resistor as required to make the meter read properly

base is near zero volts. When pin 9 of IC1 goes positive, it shuts off Q5 gate, and the gate remains shorted for the following three seconds.

After a two-second interval, the voltage on pin 14 of IC1 rises, and a spike is coupled through the 150-pF capacitor. This triggers the one-shot section of IC2, and it puts out a pulse 15 micro-seconds long. This pulse is

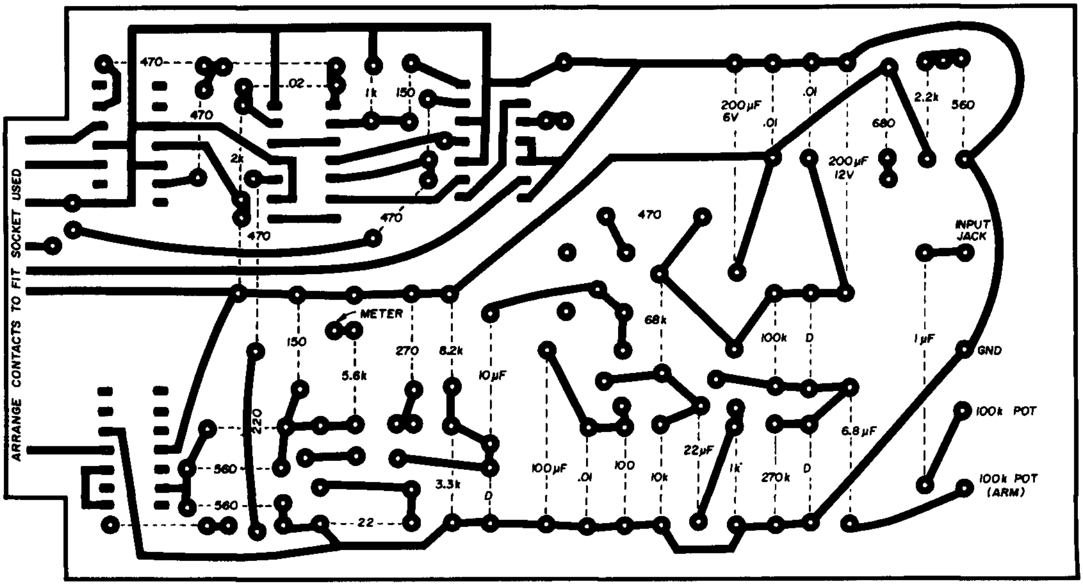


fig. 18. Printed-circuit board for the control module.

when the trigger is working under optimum conditions, then reconnect the 470-ohm resistor on Q5 base.

gate switch

Once the signal is properly processed, it must be gated on and off. This is accomplished by the three integrated circuits in the control module. The MC790P (IC1) divides the 1-Hz time base signal to 0.5 Hz and 0.25 Hz. These two signals are then applied to a gate that provides an output at pin 14 of IC2 only when both signals to pin 12 and 13 on IC2 are low at the same time. This is the count interval. It is coupled by the inverter stage of IC2 to the base of Q5 gate.

Q5 acts as a short to the count signal when its base is high, and passes the count signal during those 1-second intervals when the

inverted and amplified in both sections of IC3, and the two outputs reset 4 decades each. The reset pulse occurs during a one-second interval just prior to the count. Therefore there is more than sufficient time for the reset action (15 μs) to take place.

The "hold" button holds up the count-reset process so that a multi-digit count can be read easily from the pilot lights. It holds the voltage on pin 13 of IC1 to a "high," and therefore the counter will not reset or count until the button is released. It can be pressed to hold the next count anytime after the reset action has taken place.

Since the 4-second count time is a little slow for adjusting some equipment, the action can be speeded up by switching the "gate time" switch on the time-base module to 0.1 second. This speeds up the process by a factor of 10 and also results in the counter dropping the last digit. The whole count shifts to the right one decade. This is a convenient position when accuracy down to the last Hz is not needed. It follows that a 3-decade counter will then count four digits; however it is not practical to speed up the count by another factor of 10 to save decades, because the incandescent pilot lights will not follow the count.

blanking circuit

Since the reset pulse sets all counter decades to zero for the next count, the light action should alternate from a row of zeros to the final count alternately unless a blanking circuit is provided. It is easier to read the count if only the final count is presented and lights are off at all other times.

When the reset and count action is taking place, pin 13 of IC1 is zero. This is applied to the base of Q6, which generates a 6-V pulse at its collector that is used in module 8 to control the circuitry that turns off the lights. When the voltage at Q6 collector is high (6 V), it cuts off the 2N1305, which in turn shuts off the 2N459 power transistor. When the voltage at Q6 collector swings to zero, the 2N1305 is forward biased and turns on the power transistor. This supplies voltage to the pilot lamp circuit, and the lamps that have been enabled by the count circuitry are then lighted.

in the event of trouble

If there is any difficulty in the IC1, IC2, IC3 (control module) circuit the easiest procedure is to use a dual-trace scope and check simultaneous waveforms. Since the 15-ps reset pulse is too fast for easy observation, it is convenient to disconnect the 1-Hz output from the time base module temporarily and substitute the 10-kHz output at the junction of one of the divider pairs. This will speed up the action in the gating reset circuit so that it is more easily followed.

Many hints have already been given that will help in trouble-shooting. In particular, try to debug each assembly as it is finished. If all trouble shooting is left until last, the possible combination of difficulties can be almost overwhelming, or even doom the project.

In the event of trouble with the counter modules, you should carefully note what lights do light, and in what order. If a I-HZ input pulse is applied to the input count pins, JK flip-flop action can be verified by using a multimeter at the flip-flop output terminals. A good flip-flop will show an alternating 0 and 1-V output, while a JK that is hung up will show a fixed 1-V or 0-V output. A hang-up usually means a poor solder connection or short somewhere.

In general, if two or three lights don't light, the JK's are all right, and the trouble is in some circuit common to these lights. This can be deduced by examining the counter block diagram (the one for decades only). Do not overlook the possibility of bad transistors, but if transistors have been previously checked on an ohmmeter before insertion, it is usually more productive to signal trace than to interchange transistors blindly. Dim pilot lights are caused by low beta transistors and by defective lamps.

Most of the hard-to-find parts are available from Allied Radio, 100 N. Western Avenue, Chicago, Illinois 60680. Here is a list, along with Allied catalog numbers, price and shipping weight. Be sure to include enough for postage.

Amphenol 143-012-01 receptacle, Allied 47F2675, $1.21, 3 oz; Amphenol 133-012-21 plug, Allied 47F2857, $2.45, 2 oz; Motorola MC724P quad 2-input NAND/NOR gate, Allied 50F26MC724P-MOT, $1.08, 1 oz; Motorola RTL MC790P dual JK flip-flop, Allied 50F26MC790P-MOT, $2.00, 1 oz; Motorola MC799P dual buffer, Allied 50F26MC799P-MOT, $1.08, 1 oz; zener diode assortment, Allied 24C9340, $1.98, 4 or.

references

1. See the Hewlett Packard test instrument catalog section on electronic counters. Also refer to *Electronics World*, October, 1967.
2. D. Lancaster, "Low Cost Counting Unit," *Popular Electronics*, February, 1968, p. 27.
3. D. Lancaster, "For Low Cost, Count on RTL," Electronics, January, 1968.
4. J. Hall, "Binary-Decimal Counter," Technical Correspondence. *QST*, April, 1968, p. 54.
5. R. Suding, "Cheap and Easy Frequency Counter," 73, November, 1967, p. 6.
6. G. Jones, "An Integrated Circuit Electronic Counter," 73, February, 1968, p. 6; additional notes, *73*, *June*, *1968*, p. 74.

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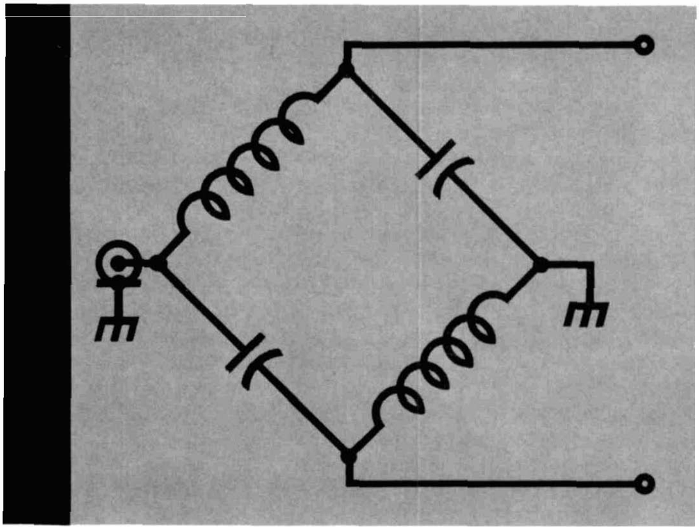

wide-band bridge baluns

B Orr, W ..., Manager, Amateu Servi e,
I. Sa ye WA6 BAN, Pr jec Enginee
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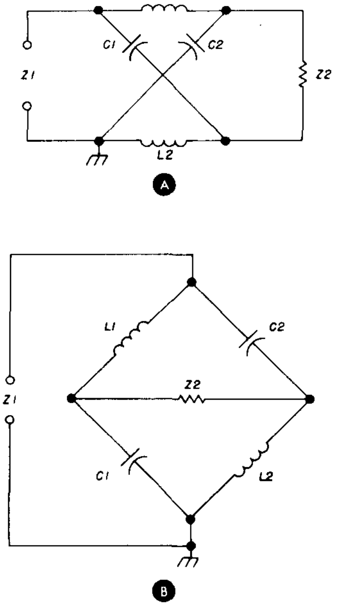
Various balun designs have appeared in the literature which provide a match between balanced and unbalanced systems over various bandwidths and which have transformation ratios of 4-to-1 or 1-to-1.[1, 2, 3]

However, these devices don't solve the problem of matching an antenna with a radiation resistance higher or lower than the nominal 50-ohm value of many coaxial transmission lines. In addition to the desired balancing action, the matching circuit may be called upon to match the transmission line to a load which does not have a 4-to-1 or 1-to-1 ratio to the line impedance.

A single-band Yagi, for example, exhibits a typical radiation resistance value of about 20 ohms, while a quad antenna may have a radiation resistance in the neighborhood of 80 to 100 ohms. How may these devices be properly matched to a 50-ohm unbalanced transmission line? The lumped-constant, symmetrical-bridge balun is the answer.

This simple balun with the unwieldly name is suitable for use as a broad-band matching device capable of coupling an unbalanced 50- or 70-ohm transmission line to a balanced load within the range of about 5 to 1000 ohms. The operational bandwidth of the balun is greater as the transformation ratio between line and load becomes smaller, and it is **theoretically** possible to maintain a 1-to-1 balun transformation over an infinite frequency range.



$L1 = L2$ $C1 = C2$ $$Z_1 = \frac{2(Z_L \cdot Z_C) + (Z_L + Z_C)Z_2}{Z_L + Z_C + 2Z_2}$$

fig. 1. The lumped-constant symmetrical-bridge balun acts as a balancing device and transformer. One side of the input may be grounded without disturbing balance. For proper operation, no coupling should exist between the coils. The circuit is redrawn in B to illustrate the bridge.

As the transformation ratio increases, the operational bandwidth of the balun decreases, as discussed later in the text. This tradeoff of bandwidth for transformation ratio should cause the user no pain, as it is common to all matching devices of this type. Realistically, it is possible to maintain good operational bandwidth (say, over a 2-to-1 frequency span) within the impedance transformation ratios normally encountered in amateur practice. Thus the bridge balun is ideally suited for a monoband- or a triband-beam antenna system.

Two articles[4, 5] provide information for designing symmetrical lattice-style bridge baluns, and these articles are the basis for the graphical information in this article. The graphs reduce calculations to a minimum and make home construction of practical bridge baluns a simple task.

the bridge balun

A symmetrical bridge circuit can be used as a broad-band matching device as shown in **fig. 1**. To design a useful balun, the required transformation ratio in terms of the input and output impedances must be known and the desired operational bandwidth specified. The following design procedure determines balun constants, defines bandwidth and indicates the magnitude of introduced swr caused by the addition of a balun to the antenna system.

The inductance (L) of each balun coil and the value of the associated capacitance (C) may be determined from the following design formulas:

$$Zm = \sqrt{Z_1 \times Z_2} \qquad (1)$$

$$Fm = f_1 + \frac{f_2 - f_1}{3} \qquad (2)$$

$$L = \frac{Zm}{6.28 \times fm} \qquad (3)$$

$$C = \frac{10^6}{6.28 \times fm \times Zm} \qquad (4)$$

$$R = Z_1/Z_2 \text{ or } Z_2/Z_1 \qquad (5)$$

where:

Z = design factor of the balun in ohms.
Z_1 = input impedance in ohms.
Z_2 = output impedance in ohms.
f_m = mid-frequency point of the design range in MHz.
f_1 = lower-frequency limit of the design range in MHz.
f_2 = upper-frequency limit of the design range in MHz.
L = inductance in microhenries.
C = capacitance in oicofarads.
R = transformation ratio.

The general steps for determining the circuit constants and typical operational bandwidth of a particular balun design are:

1. Determine f_1 and f_2 and the maximum desired swr limit at these frequencies.
2. Solve *eq.* 1 for Z.
3. Solve **eq.** 3 and *4* for balun components L and C.
4. Determine the typical operational bandwidth of the balun from appropriate charts **(fig. 4** and **7).**

bridge balun for a yagi

As a practical example, a bridge balun will be designed to match a 22-ohm balanced load (the split driven element of a three-element parasitic-beam antenna) to a 50-ohm unbalanced transmission line. The balun is specified to work at any frequency between 14 MHz (f_1) and 30 MHz (f_2). Determine the constants of the balun and the typical operating range defined by the 1.5:1 swr points* at the feed terminals.

Therefore: Z_1 = 50 ohms and Z_2 = 22 ohms. From **eq.** 1, Z = 33 ohms; from *eq.* 2, $f_m = 14 + 16/3 = 19.3$ MHz; from *eq.* 3,

* Swr limits for a transmission-line system are somewhat arbitrary and are usually chosen to provide a reasonable load for the output matching network of the transmitter, generally 2:1 or less.

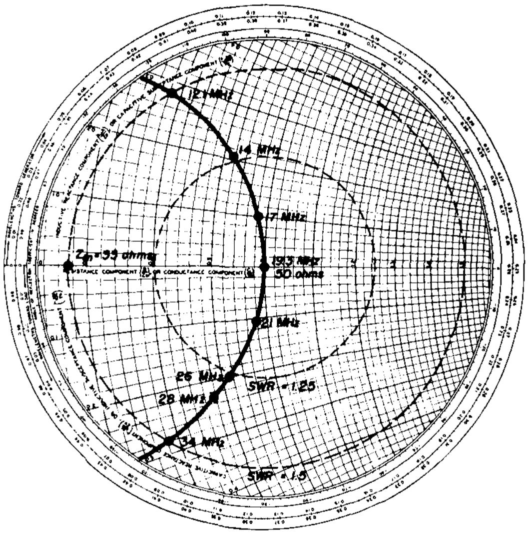
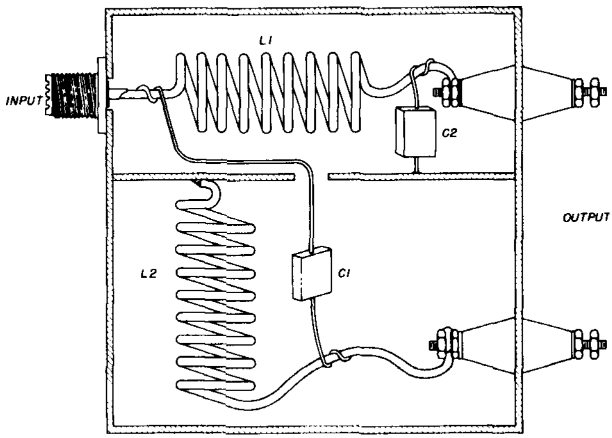


fig. 3. Smith chart' plot of a bridge balun designed to match the balanced 22-ohm **input impedance** of a three-element beam to so-ohm coax. The swr plot passes through the swr = 1.5 circle at 12.1 and 34 MHz; swr is less than **1.25:1** between 14 and 26 MHz.

fig. 2. The bridge balun should be built in a **shielded** box with a center partition running tha length of the interior. The coils ara placed at right angles to each other. For up to several kilowatts, the capacitors should be ceramic types such as the Centralab 850 series; for 100 watts or so, 500-V silver micas **are** suitable. For higher power **levels,** it's recommended that the coils be made from small commercial **air**-wound inductors wound with **number**-14 wire.



$L = 33/6.28 \times 19.3 = 0.272$ μH; from **eq. 4**, $C = 10^6/6.28 \times 19.3 \times 33 = 250$ pF; and from **eq. 5**, $R = 50/22 = 2.28$. The balun constants are shown in **fig. 2.** As a cross check on your mathematics, the derived values of L and C should be resonant at the mid-frequency point (f_m).

lower-frequency limit, $f_1 = k_1 \times f_m$ **(6)**

upper-frequency limit, $f_2 = k_2 \times f_m$ **(7)**

where k_1 and k_2 are simple unit ratios as shown in **fig. 4.**

To determine f_1 and f_2, the maximum swr

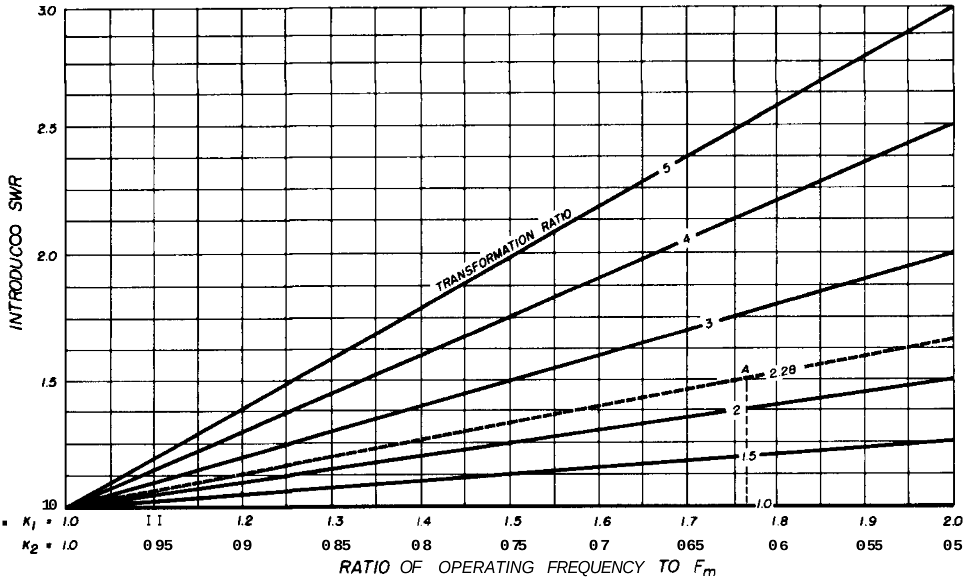


fig. 4. Introduced swr vs operational bandwidth. If the transformation ratio is known, the introduced swr of the balun can be calculated for any operational bandwidth from one-half ($0.5f_m$) to twice the design frequency ($2f_m$). In the example shown, for a transformation ratio of 2.28 and swr less than 1.5:1, the upper and lower frequency limits are 1.76 f_m and 0.63 f_m.

operational bandwidth

The broadband characteristics of the bridge balun can be determined by plotting the introduced swr of the balun against a ratio of the center frequency, f_m, as shown in fig. **4.** As mentioned earlier, the band pass of the balun is inversely proportional to the impedance transformation ratio R.

In the case of the Yagi balun just discussed, the transformation ratio is 2.28, and it is desired that the swr be held to 1.5 or less when the balun is terminated in a "perfect" load. Under these limitations, what is the operational pass band of this design?

The upper- and lower-frequency limits defined for a chosen value of maximum swr are determined by these formulas:

***An expanded** Smith **chart** is **used,** the **General Radio** 5301-7561-NE.

limit is found on the y-axis and the coordinate is traced across the graph until it intercepts the proper ratio line. In this case, no line exists for a ratio of 2.28, and the line must be estimated (dashed line). The intercept is at point A. The intercept of A on the x-axis indicates that k_1 and k_2 are 1.76 and 0.63, respectively. Accordingly, the k factors are applied in **eq.** 6 and **7** to the mid-frequency point of the design range, f_m, as follows:

lower-frequency limit, $f_1 = 0.63 \times 19.3 =$ 12.1 MHz.

upper-frequency limit, $f_2 = 1.76 \times 19.3 =$ 34 MHz.

The operational bandwidth of this design, for a maximum introduced swr of 1.5, is thus 12.1 to 34 MHz.

swr at 20, 15 and 10 meters

The typical introduced swr on the 20-, 15- and 10-meter bands may be determined by establishing k, then using fig. 4 in a reverse manner to that outlined above. The nominal value of k for each band is:

$$k = \frac{\text{desired frequency (MHz)}}{\text{mid-frequency point (MHz)}} \quad (8)$$

These swr values may be interpreted as those values read at the input terminals of the balun when it is terminated in a non-inductive 50-ohm load. As no practical antenna exhibits this magic number across any amateur band, the figure indicates that the existing swr on the transmission line will be increased by this amount when adding the balun, as compared to an imaginary balun

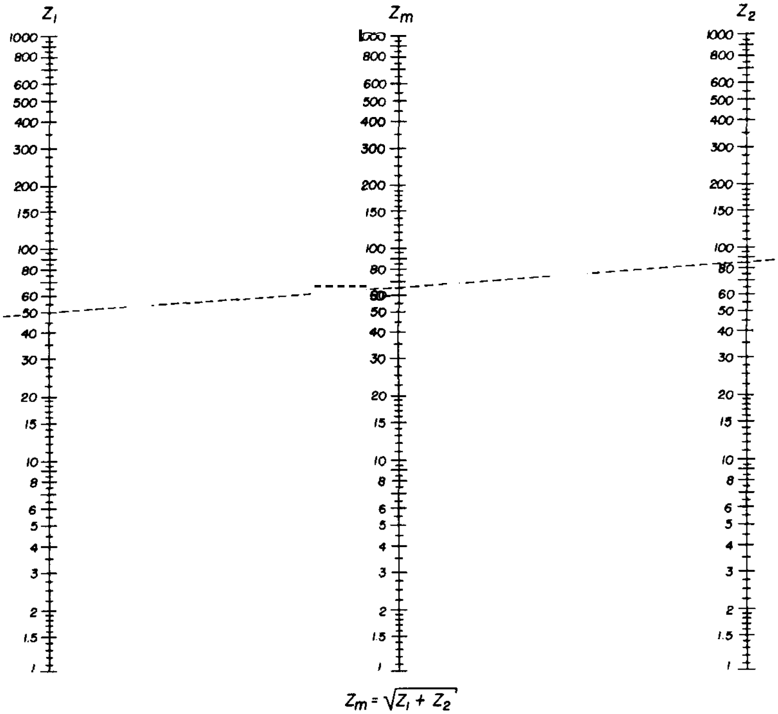


fig. 5. Nomograph to determine Z_m from Z_1 and Z_2. The design factor, Z_m, of a bridge balun can be determined by laying a straight edge across Z_1 and Z_2 and reading Z_m on the middle scale. In the example shown, Z_1 = 50 and Z_2 = 85; Z_m = 66 ohms.

For 20 meters, k = 14/19.3 = 0.725; for 15 meters, k = 21/19.3 = 1.09 and for 10 meters, k = 29.7/19.3 = 1.53.

The typical introduced swr for these bands is now found by entering fig. 4 from the ratio axis (x-axis), meeting the proper ratio line (in this case the dotted 2.28 line) and observing the intercept on the swr axis (y-axis) as follows: for 20 meters, swr = 1.26; for 15 meters, swr = 1.13 and for 10 meters, swr = 1.32.

having a perfect transfortnation ratio.

As a point of interest, **fig. 4** shows that a balun having a 2-to-1 ratio (step-up or step-down) has a frequency coverage (defined by the swr = 1.5 limits) of one-half to twice the mid-frequency design point. Larger transformation ratios are achieved at the expense of higher swr values at the frequency extremes, or by narrowing the bandwidth for a given maximum value of swr. This should cause you no distress, as similar restrictions apply to all matching devices.

bridge balun for the quad

Bridge baluns may be designed from the various charts shown in this article. A good example is the design of a practical balun for quad antennas for the 10-, 15- or 20-meter bands.

It has been claimed by the uninformed that the coax-fed quad antenna needs no balun, since the driven element loop serves as its own balun. This is not so, as the quad is a balanced antenna (both feed points above reference ground) and the coaxial line is an unbalanced device (one conductor at ground). Therefore a balancing device is required to satisfy the demands of symmetry.

In addition, an impedance transformation is called for, since the radiation resistance for a simple quad antenna typically falls about 85 ohms or so depending upon antenna design and element spacing. Taking the figure of 85 ohms as par, then, a balancing device having a transformtaion ratio of 1.7 (85/50) is required.

Since the balun should work with either a 20-, 15- or 10-meter quad (or perhaps a tri-band quad) the mid-frequency of design is again taken to be 19.3 MHz. Here we go!

1. The value of Z_m is determined from **fig.** 5 and is 65 ohms.

2. The chart of **fig. 6** may be used to determine balun constants for a center frequency of 19.3 MHz, and is useful for balun designs for 20, 15 and 10 meters.

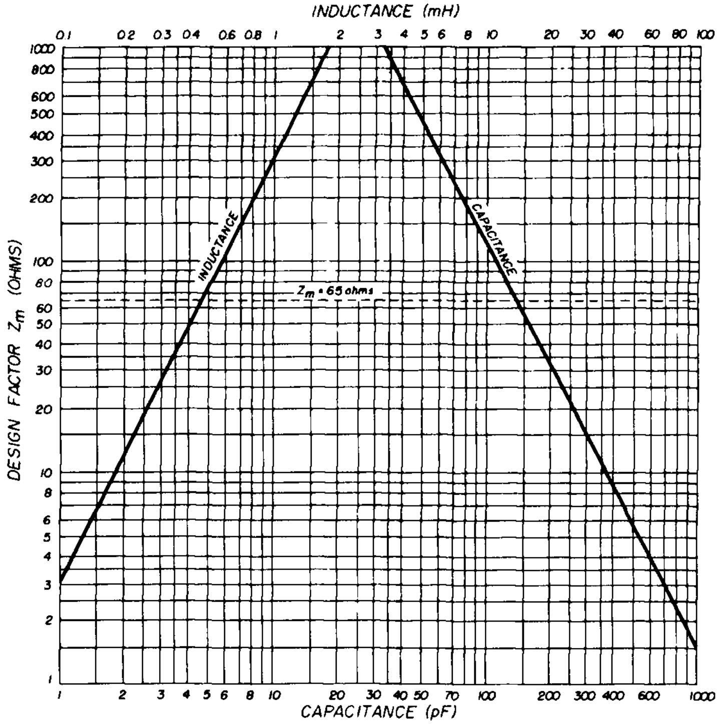


fig. 6. Chart to determine balun LC values when Z_m is known and f_m = 19.3 MHz. In the example shown for Z_m = 65 ohms, L = 0.46 μH and C = 145 pF.

Similar charts may be constructed for other values of f_m for other bands. The chart may be used for load values of Z_2 from 3 to 1000 ohms. The 65-ohm situation is indicated by a dashed line, showing that L = 0.46 μH and C = 145 pF.

The chart of **fig. 4** is again shown in **fig.** 7 in an expanded form for the 20-, 15- and 10-meter bands, useful for transformation ratios up to 4 and for swr limits as high as 1.8/1. For the case of the quad balun, the transformation ratio line of 1.75 may be used as it is sufficiently close to the actual transformation value of 1.7-to-1 to introduce no serious error in the results. The introduced value of swr for this balun design, then, is:

20 meters : 1.14
15 meters : 1.07
10 meters : 1.18

Using the design formulas, charts may be

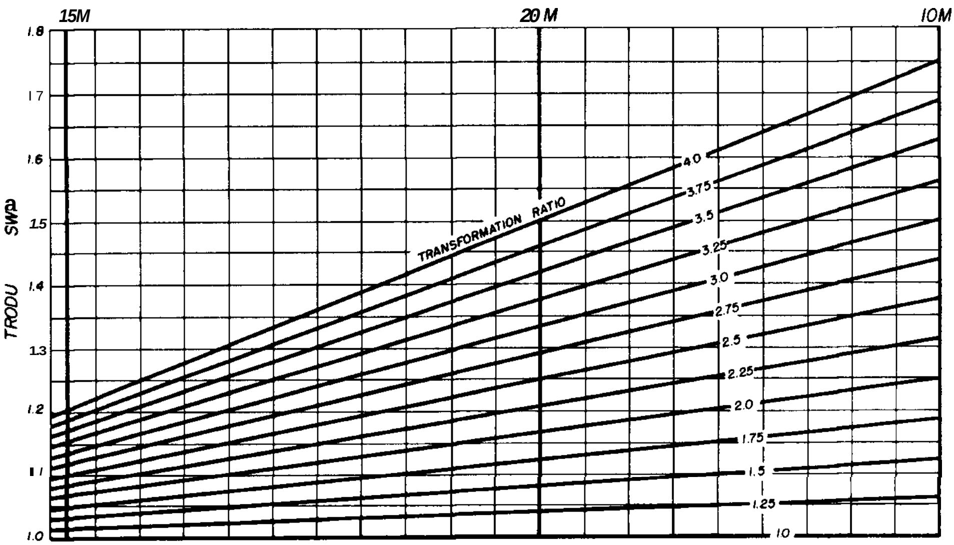


fig. 7. Expanded chart of introduced swr vs operational bandwidth for the 10-, 15- and 20-meter bands when f_m = 19.3 MHz.

constructed for balun design for various frequency spans or for various impedance transformations other than the examples just given.

building your bridge balun

The bridge balun components must be protected from the weather, and a minimum of coupling should exist between the two coils of the device. Other than this, no special precautions must be observed when building the balun. It's suggested that the unit be built in an aluminum box which has a partition placed across it **(fig.** 2).

The coils should be mounted on either side of the partition at right angles to each other. A coaxial fitting is used for the input termination, and two ceramic feedthrough insulators are used for the balanced output leads. Wiring is done with number-12 solid-copper wire using short, direct leads. The compartments should be large enough so that coil Q is not appreciably degraded—a space at least twice the coil diameter is recommended.

The seams of the box should be protected in some way so that water doesn't enter the enclosure. They may be sealed with roofing compound or taped and coated with epoxy cement. When completed, the balun may be mounted to the antenna structure in proximity to the drive point of the antenna. The box should be considered to be at ground potential.

references

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5. J. Frankel, "Coupling Networks Between Unbalanced and Balanced Circuits." Proceedings of the IRE, September, 1941, p. 486.

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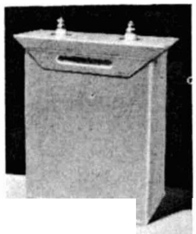

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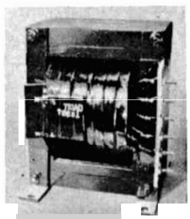

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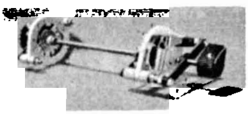

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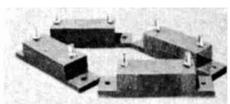

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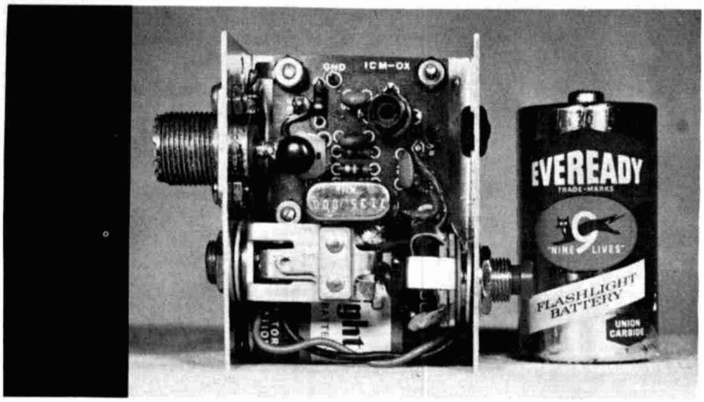


the needle swings to QRP

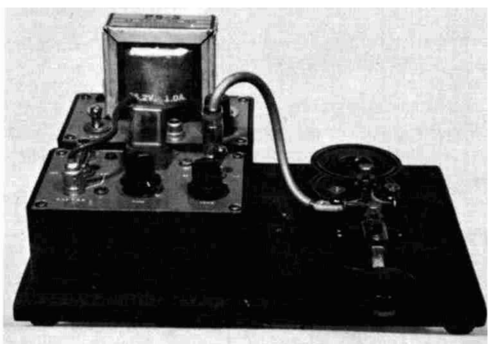
A run down on QRP operation and some of the equipment being used to set new DX records

Howard S. Pyle, W7OE, 3434-74th Avenue, SE, Mercer Island, Washington 980

Are you aware of the rapidly increasing interest in low-power operation on the ham bands? "CQ QRP" is heard with increasing frequency; as more and more records are publicized, there's a tendency to experience new thrills and exciting contacts through the medium of flea-powered gear.

When we hear of break-throughs such as the recent performance of W6TYP who worked New Jersey with 50 milliwatts input on 40 meters and earned the top award of the International QRP Amateur Radio Cluh crediting him with 210,000 **miles-per-watt,** it

The Omega Electronics 5-watt 80- and 40-meter transmitter and power supply.



is exciting news indeed! To the best of my knowledge, Art currently holds the world's record for miles per watt. . . any challengers?

What constitutes QRP? As one of the 'Q' signals established by the Radio Act of 1912, it is interpreted as, "Decrease power" or, if followed by a question mark, "Must I decrease power?" It was originally used to afford relief from inherently broad spark signals in the early maritime wireless service.

Loosely interpreted in amateur practice, QRP simply means that the station using it is in the low-power class as distinguished from the higher powered hams with their kilowatts. While no definite criterion has been formally established as a dividing line, hams generally consider 100 watts input (200 PEP) as maximum power input in the low-power category. This is the figure used by the QRP Amateur Radio Club. Between 100 and 500 watts is generally classified as medium power and anything from 500 watts to the legal maximum automatically falls into the high-power group.

With the steady growth of amateur radio and the proporttonate increase in high power stations a few years ago, the interference problem was becoming acute. A small proportion of hams were able to meet the high power competition by acquirtng a kilowatt of their own; the little fellow found it increasingly clifficult to fight through the powerful signals. As a result, a group of western hams, led by K6JSS, organized the QRP Amateur Radio Club eight years ago. They were committed to a maximum power input of 100 watts (200 PEP). Their primary aim was to encourage more extensive use of low-power in an effort to reduce QRM ancl provide more equitable communication status for the modestly powered ham station.

Initial enthusiasm wasn't great. "How are you going to buck a kilowatt with 100 watts or less?" was the question on many ham

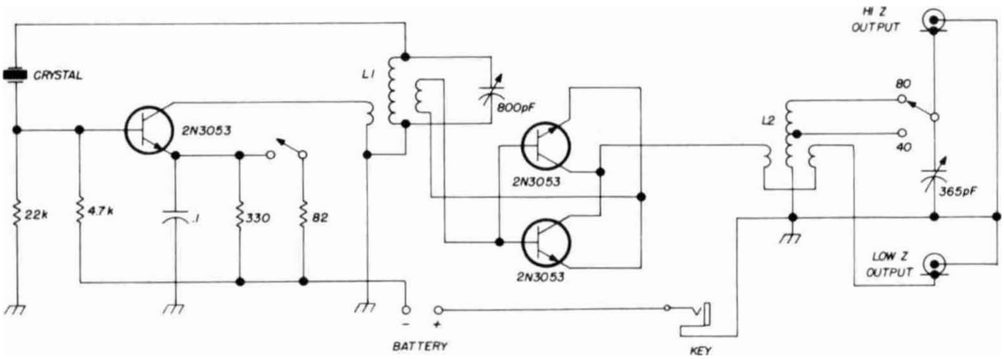
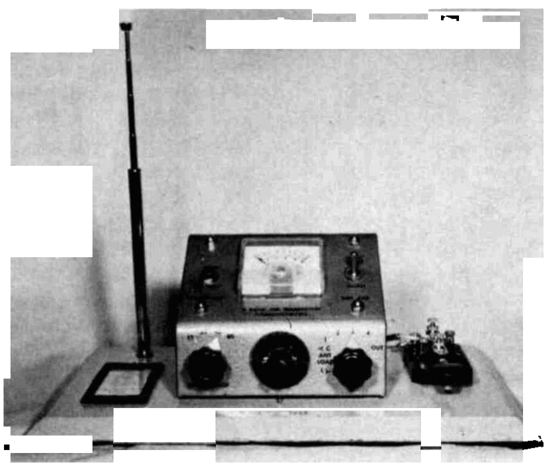


fig. 1. **QRP rig used at W7IGV.** Power input depends on battery voltage and runs from 1 to 5 watts. L1 is number-30 enameled wire on an Amidon* T-50-2 toroid, wound to fully fill the circumference. L2 is 32 turns number-18 wire on an Amidon T-80-2 toroid: 40-meter tap at 14 turns.

Four-band QRP transmitter uses 4 International Crystal OX oscillators and has a built-in whip antenna and antenna tuning unit.

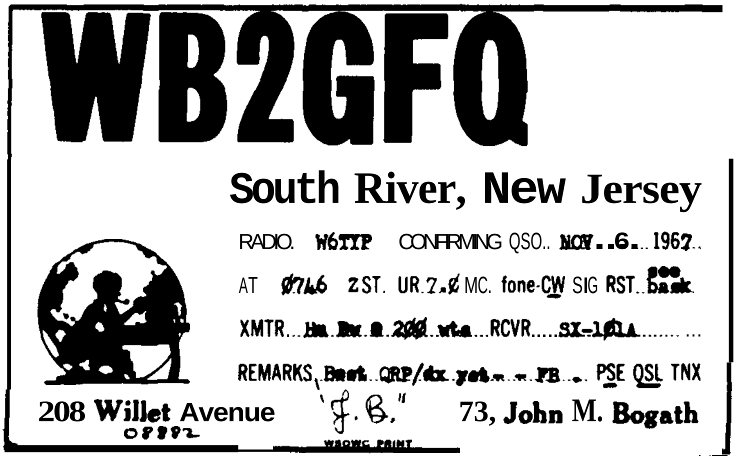


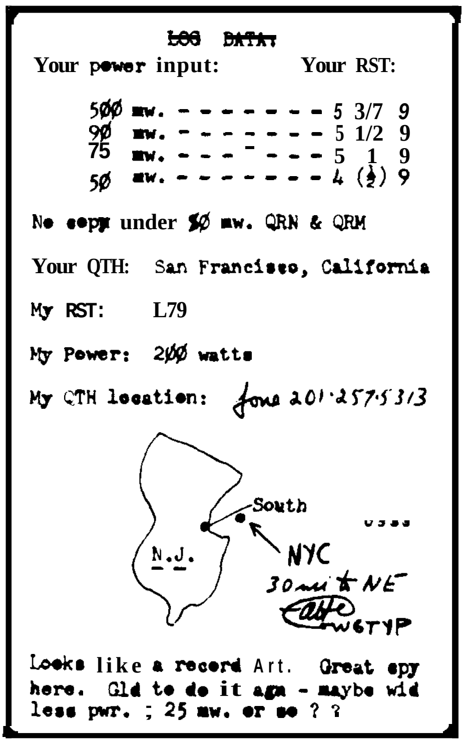
* Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607. T-50-2 toroid is 45c; T-80-2 toroid is 60c. Add 25c per order for packing and shipping

tongues. Rolling up their sleeves, the club members went to work. Their objective was to increase equipment and antenna efficiency to the greatest possible extent and improve their operating skills at the same time.

Careful impedance matching between transmitters and antennas, improved output coupling and more effective antennas were all given careful attention. Operating proficiency was upgraded — not only in actual communication techniques but by choice of bands vs time of day, frequency selection for the desired distance and by finding holes that were relatively interference free.

All these points generated increased interest in QRP operation. As proficiency and





Front and back of QSL card confirming **W6TYP's 210,000-miles-per-watt contact on 7 MHz.**

technical advancement improved, more and more of the low-powered class began gradual power reductions; going gradually downward from 100 watts to 75, 50, 25 and finding that with their new skills they were establishing some rather amazing records. A few hardy souls even dipped into the 5- and 10-watt field and experienced surprisingly good coverage.

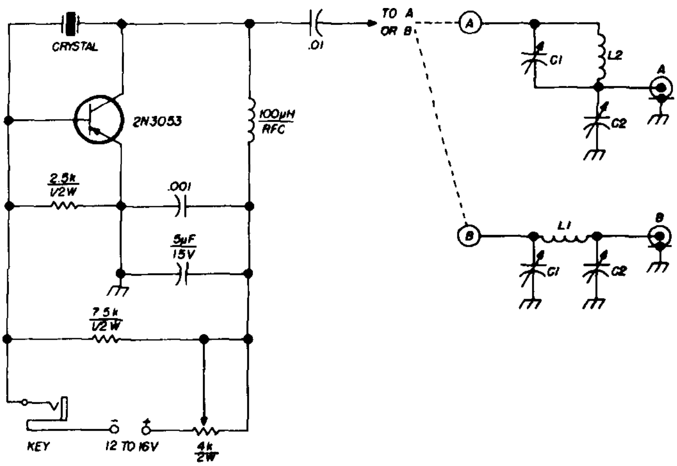
However, a plateau was reached in the range of 5 to 25 watts. They had just about reached their limit in low-power performance with available vacuum tubes; what now? How about those little transistors which had been nudging the electronics field the past few years? Were there hidden possibilities there? After all, transistorized, pocket-sized broadcast receivers were beginning to make more than a casual dent in radio reception equipment, and toy shops and department stores were offering walkie-talkies.

A few hams took a second look and with tongue in cheek gingerly tackled the problem. Some made a couple of half-hearted tries with little or no success and abandoned the project. Those with more determination stuck with it and began to encounter some delightful surprises! Most of these latter experimenters were QRP club members who had the advantage of working as an organized, though separated, group who could exchange experiences on common ground.

Probably the most dedicated member of the group was Art Child, W6TYP, who envisioned the untouched possibilities of the little genies and went seriously to work with them. We mentioned Art's phenomenal 210,000 miles-per-watt record earlier. Suffice it to say that his steady progress has gained his unqualified acceptance by the QRP fraternity as "King of the QRPP'ers." The extra P is applied to rigs that run with an input power of one watt or less!

fig. 2 Rig used by **W6TYP** for his record-breaking 210,000-miles-per-watt contact with **WB2GFQ.** The **4k, 2W** potentiometer permits input power adjustment from 1 to 500 **milliwatts. Output** circuit A is used with **50-** or 75-ohm coaxial lines: **B** is used with 600- to 1000-ohm feedlines.

C1 is a **970-pF** trimmer **(Arco L-314);**
C2 is a **2830-pF** trimmer **(Arco L-306);**
L1 is 50 turns number-28 enameled, close wound on 1/4" form;
L2 is 12 turns number-28 enameled, close wound on 1/4" form



equipment

Let's take a look at some of the gear being used to make 1000-mile-per-watt contacts every day. These are just a few of the mini-rigs currently being used and have demonstrated exceptional performance. Almost without exception, equipment is all homebrewed. At this writing I know of only one manufacturer who offers a commercially-built transistorized amateur transmitter in the 5 watts or less category. Omega Electronics* pulled the rug out from under the other manufacturers with their little LT-5 transmitter and its companion power unit, the PS-5.

It's not my intent to give you complete construction details for any of these little rigs. However, with the schematics you can use your ingenuity and work up your own component layouts. I'll wager you'll find a whole new world opening before you!

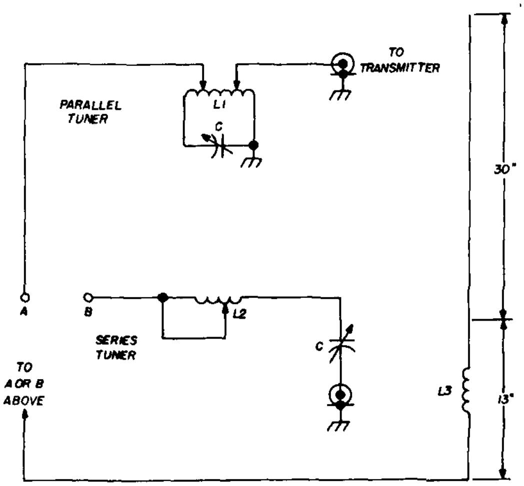


fig. 3 Simple vertical antenna **used** by **W6TYP** with **milliwatt** transmitters

C **365 pF to** cover through 16 meters: **50 pF adequate** for 40 **meters.**

L1, L2 36 turns number-14 enameled. **2" diameter.**

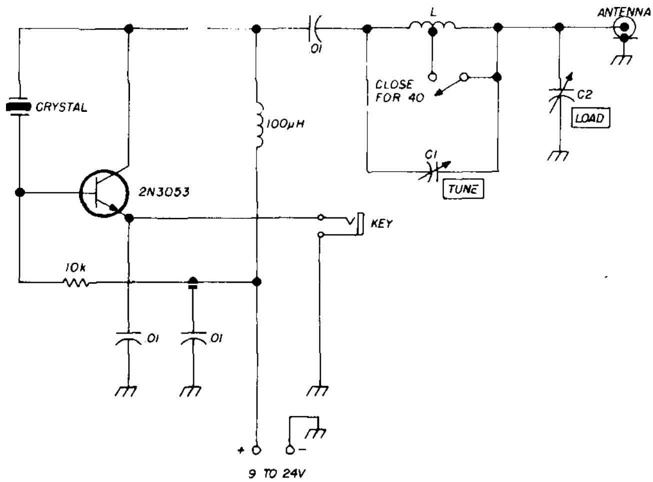
L3 **195** turns number-14 closewound on **3/4"** wood dowel 48" long.

* Omega Electronics Company, 70463 Roselle Street, San Diego, California 92121.

Naturally, you'll have some questions; I'll try to anticipate them and offer some answers. First will undoubtedly be, "What about an antenna . . . what do I use?" That is an easy one; if you have a satisfactory antenna which produces results on your big rig, use it! That's what 99% of today's transistor experimenters are using. If it performs well on your main transmitter, you can be pretty sure it will deliver equally well with mini-power. But be sure it's a good antenna first; well-matched, low swr, in the clear and built to stay put.

I've been asked if putting the mini-powered rig directly at the base of the transmission line just below the antenna or, if vertical, right at the base insulator is desirable. Sure, if you can manage it; you'll lower your losses in the transmission line. However, from the experience many of the boys have had, what you gain isn't worth the mechanical complications.

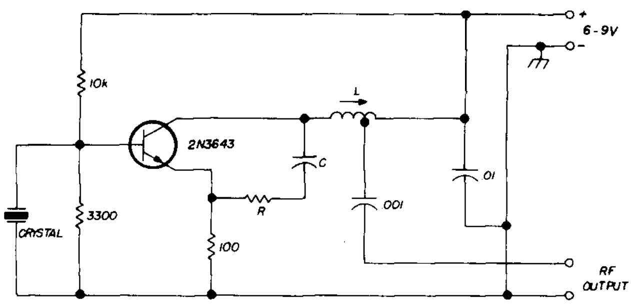
fig. 5. Circuit used by Omega Electronics in their LT-5 five-watt transmitter. C1 is 100 pF; C2 is 365 pF; L is 36 turns 1" diameter, tapped at 18 turns for 40 meters.



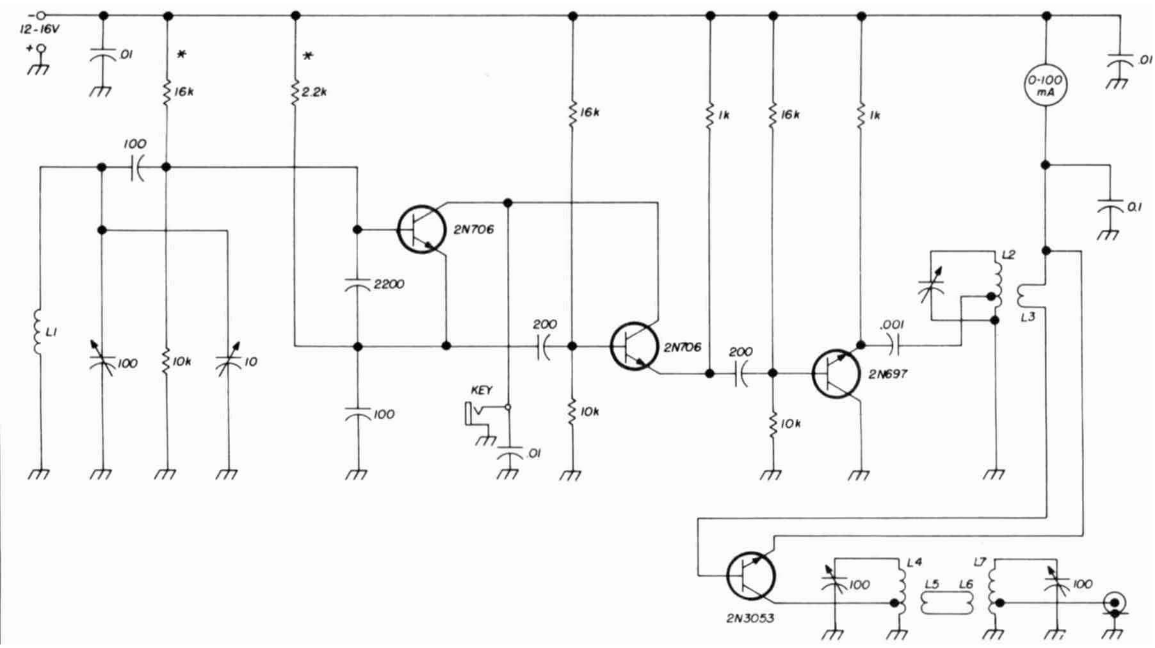
W6TYP uses a random length of transmission line—15 to 25 feet. I use my 40-meter half-wave dipole with one hundred feet of RG-59/U and have worked over 900 miles with .185 watts input! So, if you have an effective antenna now, you have no problem.

W6TYP is handicapped for space for a wire antenna, and practically all his work has been done with a Joy Stick extending from the fourth-floor window of his hotel room.

fig. 4. International Crystal Company's OX oscillator circuit. The values of L, C and R vary with frequency and are furnished with the kit.' Circuit may be keyed in either the plus or minus battery lead.



* $2.35 postage paid from International Crystal Manufacturing Company, Inc , 10 North Lee, Oklahoma City, Oklahoma 73102. EX crystals for the OX oscillator are $3.75 each postage paid.



L1 18 turns number **20.** aimound on **1"** diemeter.

L2 **30** turns number **20.** **airwound** on **1"** diameter. tapped at **12** turns.

L3 **3** turns around cold end of **L2.**

L4, L7 **30** turns number **20.** aimound on **1"** diameter, tapped at 6 turns.

L5, L6 **3** turns around cold end of **L4** and **L7** **respec-**tively.

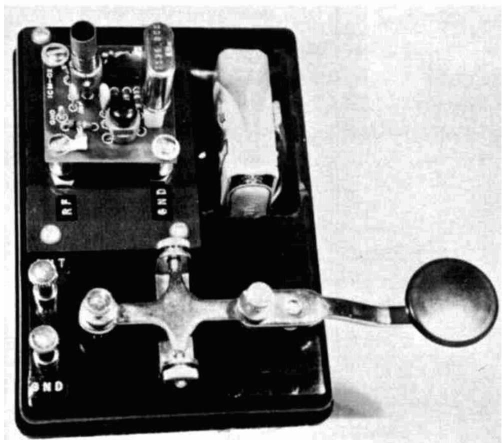
fig. 6. This little rig developed by **W6EAC** runs one to five watts with vfo control. Resistors marked with an asterisk should be adjusted for **best** keying.

He also uses a modified version of this antenna which is shown in fig. 3. Art has tried dipoles and Windoms on the hotel roof, long wires from here to there and variations of these, but he always comes back to a single, compact vertical for assurance of real success.

Next you'll ask, "What bands are being used?" For the most part, QRP operation has centered around 40 meters, but results are just as satisfying on any of the popular bands. For example, with four little OX oscillators from the International Crystal Company I have realized excellent performance on 15, 20, 40 and 80; a four-band arrangement with a separate 2-inch-square transmitter on each band! A few experimenters have had satisfactory reports up to several hundred miles on 160 meters with inputs of one watt or less!

And now, "Are all of the rigs crystal controlled!" Not by a long shot; W6EAC is

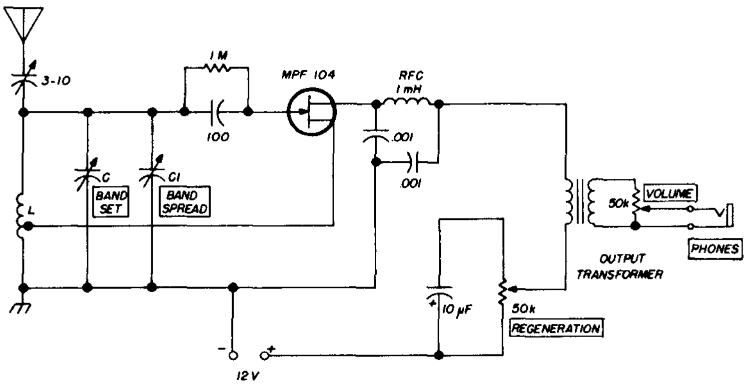
QRP transmitter at W7NUN uses an International Crystal OX oscillator. The oscillator **board plugs into the base; band changing** is **a** matter **of plugging** in another board.



getting excellent results with his home-brew transistorized one-watt vfo! W7IGV in Idaho with his own vfo design equals W6EAC's results. Others have built their own vfo's and encountered no problems. Crystal control is the logical first step; after you get your feet wet, see what you can do with a vfo.

Antenna tuning unit? No problem; make up a simple series or parallel arrangement such as W6TYP uses or use a conventional pi-network. The various Z-matches, transmatches and similar are just as effective with

fig. 7. Simple regenerative receiver used by W7GNT for portable operation. L and C values chosen for desired amateur band.



these little rigs as they are with their big brothers. Handle the output just as you would that from a more powerful tube transmitter.

What about modulation? Well, why not? The present trend is to concentrate on the rf angles; modulation is just around the corner. Of course, there is no reason why the little rigs can't be expanded to include modulation as well. Some efforts have been made along these lines; it's been reported that one W6 station has worked several hundred miles with voice on 160 meters with less than one watt input!

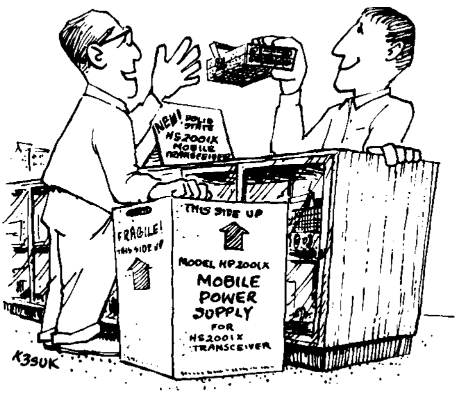
It's my guess that modulation experiments will be confined to a-m initially; there are too many hurdles to jump with ssb at this early stage. Take a lesson from the walkie-talkies and the increasing number of transistorized CB rigs; there are a lot of hidden tips there if you want to pursue this angle.

I have only scratched the surface of this exciting new field. I have shown you a few examples of accomplishment by Western hams because I am intimately familiar with their results. Undoubtedly you will find one or more hams in your own vicinity who are deeply engrossed in the art of QRP operation. Contact them and work with them; remember that many of the significant developments in the electronics field were ham conceived and ham developed.

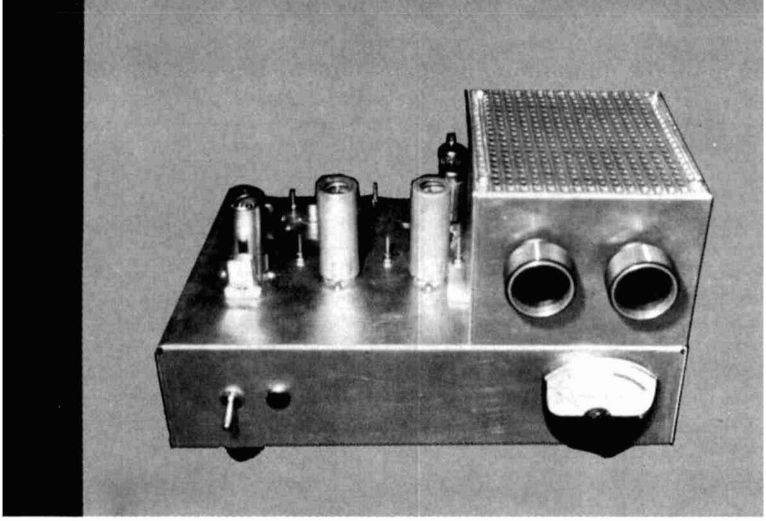
If you really want to keep abreast of what the QRP and QRPP gang are doing, the QRP Amateur Radio Club-International, with 3000 members in 57 countries, is the spot for you! Drop a postcard to Jim Loring, WA1BEB,* and ask for an information sheet and application blank to a club that offers a life membership for only $2.00 and no dues!

* Jim Loring, WA1BEB, General Secretary, QRP Amateur Radio Club--International, RFD 2, Gilead, Bethel, Maine 04217.

ham radio



"It's just loafing along at 2000 watts PEP . . . power supply is external of course . . ."



medium power six meter transverter

Construction details of a 14-MHz to 50-MHz transverter with greater than 150 watts output

Rick Davis, K8DOC, RD 2, Creek Road, Jefferson, Ohio 44047
Alex Dolgosh, K8EUR, 3905 Station Ave., Ashtabula, Ohio 44004

Looking for new horizons to conquer? With the predicted sunspot activity, six-meter DX may be more a reality than a phenomenon. Older vhf enthusiasts will recall the fine E-layer openings of 7957 to 1959 and the occasional F2 openings to both Europe and Asia. Now, with ssb facilities, you can enjoy the openings with longer and more reliable contacts than ever before.

Here is a low-cost, medium-power transmitting and receiving converter or transverter that is adaptable to most 14-MHz ssb stations, whether equipped with a separate transmitter and receiver, or a transceiver. Sufficient output power is available to drive a high-power grounded-grid linear amplifier more than adequately.

circuit

This circuit has evolved over a period of several years from the original W6RET design in the August 1965 issue of QST. It has been redesigned to provide greater power capabilities with improved receiving converter performance. In fact, it bears little resemblance to the original. Three models of this transverter are currently being used with good results.

The overtone crystal oscillator uses the triode section of a 6U8A, while the pentode section is used as an isolation amplifier to minimize loading effects on the oscillator and furnish sufficient drive to the control grid of the 5763 mixer. Drive is introduced in the screen circuit of the mixer through L6 and L7.

The 50-MHz output of the mixer is developed across L3 and coupled to the grid of the 5763 driver. The 10-ohm resistor loads the circuit sufficiently to insure stability. Cathode bias is obtained by the drop across a 15-volt, 1-watt zener diode, simplifying bias circuitry and providing automatic regulation. The diode is bypassed to prevent possible damage by rf energy.

A small metal shield should be installed across the driver tube socket to prevent coupling between the grid and plate circuits. A parasitic choke is used to further insure stability. The signal from the driver plate is coupled to the grid of the final amplifier with a capacitor. A parasitic choke is used in the final plate circuit for stability.

The bias for the 4X150A (or 4CX250B) is regulated at 47 volts by a 1-watt zener diode. A set of relay contacts breaks the bias network ground return on receiver, biasing both the driver and final beyond cut-off. In transmit, the bias drops to normal operating values. The 4X150A, operating in class AB_1, idles at about 50 mA.

The 4X150A screen by-pass capacitor shown is an integral part of the socket I used. However, less expensive sockets may be substituted if an external .001-μF bypass is installed at the screen terminal. The concentrically bypassed socket is not an absolute necessity at 50 MHz.

The plate tank is a conventional pi network. The 33-pF input capacitor should be spaced for the chosen plate voltage, but the output capacitor may be a close-spaced broadcast variety. No part numbers are given since these were salvaged from the junkbox. The coil dimensions were carefully selected using a wattmeter to determine the best L-C ratio. An Ohmite 2-50 choke is used as a safety measure to keep high voltage off the tank coil if the blocking capacitor fails in service.

Cooling air is provided by a Dayton 2C782 15-cfm blower. Adequate air flow through the heat-radiating anode fins of the final amplifier is a must. The chassis should be airtight. If you use another blower, greater air flow should be considered. I wouldn't use a fan under any conditions (as opposed to a centrifugal blower) since a fan cannot deliver adequate air against the back pressure created by the 4X150A.

receiving converter

The receiving converter circuit is a straightforward triode cascode rf amplifier coupled to a triode mixer. For simplicity, I used 6CW4 nuvistors throughout. The 50-MHz signal is coupled to the rf amplifier through a parallel-tuned 14-MHz trap which eliminates any 20-meter leakthrough. The rf amplifier plate/mixer grid tank is tuned to 50-MHz by L10; 36-MHz energy is coupled to the mixer grid from the 6U8A isolation amplifier.

construction

I won't attempt to provide exact carbon-copy construction details—this will be left to the resources of the individual builder. However, the Bud AC408 7x12x3 chassis and Bud AU1029 4x5x6 utility box used afford almost perfect sizing for good component layout. One variation was built with an 8x17x3 chassis to provide top mounting for the blower and rack mounting of the entire unit.

It's suggested that the transverter be constructed in stages. First, the crystal oscillator and isolation amplifier should be made operational, followed by the receiving converter. After this, the mixer, driver, and final amplifier stages should be completed and tested individually. Make sure the zener diodes are correctly installed—observe polarity. Good vhf construction practices should be followed throughout although there is no magic to making tuned circuits work well on six meters. A grid-dip oscillator should be used extensively at all points to avoid resonating tuned circuits on the incorrect sum, difference, or multiple frequency.

Only the minimum relay and control circuitry is shown since you will want to design your own controls to suit the needs of your station. It should be noted that one more set of relay contacts may be needed with some transceivers for receiving antenna switching.

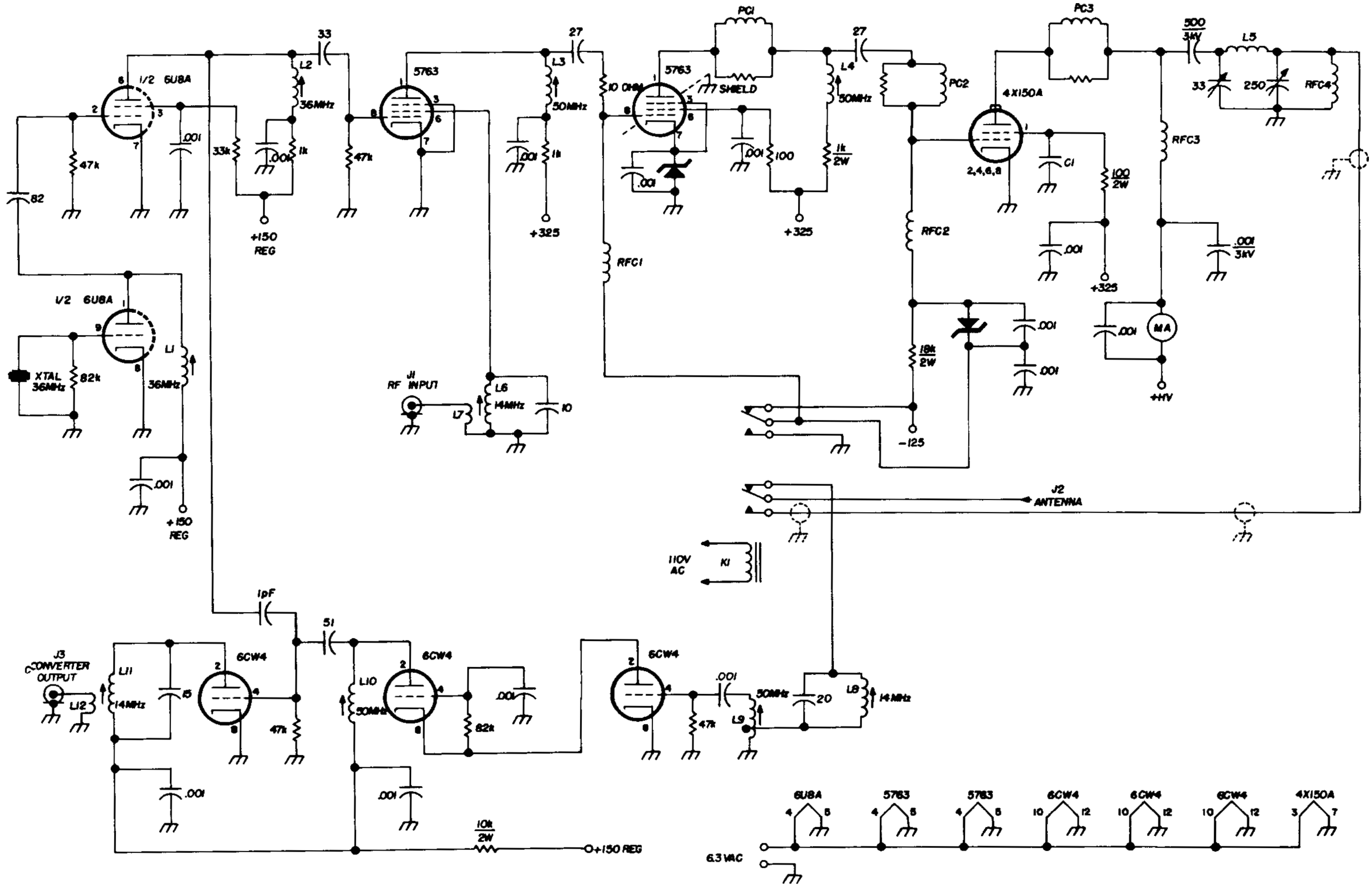


fig. 1. 150-watt six-meter transverter. There should be a .001-μF bypass capacitor from the lower end of L4 to ground.

D1 15-volt, 1-watt **zener** diode **(Mallory** ZA-15)

D2 47-volt, 1-watt zener diode **(Mallory** ZA-47)

J1, J2 SO-239 coaxial raceptacle

J3 RCA phono jack

K1 dpdt relay, 155-Vac coil (Potter and **Brumfield KRP-11AG,** removed from plastic case)

12 turns number-30 enameled on **¼"** slug-tuned form

15 turns number-30 enameled on **¼"** slug-tuned form

12 turns number-24 enameled on **¼"** slug-tuned form

10 turns number-24 enameled on ¼" slug-tuned form

5 turns number-10 space wound, **1"** diameter, 1" long

L6, **L8, L11** 40 turns number-30 enameled on ¼" slug-tuned form

L7 4 turns number-22 hookup wire wound around cold end of **L6**

L9, L10 11 turns number-24 enameled on **¼"** slug-tuned form. **L9** tapped at 2 turns from cold end

L12 **4** turns number-22 hookup wire wound around cold end of **L11**

PC1, PC2 4 turns number-24 enameled wound on **47-**ohm, ½-watt resistor

PC3 4 turns number-14 enameled wound on **47-**ohm, 2-watt resistor

RFC's **7-µH** rf choke **(Ohmite** 2-50)

Use Eimac SK-606 chimney or equivalent for **proper** air ducting.

alignment

If you check out the transverter one stage at a time, you shouldn't run into any difficulties. Before final **tuneup,** resonate all coils to the desired frequencies with a grid-dip meter. It's possible to tune the whole works up on 36 MHz if you're not careful. All receiving converter coils except the 14-MHz trap should be peaked for maximum received signal strength. The 14-MHz trap is adjusted for minimum 20-meter leakthrough.

All slug-tuned coils in the transmitting section should be peaked for maximum plate current indication with 14-MHz drive applied to the mixer. Needless to say, the final amplifier plate current should not rise above the idling value unless 14-MHz drive is applied to the mixer. The final is tuned up in the normal manner. About 250 mA of plate current is optimum under full carrier.

It's best to use an swr bridge or wattmeter and tune for maximum output with the key down. When driven by a ssb signal, the indicated peak plate current should not exceed 50% of the key down value. Since only a small amount of 14-MHz drive is required, a pad will be necessary to attenuate the output of most exciters.

Neutralization has been found to be unnecessary on any of the three transverters now in use, even at plate potentials as high

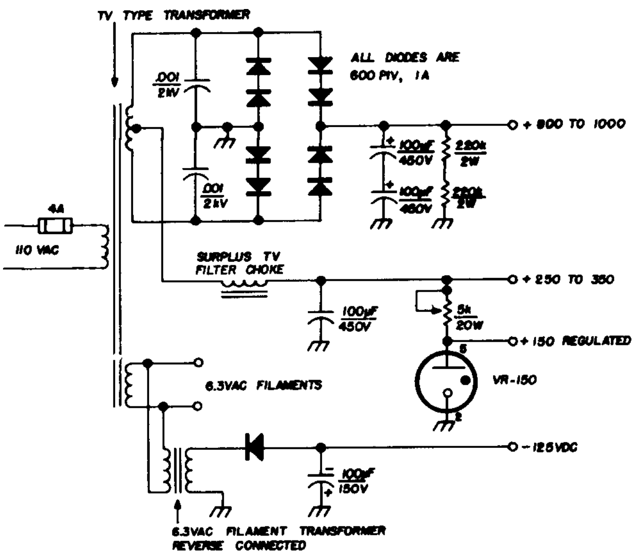


fig. 2. Power supply for the **six-**meter transverter uses a transformer from an old television **set.**

as 1400 volts. This is attributed to final amplifier shielding and complete isolation between grid and plate tuned circuits. The transverter may be run at full key down input power on CW, and may even be operated on am at reduced power and efficiency.

power supply

Power supply requirements may be met most economically by using an old TV power transformer with a 6.3-volt filament winding and a center-tapped secondary winding of 600 to 1000 volts. A bridge rectifier and capacitive filter circuit is used for the high voltage. With this circuit, the average TV transformer will provide about 800 to 900 volts under load. The low voltage for the mixer and driver plates and final amplifier screen is taken from the center tap of the high voltage winding.

A 5000-ohm, 20-watt adjustable resistor is used to drop this voltage to the proper value for the VR-150 regulator. The receiving converter, overtone crystal oscillator and isolation amplifier are also powered by this regulated voltage. Bias voltage is obtained by reverse connecting a small 6.3 volt filament transformer to the filament winding of the TV transformer and rectifying the output.

It's suggested that the 110-volt ac line be brought to the transverter chassis; install an outlet on the rear for the power supply. This way, 110 volts will be available for the blower, and the power supply may be located remotely. Although the filament voltage is fractionally higher than the recommended 6.0 volts, no shortening of tube life has been experienced.

operation

This transverter makes an excellent driver for a high-power grounded-grid linear amplifier. Output in excess of 150 watts may be expected, so it also provides a very substantial barefoot signal. In addition, this simple design will put six-meter operation within reach of most amateurs without the half-kilobuck expense of a commercial transceiver and power supply.

ham radio

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 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-wavelength 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. 1A**). This condition is easily remedied by the rather obvious technique of building a larger diameter coil whose

table 1. Television and fm frequencies which can be used for calibrating a grid-dip meter.

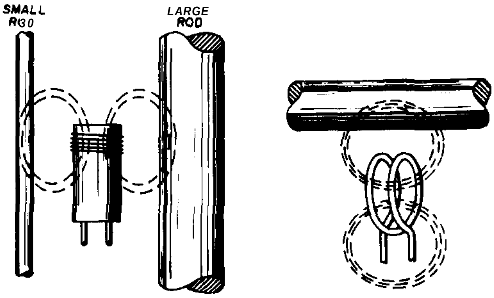
channel	video	sound	video —4.5 MHz	video second harmonic	sound second harmonic
2	55.25	59.75	50.75	27.6	29.9
4	67.25	71.75		33.6	35.9
5	77.25	81.75		38.6	40.9
fm	88-108			44 to 54 MHz	

larger field pattern can be coupled to larger diameter rods (see **fig. 1B**).

antenna gdo

One method of solving the problem is shown in **fig. 2.** 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 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.

fig. 1. 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.



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 1.** 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

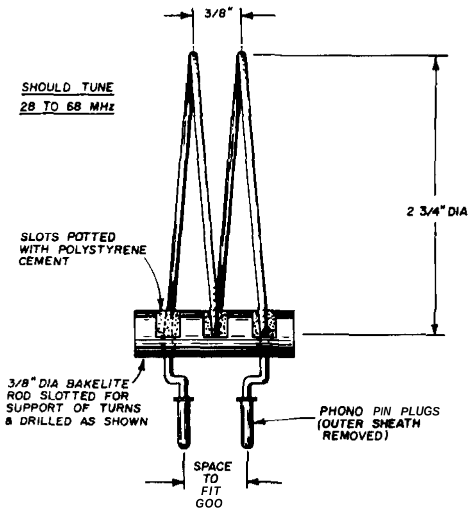
Let's try an example. A good three-element beam is the one with 0.15-wavelength spacing from driven element to reflector; the reflector is approximately 1.06 times the length of the driven element. The director is 0.943 the length of the driven element and spaced 0.1 wavelength away.

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 X 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.

fig. 2. Method of building a large coil for your grid-dipper.



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

time-to-reply statistics for dx qsl's

Now that you have worked that rare one, what's the best approach for getting a QSL?

Lee Vogel, WB6IUH, 1048 W. Hill Avenue, Fullerton, California 92633

How long should you expect to wait for that QSL card you need for DXCC? What are your chances for receiving a QSL in a month? . . . in six months? . . . in a year after the QSO? How much will you improve the chance of eventually receiving the card if you send IRC's? How about an addressed envelop—complete with foreign stamps?

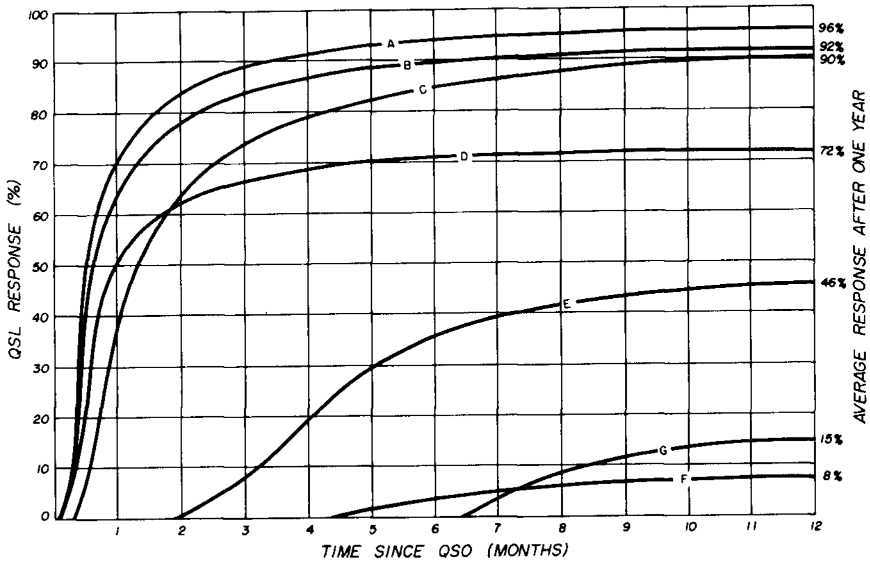
This article answers these questions and is based on statistics gathered during the past few years. During this time I have tried all of the various QSL'ing methods proposed in previous articles and kept an account of their success. There is nothing magic about a WB6 call so you should realize at least the same results. If you are rare for the DX station then you can expect to do better. In any case, the relative merits of various QSL'ing procedures should still apply.

The QSL courtesy is recognized throughout the world, and a number of articles have been published on the art and ethics of QSL'ing.[1,2,3,4,5] Most authors discuss the problems faced by the rare DX station and describe techniques which will insure the QSL by alleviating his load.[6,7,8] However, I know of no published data available on the effectiveness of the various methods.

This article presents statistics gathered over a three-year period from 1500 DX QSO's. Several methods of requesting a QSL are compared including: direct request with IRC's enclosed, direct request with a self-addressed envelope and foreign stamps from the DX country, direct QSL card request only, request via the bureau (by air and surface mail) and request through a QSL manager.

the problem

After an exciting QSO with a rare DX station you are immediately faced with how best to obtain a confirming QSL. The most economical choice is clear if the station has a QSL manager; check all of the latest lists.* However, if there is no manager then you have several options. You can QSL via the foreign bureau and wait very patiently; send



Curve A Response to QSL request accompanied by self-addressed stamped envelope

Curve B Response to QSL with IRC's enclosed

Curve C Response via QSL manager when self-addressed stamped envelope supplied

Curve D Direct reply to a routine direct request with your card

Curve E QSLs received via the bureau before the DX station has heard from you

Curve F Reply via the bureau to a direct request

Curve G Bureau replies to your request via the DX bureau

fig. 1. Graph of time-to-reply statistics for various methods of QSL'ing. Curves F and G eventually reach 67%.

The average time required to receive an answer to each of these methods is compared as well as the over-all percentage reply.

Since I am a professional statistician, it was more or less second nature for me to keep complete DX contact data in my log. In addition to the entries required by law, I also record the QSL method I use and the time required to receive a reply. A study of this data over the past few years clearly demonstrates the timesaving which results from the proper QSL'ing procedure. Let's take a look at my QSL'ing experience.

your card directly to him hoping he'll be a good guy and reply in kind; send your QSL card with IRC's to help pay postage or obtain stamps from his country and send a self-addressed, stamped envelope with your QSL card—possibly enclosing a photograph and personal note in his language!

During the last three years I have tried all of these methods. Let's look at three ways of comparing QSL methods; you can decide which is most important.

1. The average elapsed time between QSO and receipt of QSL

2. The elapsed time to 50:50 chance of receiving a given card; the time half of the cards will arrive

3. The over-all percentage response one year from the QSO

These answers are summarized in fig. 1 which gives the percentage response versus time for each QSL'ing procedure. The curves are identified as follows:

Curve A Response to QSL request accompanied by a self-addressed stamped envelope.

The best and fastest method on all counts is to supply the DX station with a completely prepared envelope! The average time for receipt of a QSL was only about 5 weeks; there is a 50:50 chance of receiving a given card in only two weeks. Better than 95% replies were received within a year.

Curve B Response to QSL with IRC's enclosed.

This is the second best QSL'ing method by all accounts. The correct number of IRC's varies from country to country and is given in the **Radio Amateur Callbook.** The average time to receipt was only $5^1/_4$ weeks with a 50:50 chance of receiving the card in three weeks. Over-all response within a year was a creditable 92%.

Curve C Response via a QSL manager when you furnish a self-addressed stamped envelope.

This was the next best method. Generally it is the cheapest since many DX stations have stateside managers. Average time to respond was $8^1/_2$ weeks with a 50:50 chance for receiving a card in about 5 weeks. About 90% of the cards were received within a year of the QSO date.

Curve D Direct reply to a routine direct request with your card.

This method rated next. The average time to respond (for those who eventually did!) was $6^1/_2$ weeks with a 50:50 chance for a reply in 5 weeks. However, there was only 72% response within a year after the QSO. For this reason the effective average time will be longer to account for this increase in failure to reply. A side note: Sending your card via air mail will, on the average, save about one and one-half weeks.

replies via qsl bureaus

The remaining curves, marked E, F and G, represent results for cards sent via the bureau. When the address of a DX station is not known you can only send your card to the QSL bureau in his country. Addresses for these bureaus are listed in QST periodically as well as the foreign edition of the **Callbook.** Your card is forwarded from the bureau usually to club stations for eventual delivery to the amateur himself.

Returning cards are sent by the foreign bureau to ARRL headquarters where they are bundled and forwarded to the QSL bureaus in each W/K call area. Cards received locally are sorted and periodically forwarded in self-addressed stamped envelopes furnished by the amateur. The average delay in the local QSL bureau is often a month or more, depending upon the voluntary club members who run the bureau.

In general, the response through the bureau is slower and not as reliable as the direct methods are. You can expect to receive only

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Name .... ***Call*** ....
Address ....
City .... ***State*** .... ***Zip*** ....

Name .... ***Call*** ....
Address ....
City .... ***State*** .... ***Zip*** ....

Name .... ***Call*** ....
Address ....
City .... ***State*** .... ***Zip*** ....

Name .... ***Call*** ....
Address ....
City .... ***State*** .... ***Zip*** ....

If you need more space please attach a separate list.

Your Name .... ***Call*** ....
Address ....
City .... ***State*** .... ***Zip*** ....

This special offer expires December 24, 1968

about two out of every three QSL's within a year of the QSO. By then the response rate is practically nil. Therefore the percentage of replies will not increase much in another year or so. If you haven't heard within a year, you had better send a second card with special inducement if you want the card!

Curve **E** QSL's sent to you via the bureau before the DX station heard from you.

In countries with outgoing bureaus, the majority of amateurs send their card via the bureau before they receive yours. Usually the card is marked "pse QSL" to remind you that he would like to receive your card. Curve E presents results tabulated for these replies. Thus it represents the one-way time via outgoing bureaus. The average time to reply (assuming he replies within a year) is over 22 weeks. A 50:50 chance of reply occurs at about 15 months after the QSO with only about 46% reply response after one year. Note that it takes almost 2 months before the first cards begin to arrive.

Curve F Reply via the bureau to a direct request on your card.

This represents those DX stations who wait for your card and then send their card to you through the outgoing bureau. The first reply of this type arrives about 4½ months after the QSO. After a year, about 8% of the overall 67% bureau response will reply in this fashion. Note that sending your card by air mail will result in an average saving of 3 weeks for these replies.

Curve **G** Bureau replies to your request via the DX bureau.

This last category represents the slowest method — both amateurs using the DX bureau. The first cards begin to arrive after about 6½ months. About 15% out of the 67% bureau responses will be received this way a year after the QSO. While these last two curves are not as flat after a year as the other methods, it is obvious that replies will not increase rapidly thereafter.

final comments

The curves in fig. **1** can be used also to estimate your chances for receiving a card at various times after a QSO. Simply enter the graph at the time of interest and move up vertically until you intersect the curve applying to the method you used. Read the percentage after this time on the left-hand axis.

These statistics graphically demonstrate the payoff for various QSL'ing options. This data should help you to decide when the added expense is worth the savings in time to reply. Perhaps even more important, how the added effort on your part increases the chances of receiving an important card.

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4. T. Hannah, "QSL Via Box 88," *QST,* September, 7967, p. 77.
5. I. Kazaisij, "On Sports Ethics for Short-Wavers," Radio (USSR), Volume II, February, 1967.
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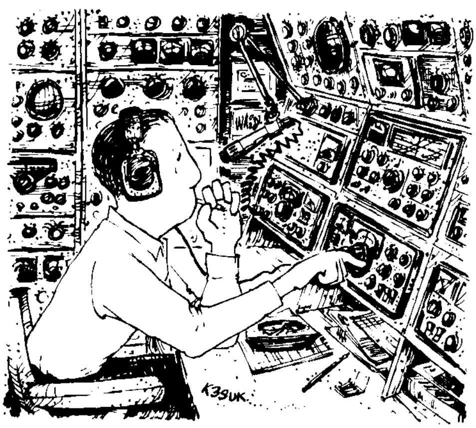
8. T. Wilds, "On the Art of QSL'ing," *QST,* October, 1963, p. 61.

* W6GSV's *QSL* Managers' Directory for example. $3.95 from Bookbinder Publishing Company, Box 54222, Terminal Annex, Los Angeles, California 90054.

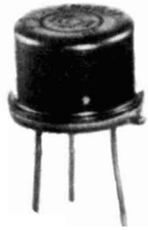
Foreign postage is available from the DX Stamp Service, Webster, New York 14580, run by W2SAW. If you'll write to him, he'll send you a complete price list.

ham radio



". . . OSS . . . OSS . . . OSS
I wonder what that means?"

State of the Art for '69 ?



YES !

The hottest thing to come along in years for high performance receiver design is the MOSFET (Metal Oxide Silicon Field Effect Transistor). New types–only recently available for general use–are capable of equalling or surpassing the performance of the best popular vacuum tubes in critical front end applications.

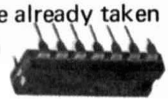
Properly used, they finally lick the old transistor hang-up of achieving low noise plus extremely wide dynamic range. (Without the ability to handle very large as well as very small signals, your receiver is susceptible to cross-modulation interference from strong off-channel stations, and to distortion of strong on-channel stations due to inadequate AGC range.)

Certain of the newer **MOSFET's** combine the **large-signal** linearity necessary to avoid **cross-modulation** and the **ability** to be **gain-controlled** over a wide **range** without compromising that linearity–no more auxiliary RF attenuators to supplement marginal AGC!

Transistors are still pretty hard to beat for low voltage, relatively high dissipation applications, too . . . like power amplifiers and voltage regulators.

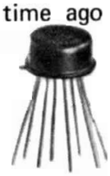
. . . AND NO

Digital integrated circuits have already taken the computer field away from transistors, and they're now revolutionizing instrumentation as well.



Digital frequency counters using IC's, for example, do the job of much larger transistor units at a fraction of the cost.

Intensive development of Linear **IC** technology has paid off more recently. A short time ago linear IC's were very limited in performance, expensive, hard to get. Now they're readily **available** in a multitude of varieties . . . and inexpensive. . . doing for applications like I.F., audio, and control amplification what the digitals did for computers.



Linear or digital. an IC does the job **of** several transistors plus the diodes, resistors, and interconnections that make them into a complete ready-to-use functional block. Be it a complex JK flip-flop or a high-gain I.F. amplifier, the **IC** often costs less than just the resistors of the conventional circuitry it replaces! (Equipment reliability benefits, **too . . .** because of all the separate components and soldered joints that are **eliminated.**)

The POINT is, modern component technology has made available a remarkable variety of sophisticated devices.

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ramp generators

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

"It looks like a sawtooth oscillator to me, Charlie. I wonder why this guy calls it a ramp generator?' Actually the question of whether a wave form like that shown in **fig. 1** is a ramp or a sawtooth is merely a matter of word choice. Some might argue that "ramp" is the more general description since it does not infer a sequence of waveforms like that between "0" and "t" in the example. But the sawtooth advocate could claim that the word "sawtooth" is singular, too, and possibly claim prior use historically. In this discussion I will refer to waveforms like that in **fig. 1** as ramps (choosing this term for brevity only) and hopefully will not be accused of taking sides in the semantics battle.

Ramp generators probably were first designed for horizontal sweep use in cathode-ray oscilloscopes. The ramp-generator circuit used in most of the early oscilloscopes was the thyratron relaxation oscillator. This circuit is really just a modification of the old neon-lamp relaxation oscillator shown in **fig.** 2. The neon lamp circuit, used for a novelty type flasher, has probably been built by every electronic experimenter alive. **Fig.** 3 shows how a thyratron (a neon tube with a grid) is substituted in the circuit. The grid allows adjustment of the firing voltage of the tube so t_1 can be varied by adjusting grid bias. Of course, sync pulses can be easily injected at the control grid.

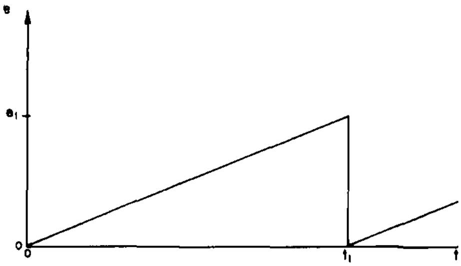
You'll notice in **fig. 2** and **3** that the waveform is not a linear ramp as desired, but rather a portion of an exponential curve—the curve followed by the voltage of a capacitor charging through a resistor. To make the waveform of **fig.** 3 more nearly linear, it's possible to return the top of R to a voltage higher than E_{BB}, which better approximates **constant-current charging.**

This constant-current charging is an important concept in understanding ramp generation. For example, suppose you have a black box with a one-million-volt battery inside it. In series with the battery is a 10-megohm resistor, also inside the box. The resistor and the common side of the battery are connected to two terminals in the side of the box as in **fig. 4.** On the outside of the box a 1000-ohm variable resistor is connected across the terminals.

An observer is invited to adjust the variable resistor and measure the voltage across it while doing so. He finds that the voltage across the variable resistor is proportional to its resistance since the current flow in the whole circuit will hardly change at all, percentage-wise, as R changes.

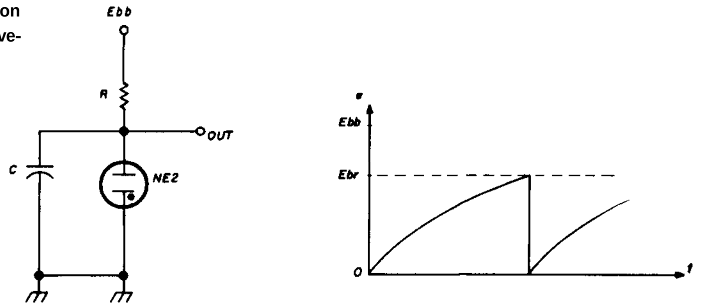
The current through R will be approximately 100 milliamps regardless of the setting of R. The observer would say, "In this black box is a constant-current source

fig. 1. A ramp waveform.



of 100 mA." I'd like to point out that this is an illustrative example only and not for actual construction. (If you do build it, for goodness sake don't ever disconnect the resistor!)

fig. 2. Neon-lamp relaxation oscillator and its output waveform.



The constant-current generator can be conveniently thought of as a large voltage in series with a large resistance, at least for a mental handle. In actual practice, more realizable approximations to constant-current generators are used in circuits. One such approximation is the pentode vacuum tube. If a pentode is substituted for the charging resistor, as shown in **fig. 5,** near-linear ramps are obtained. This method was used in some of the early ARRL Handbooks in their oscilloscope sweep circuits but dropped in later editions.[1]

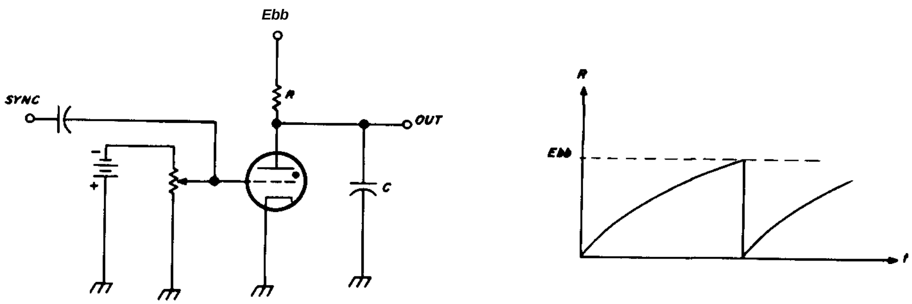


fig. 3. Thyratron ramp generator.

By now you should realize that the neon or thyratron is nothing more than a voltage-sensitive switch. The fact that the switch is sensitive to voltage means that ramp generators are self-resetting; that is, they put out ramp after ramp in a continuous sawtooth-like wave train.

By substituting a triode for the thyratron, it's possible to change the free-running ramp generator to a driven ramp generator. In **fig.** 6, when V_1 is biased "off," the ramp grows linearly with time. But at t_1, V_1 is made to conduct by the positive pulse applied to its grid, and C is discharged. This "triggered sweep" is the type of ramp generator that is used in most modern oscilloscopes.

The bootstrap circuit (so called because it "pulls itself up by its own bootstraps") is shown in **fig.** 7. This circuit is very similar to that of **fig.** 3 except that a cathode-follower has been added that couples the output ramp voltage back to the charging resistor. Therefore, the voltage which operates the RC circuit rises as charging proceeds. This is an effective method of obtaining constant-current charging and yields very nearly linear ramps.

The last variety of tube-type ramp-generator to be discussed here is the Miller integrator. The sanatron and phantastron circuits are variations of the

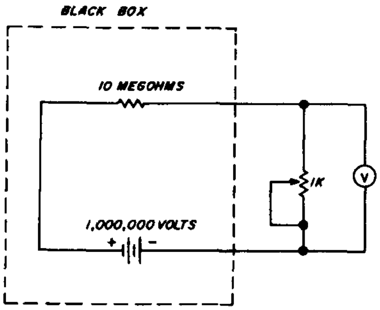


fig. 4. Illustrative example of a constant-current generator.

Miller integrator and will not be described in detail. The Miller integrator is also similar to fig. 3, except that C is replaced by the Miller-effect capacitance —one of the facts of life that is undesirable in building rf amplifiers; it is the apparent input capacitance "seen" from grid to cathode due to grid-to-plate capacitance (C_{gp}).

The fact that C_{gp} is connected between the input and the amplified and out-of-phase output of the amplifier makes it appear much larger than C_{gp}. In fact, the apparent capacity to the cathode is approximately μ x C_{gp} the voltage gain of the tube times the grid-to-plate capacitance! Since the μ of a pentode (the type of tube usually used in Miller integrators) is often 1000 or more, the effective C can be quite large. Is it any wonder that neutralization is so important in rf amplifiers?

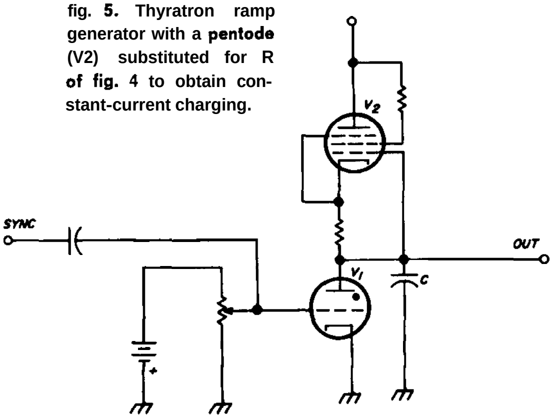


fig. 5. Thyratron ramp generator with a pentode (V2) substituted for R of fig. 4 to obtain constant-current charging.

By adding a lumped capacitance across the grid and plate, we can make μC_{gp} really of useable size. Then a very large capacity is charged to a small voltage (at the grid) so it stays on the nearly linear portion of the exponential charging curve near zero voltage. The amplification of V_2, however, provides a larger useable ramp. It is possible to analyze the Miller integrator in other ways and even to show that it is equivalent to the bootstrap circuit, as shown in reference 2.

It may seem that a great deal of space has been used in describing "obsolete" vacuum-tube circuitry. But we will see that these tube circuits form the basis of many of the newer solid-state circuits. In fact, now that FET's are coming into wide use, many of the old tube circuits can be used directly.

A simple solid-state circuit that is very much like the neon-lamp relaxation oscillator of fig. 2 is shown in fig. 7. The diode is a Shockley four-layer type which displays a negative resistance on breakover.

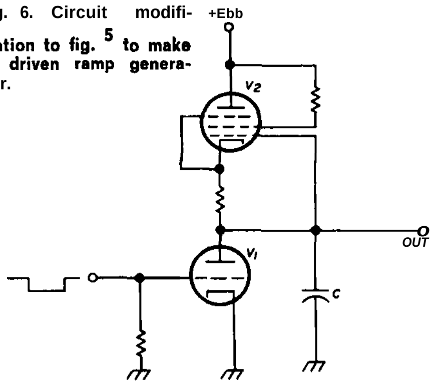


fig. 6. Circuit modification to fig. 5 to make a driven ramp generator.

Fig. 10 shows how a unijunction transistor can be used as a relaxation oscillator. The breakover voltage can be controlled by choosing the zener voltage, since the ujt breakdown is always a constant fraction of the voltage between B_1 and B_2. This circuit can be modified (much as fig. 3 was modified to fig. 5) for constant-current charging by replacing R with a bipolar transistor, an FET, or a field-effect diode as shown in **fig. 11.** An example of a solid-state bootstrap circuit is shown in **fig.** 12. This particular bootstrap circuit is self-resetting. It is described in more detail in reference 3.

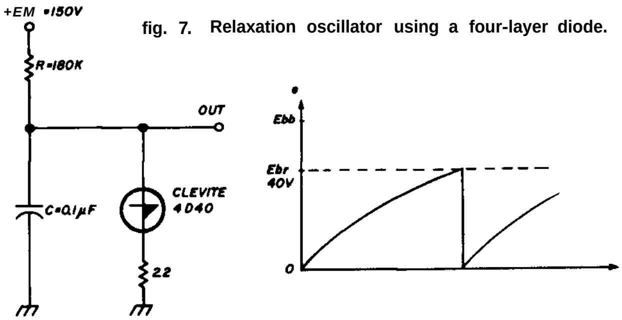


fig. 7. Relaxation oscillator using a four-layer diode.

fig. 8. The bootstrap ramp generator. The ramp waveform starts when the switch tube (V1) is biased off; V2 is a cathode-follower that controls feedback.

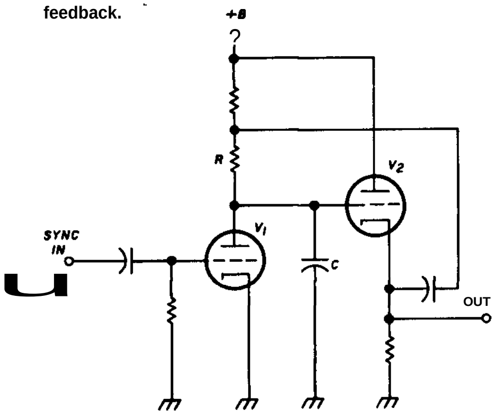
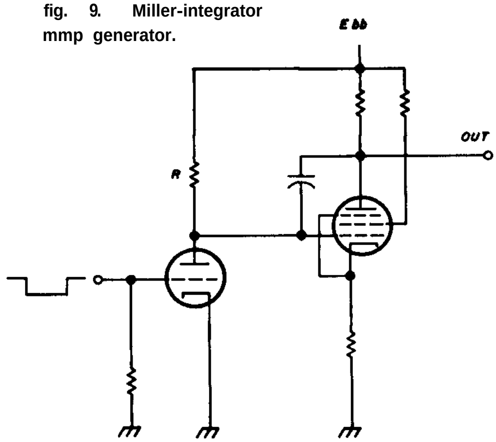


fig. 9. Miller-integrator mmp generator.



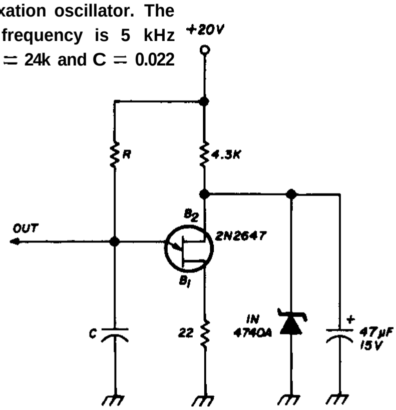
The Miller integrator is easily constructed using an FET as the amplifier stage, as shown in **fig. 13.** The Miller integrator is only one example of this general class of feedback ramp generator; as mentioned before, the bootstrap is a similar circuit. You will find that the amplifier tube or FET may be replaced with nearly any high-input-impedance, low-output-impedance, inverting, voltage amplifier.

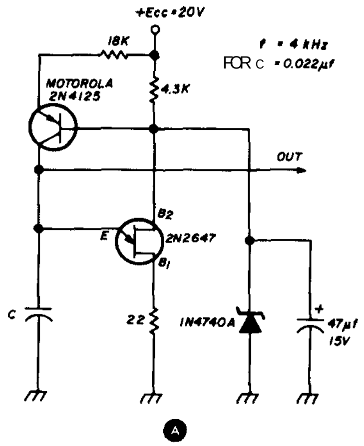
Such a device is the so-called operational amplifier that is available in packaged form from at least half a dozen firms. Operational amplifiers were originally designed for use in analog computers but now find wide use in other electronic systems.

Wider usage of operational amplifiers, and the tremendous potential price cuts promised by monolithic integrated operational amplifiers, should make the feedback method of ramp generation more popular in the future. The higher the voltage gain of the op-amp, the more nearly the ramp approaches perfect linearity, providing the input-output phase can be made to stay at 180°.

The uses of the ramp generator are many. Every oscilloscope uses a ramp generator for horizontal sweep and so do tv cameras and tv receivers and panoramic receivers (spectrum analyzers). By applying the ramp voltage to a voltage-controlled oscillator (VCO) a swept frequency may be produced. Such voltage-

fig. 10. Unijunction-transistor relaxation oscillator. The output frequency is 5 kHz when R = 24k and C = 0.022 μF.





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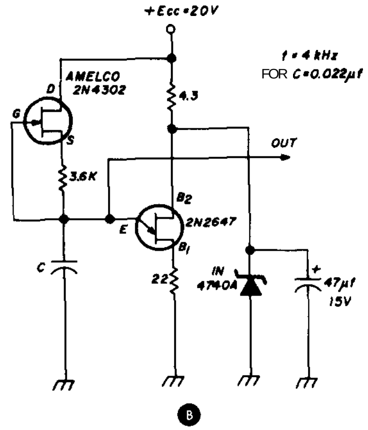
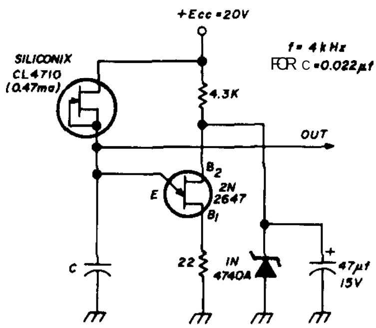


B

fig. 11. Modifying the relaxation oscillator shown in fig. 10 for constant charging with a bipolar transistor (A), an FET (B) and field-effect diode (C).



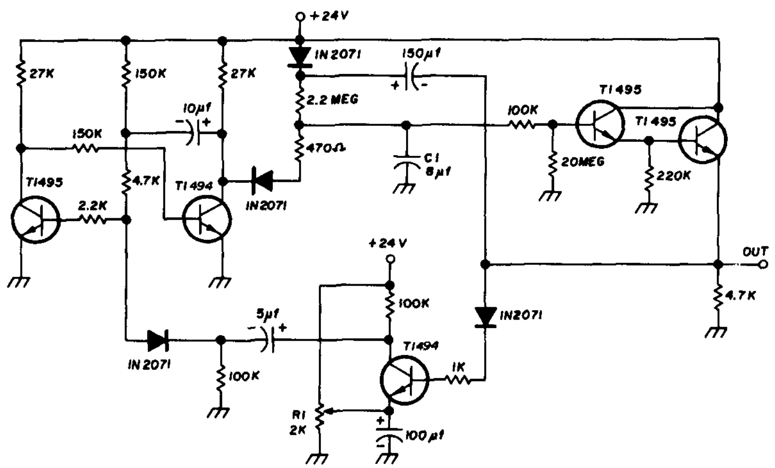
controlled frequency sweeps are used in sweep generators and spectrum analyzers.

The ramp generator is the basis for most electronic timers and similarly can be used to build a pulse generator with pulse length as a function of voltage. This is shown in block diagram in fig. **14** and an actual circuit is shown in fig. **15.**

The testing of components at various voltages can be a tedious job if done

fig. 12. Bootstrap ramp generator. As **C1** starts to charge through the **2.2M** resistor, the ramp voltage is coupled out **through the two T1495 emitter-follower output stages. The ouput** ramp is coupled back to the top of the **2.2M** resistor through a **150-μF** capacitor, reverse biasing the **1N2071** and allowing the voltage operating the resistor to exceed 24 volts.

NOTE: C1 IS MYLAR AND NON-POLARIZED. ALL RESISTORS ARE 1/2W 1%. TIME CYCLE IS SET BY R1. TIME CYCLE FOR CIRCUIT SHOWN IS 2-20 SEC ± 5% IF AMBIENT TEMPERATURE RANGE IS -20°C TO +65°C.



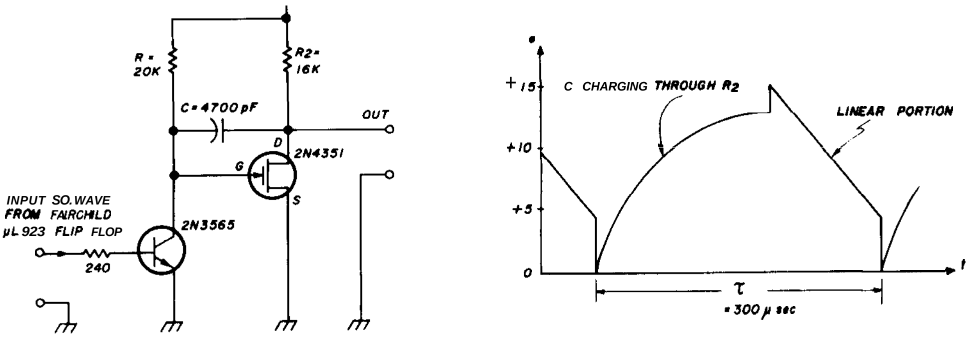
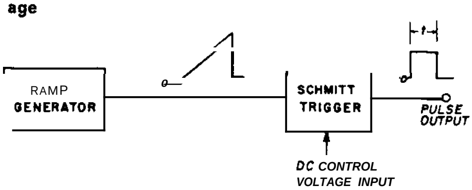


fig. 13. Miller-integrator ramp generator using an enhancement-mode MOSFET.

manually. By applying a ramp of voltage to the device under test while monitoring the performance on an oscilloscope with an identical sweep, the testing can be done automatically and recorded photographically. This is the sort of thing done in transistor curve tracers.

fig. 14. A voltage-controlled pulser where the pulse width of the pulse train is controlled by a dc voltage



There are many more uses for the ramp generator; your imagination and the linearity of various voltage-to-whatever transducers are the only limitations. I want to thank Texas Instruments for permission to use fig. 12.

references

1. ARRL, "The Radio Amateur's Handbook," 1945 Edition, p. 428.
2 M.I.T. Radiation Lab Series, "Waveforms," Boston Technical Lithographers, Inc., p. 35.
3. Texas Instruments, "Transistor Circuit Design," McGraw-Hill, New York, 1963, p. 415.
4. J. Pettit, "Electronic Switching, Timing, and Pulse," McGraw-Hill, New York, 1959, Ch. 6.

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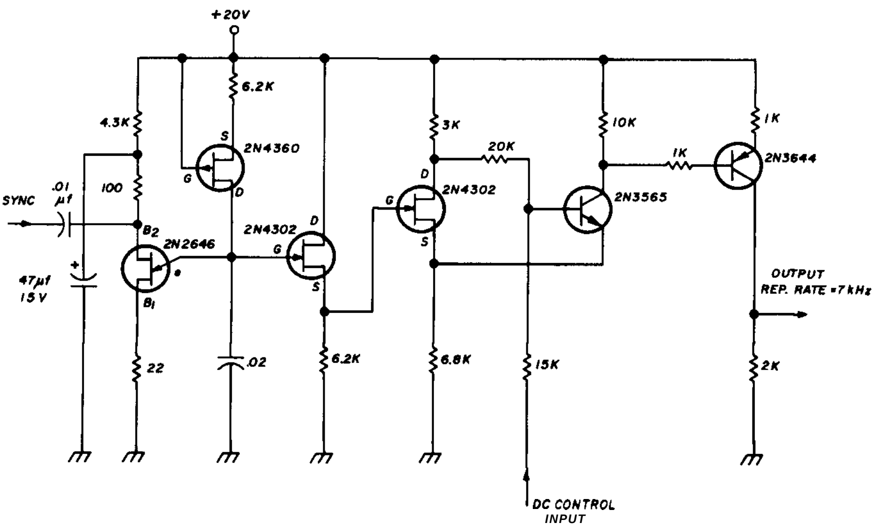
fig. 15. Actual circuit of the block diagram shown in fig. 14.



silver plating for the serious amateur

make your own lab for silver plating vhf components; it's easy and inexpensive

Ralph W. Campbell, W4KAE, 316 Mariemo t Drive, Lexi gton, Kentucky 40505

The advantages of using plating on electronic components are well known. At W4KAE I've found that vhf tank circuit efficiency improved considerably after silver plating. Silver oxide on critical connections such as mike connectors and antenna conducting bolts pays off in increased reliability. Transmitting tube pins, uhf cavities, and klystron connections all benefit from silver replating. The movable parts in tuned cavities in fm transmitters, for example, frequently break down because of worn plating on metallic interfaces.

Successful silver plating has been a problem ever since I started to build amateur equipment for 2 meters. Of all the popular methods tried, brush plating (described below) was closest to yielding acceptable results, even though it was too slow for amateur purposes. This kind of plating, I thought, was also more potentially dangerous since it used cyanide salts in a water-soluble base.

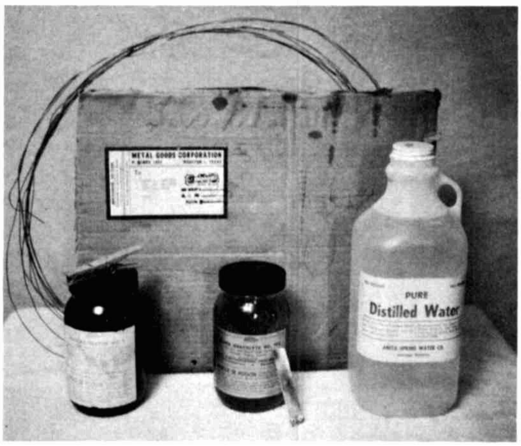
In this article, silver plating is shown to be a practical means for improving conductors and conductive surfaces. And the method can be used with safety by following the rules printed by the manufacturer on containers. An equipment description and a step-by-step set of instructions are presented. For those unaccustomed to plating terms, a glossary of by-word-of-mouth shop-talk definitions appears at the end of the article.

plating methods

There are three basic methods for depositing metal onto metal: brushing, tank plating, and immersion. The first two are true electroplating methods. The third, also known as electroless plating, is an ionic displacement method where the plating solution is heated and electricity is not used.

Brush plating is probably the simplest to use. The work to be plated (called the cathode) is brushed repeatedly with the anode until the desired plating buildup is obtained. The anode, which is generally made of the metal to be deposited, is wrapped with a gauze-covered wad of cotton. Before each brushing this is dipped into a solution containing cyanide salts: of the metal to be deposited. Both cathode and anode are connected to a dc source to complete the electroplating circuit. Voltages between **1.5** and **6** Vdc are normally used for silver plating.

fig. 1. The materials needed for high-quality silver plating.



Tank plating, which is the most practical method, is the one described here. Experiments have shown that tank plating is probably best for most amateur work. If you have enough patience and time, however, brush plating would be the logical choice for use on circuit boards, which have discontinuous conducting elements and are thus difficult to plate using the tank method.'

what's available

Having had some success with the brush-plating method I wrote the people who made these kits to see if there was anything better available.* There was. Kits of chemicals (with plater) and metallic anodes could be supplied at reasonable prices over most other sources. For industry they offer their "Rapid-Plater." For amateur, design-lab and experimental use, small amounts of plating supplies can be obtained. My initial inquiry resulted in quick response plus their latest catalog.

As it turned out, many different compounds and metals were available. To gain experience, I tried copper plating first, then bright nickel and "Rapidmetal;" then silver. Electrolytes other than the cyanide salts in solution were tried, but they had almost no throwing power.

Organic compounds have apparently been added to Rapid's electrolyte number **316** to prevent formation of hydrogen cyanide. (This stuff, incidentally, is the same gas used in prison death chambers, so it's not hard to see why the manufacturer has taken pains to inhibit its formation).

For my tank plating, I used the following chemicals and materials: Coatalyte number **316** electrolyte, silverizing activator number 4 and number **536** without sleeve. These form the basic elements of a simple but effective homemade plating lab.

equipment

Fig. 1 shows most of the apparatus needed for professional quality silver-plating. Starting at the left, a bottle of Rapid activator number 4 is shown. This is used for preparing copper, brass and bronze for silver and gold plating. No water rinse is needed with this activator. The chemical action of the activator is known as "striking or silverizing" nonferrous base metals for high-integrity deposition of silver.

* Rapid Electroplating Process, Inc., 1414 S. Wabash Ave., Chicago, Illinois 60605.

On top of the activator bottle is a wooden clothespin — the single most handy device I used. Next is electrolyte number 316, with a piece of number 536 pure silver anode resting on the container; then a half-gallon of distilled water.

In tank plating, the Coatalyte should be used full strength. Dilution by more than 10 percent will destroy the throwing power of the electrolyte for future use. But if you must dilute, don't attempt to use the weakened solution again for smaller work.

I bought two bottles for this reason. One jar is used for large surfaces (using meticulous care and agitation), and the remaining jar is kept at full strength. Only undiluted Coatalyte will have enough throwing power to deposit 1 mil total plating thickness on piecework in fifteen minutes. Then the work must be stripped down for final or additional coats.

Behind the jars is some stainless steel wire which is used to connect the piecework and anode to the voltage source. Stainless steel is preferred because it minimizes buildup of ferro-cyanide, which contaminates the solution. Loose particles of foreign material floating in the electrolyte tend to cause roughness in the deposition of silver onto the work.[1] Ordinary pure copper wire can be used; however, number 18 stainless can be obtained COD on a "minimum billing" from the Metal Goods Corporation for less than $5.00.*

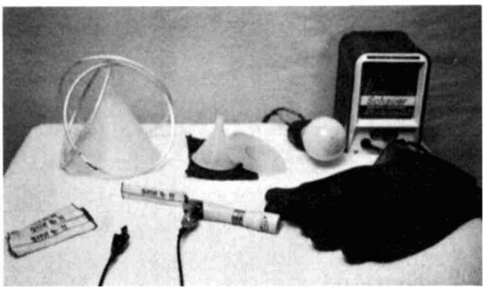
Fig. 2 shows several acrylonitrile number 31 sleeves used to cover boot-shaped anodes (for Rapid or "sleeve plating"), and in the background is a nylon funnel with three rings which I silver plated. To the front is an applicator (anode + sleeve + handle) and two heavy-duty Mueller clips which are used for connections to the dry cell or battery charger. Also, there is a smaller nylon funnel shown with a polyethylene lid, and a homemade separator, resting on a piece of Scotch-brite abrasive cleaning pad.

Fig. 3 shows an in-tank view of my plating method. Only 1 1/2 volts are used from a number 6 dry cell. Normally, the concentrated electrolyte is kept in its jar or in a roll-film tank. Cost of silver solutions is typically 54.80 per 16-oz container of number 316. For this reason, small containers are preferred. Stripping the work for successive coats should be done every fifteen minutes. Lightly sanding with crocus cloth is preferable. Rapid Electroplating Process "portable platers" use sleeves which negate the chore of stripping at regular intervals. This yields an even higher integrity coat than tank plating, but you must buy a plater to do this well.

* P. O. Box 1452, Houston, Texas.

Ed. note. The stainless-steel treble "E" string for plectrum banjos is about number 22 AGW. It is three feet long and costs about 30¢ in music stores.

fig. 2. More of the materials used for plating along with three rings that I silver plated.



step-by-step plating

1. First take the work to be plated to a clean, well-ventilated area and wash the work with a strong detergent. Take off any scale or dirt with a single-edged razor blade, preferably by scraping. Then rough up with abrasive.

2. Place work in metal cleaner. Remove after tarnish is brightened, wash, and dry. Use a towel!

3. Using rubber gloves, open a jar of activator number 4. Place work in solution.

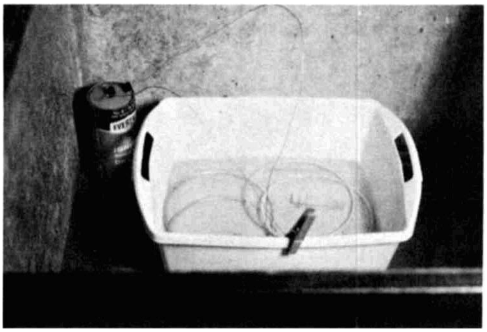
4. Cut off two lengths of wire and connect them to the dry cell or battery charger; positive connection is made to the silver anode. Mark with red nail polish.

5. Repeat step three until surface shows a white silver haze. If it is difficult to get this,

try grinding in the activator or striker with Scotch-britc saturated in the appropriate solution. If you are plating aluminum, it will be necessary to strike the work first with activator number 5, and then wash and repeat the silverizing process with activator number 4. Rubber gloves are essential!

6. Attach other wire (negative) to piecework; next place both in electrolyte number 316. Activator solutions must be kept separate from Coatalytes, except for number 4, which need not be rinsed off completely.

fig. 3. In-tank view of my plating setup.



7. Agitate while plating. Thick coats can be built up faster than the first, and critical, one.

8. Remove work, wash, and dry. Work set aside for later use should be resilverized immediately before an additional coat.

9. Plating time: up to four hours can be used with a current of 100 milliamperes, as a general rule.

some after thoughts

If you plan to do a lot of silver plating, it's a good idea to salvage the solutions. Reference 1 lists companies that buy plating solutions for recovering the precious metals in them.

Reference 1 also recommends alternate plating and deplating. A short period of deplating (by reversing the polarity of the voltage source) will remove any metal that has not been soundly deposited during the previous plating cycle. Thus each plating cycle deposits onto a base of firmer, smoother metal. This results in a brighter finish. The deplating cycle is very short: of the order of a few seconds.

glossary

building up	silver deposited in excess of 1 mil on the base metal
electroless plating	a molecular displacement chemical reaction where silver is deposited without electricity
coatalyte	electrolyte
integrity	quality of deposited silver
nonferrous	any metal not iron
silverizing	activating copper, brass and bronze for silver plating. Aluminum is first struck; then it is silverized
striking	activating the work or base metal for silver plating
stripping	process of lightly scraping or sanding successively plated work before building up
tank plating	immersion plating with electricity
throwing power	ability of a given electrolyte solution to deposit silver from an anode, through the solution used as electrolyte, to the properly activated base metal

references

1. Metal Finishing Guidebook Directory for 1968, Metals and Plastics Publications, Inc., 99 Kinderkamack Road, Wcstwood, N. J. 07675.

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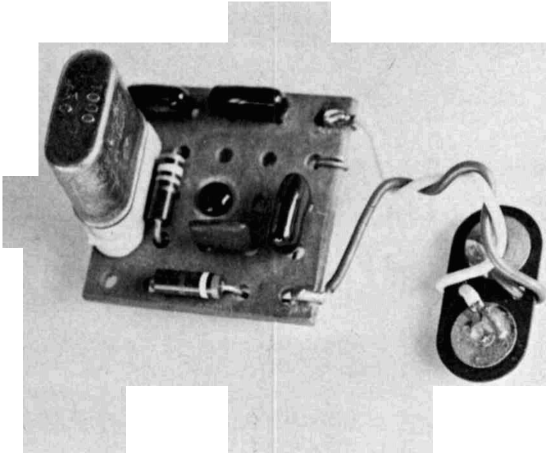


miniature crystal oscillator

A reliable low-cost crystal oscillator that isn't much larger than a postage stamp

Ev G. Taylor, W6DOR/W7BYF, 4100 Worthin n Dive, North H'ghlands lif 'a 95660

I **wonder how many** frequency standards were never built for the want of a special choke that is called out, lack of time to finish soldering all of the wires to the tube socket or need of a spare power supply to try it out? This oscillator doesn't require any of these things. The circuit is simplicity itself. The part requirements are four capacitors, two resistors, a crystal socket and a transistor. It's so simple that you don't even need hookup wire—the parts are placed so the component leads serve as the connections. The few parts that are used were placed on a piece of pre-punched terminal board slightly larger than a postage stamp. To be exact, it measures 1-3/8 inches square!

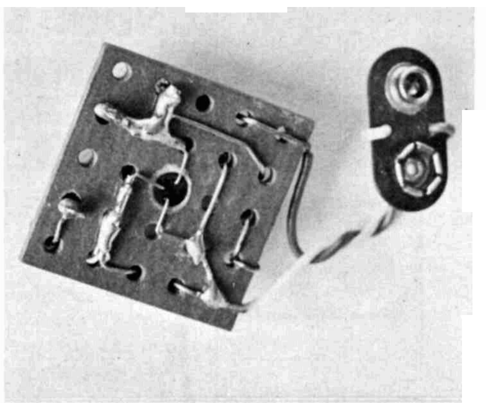
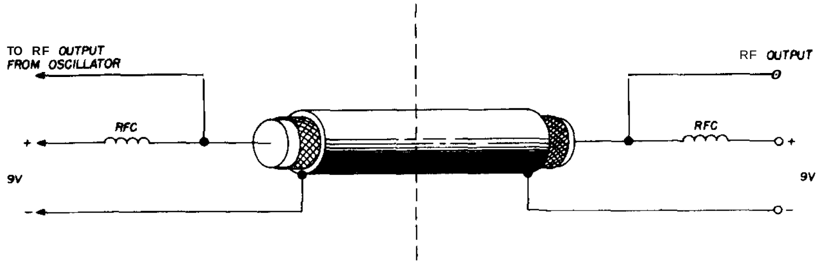


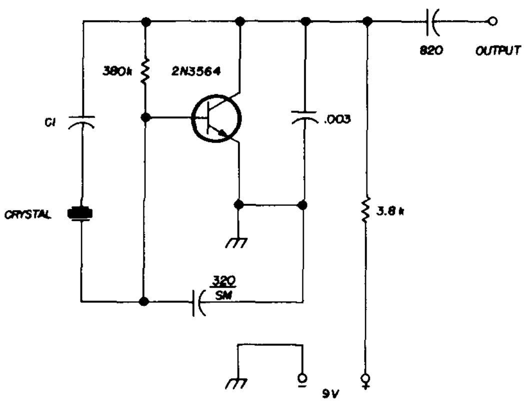
fig. 2. Method of getting dc power to the oscillator if you bury it in the ground for stability.



The power source is a 9-volt battery that can be picked up at your corner supermarket or drug store. If you use the oscillator as a 100- or 1000-kHz marker as a net-frequency "spotter," you can pick up the supply voltage from the cathode bias on the output tube. If your receiver is transistorized, you can use the battery in the set. If you put a low reading milliammeter in series with the voltage source, you can use this circuit as a crystal checker the next time you go looking for a good rock at your favorite surplus store.

This is a simple but reliable circuit. Even overtone crystals will work in it; 100-kHz crystals will oscillate with a 3-volt supply.

fig. 1. Circuit of the postage-etamp oscilator.



The pre-punched terminal board is cut so that one of the holes is in the center. This hole is enlarged sufficiently to accommodate the transistor.

The transistor is a vhf type that is available for less than a dollar. The board somewhat stabilizes what little heat is generated by the transistor since the board and transistor are in direct contact if you use the correct size drill. The other parts fit in like one, two, three.

A fixed capacitor is put in series with one lead of the crystal socket so you can set the frequency on the nose—its value may be different with various crystals and different trequencies. With a 1-mHz crystal that I used, the capacitor turned out to be 120 pF. It might be simpler for you to use a small trimmer in case you want to use this circuit with various crystals. For long-time stability be sure to use a silver-plated trimmer.

If you use this oscillator for checking WWV, I have a little suggestion for further stabilization. Put the unit in a glass jar, fill it with oil and bury it a few feet in the ground. The voltage supply may he fed through coax if you use isolation chokes at each end. This is the system used to feed power to CATV-system amplifiers.

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new dx record for 40,000 mhz

WA7EDI and W7CAF work over a 3720-foot path on 40,100 MHz

Hal Jackson, K7PMY, 6626 East Coronado Road, Scottsdale, Arizona 85257

A new DX record for two-way communications above 40,000 MHz was established at the Southwest Division Convention in Phoenix on the first of September. The new distance record, 3720 feet, was accomplished on 40,100 MHz; the old record, set in 1967 by W6FUV and W6SJO was 700 feet.

One station, WA7EDI/7, operated by Lorraine Cripps, was located on top of an office building in downtown Phoenix; on hand as witness was Ray Cripps, WA7EDH. In front of the Townehouse Hotel at the other end of the link was Gary Hamman, W7CAF/7; the witness at the hotel location was John Huntoon, W1LVQ, ARRL General Manager.

Similar equipment was used at each end of the 3720-foot path and consisted of microwave laboratory equipment furnished by Motorola's Aerospace Center of Scottsdale. The transmitter was made up of a Hewlett-Packard signal generator at 20,500 MHz driving a frequency doubler to 40,100 MHz; power output was 1 milliwatt. Modulated CW was used as shown in **fig. 1.**

Each receiver used a crystal detector with the output coupled to an HP-415E vswr meter which had an internal high-gain amplifier. Output from the high-gain preamp was coupled into an audio power amplifier to drive a speaker. Full break-in CW operation was provided by a 10-dB direction coupler as shown in **fig. 1.**

The antenna system at each end of the link was a three-foot parabolic reflector fed with an open-ended piece of waveguide. The measured half-power beamwidth of each antenna was 0.7 degrees; calculated gain was 48 dB above an isotrope or about 46 dB above a dipole. The mounting stands for the antennas and other equipment were designed and built by WA7EDH.

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Gary Hamman, W7CAF/7, with John Huntoon, W1LVQ, and John Griggs, W6KW.

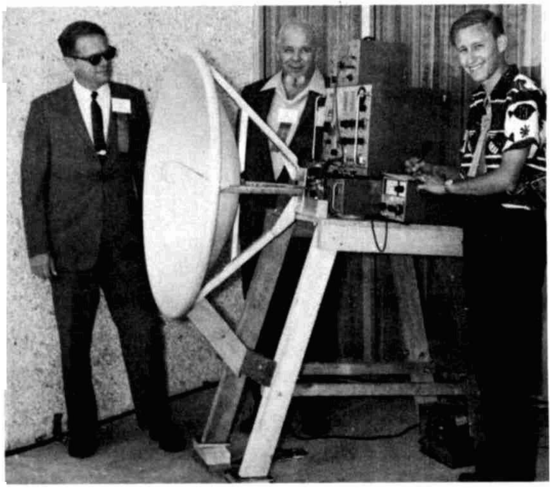
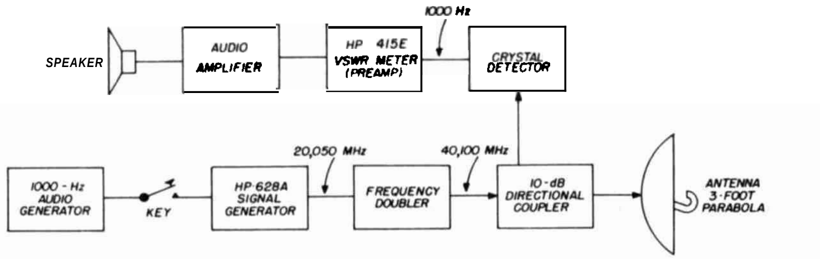
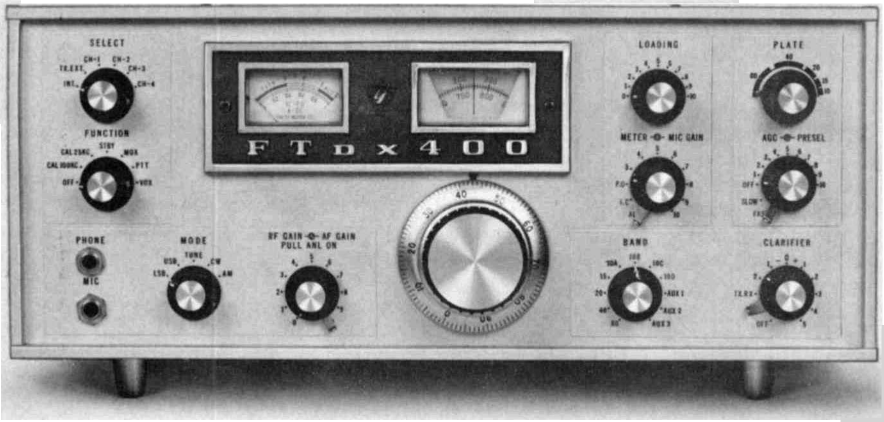


fig. 1. Block diagram of the system used at both ends of the path. Full break-in operation was provided by the 10-dB directional coupler.



Lorraine Cripps, WA7EDI/7, at the key on ona end of the path with her husband. Ray, WA7-EDH, looking on.

NOT FOR THE NOVICE



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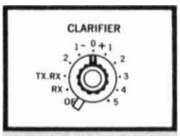
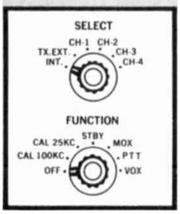
Design features include double conversion system for both transmit and receive functions resulting in, drift free operation, high sensitivity and image rejection * Switch selected metering * The FT dx 400 utilizes 18 tubes and 42 silicon semi-conductors in hybrid circuits designed to optimize the natural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

This imported desk top transceiver is beautifully styled with non-specular chrome front panel, back lighted dials, and heavy steel cabinet finished in functional blue-gray. The low cost, matching SP-400 Speaker is all that is needed to complete that professional station look.

SPECIFICATIONS: **Maximum input:** 500 W PEP SSB, 440 W CW, 125 W AM. **Sensitivity:** 0.5 uv, S/N 20 db. **Selectivity:** 2.3 KHz (6 db down), 3.7 KHz (55 db down). Carrier suppression: more than 40 db down. Sideband suppression: more than 50 db down at 1 KHz. Frequency range: 3.5 to 4, 7 to 7.5, 14 to 14.5, 21 to 21.5, 28 to 30 (megahertz). Frequency stability: Less than 100 Hz drift in any 30 minute period after warm up.

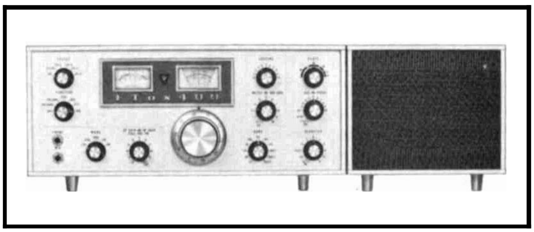


CLARIFIER CONTROL— Does the work of an external VFO — allows operator to vary receive frequency 10KHZ from transmit frequency, or may be used as an extra VFO combining transmit and receive functions.



SELECT CONTROL — Offers option of internal or outboard VFO and crystal positions for convenient preset channel operation.

FUNCTION CONTROL—Selects crystal calibration marker frequency and desired transmit mode of operation.



FT DX 400 $599.95 — SP-400 $14.95

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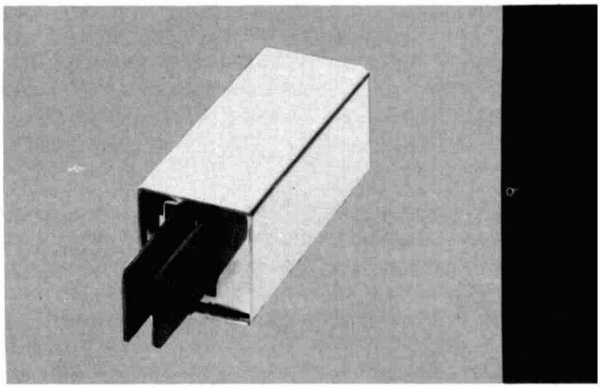


the james research permaflex key

A new design for solid-state keyers

Jim Fisk, W1DTY, Post Office Box 25, Rindge, New Hampshire 03461

A number of different keys have been introduced and manufactured over the years, and each new design borrows a little something from previous units. First came the hand key, then the bug and the cootie key (or was it the other way around), and finally, adaptations of the basic bug for use with electronic keyers. The new James Research Permaflex key represents a radical departure in key design. It doesn't even look like a key in its handsome chrome-plated case. The function is still the same, but modern materials, design and technology are used to manufacture a key to meet the needs of today's solid-state electronic keyers.

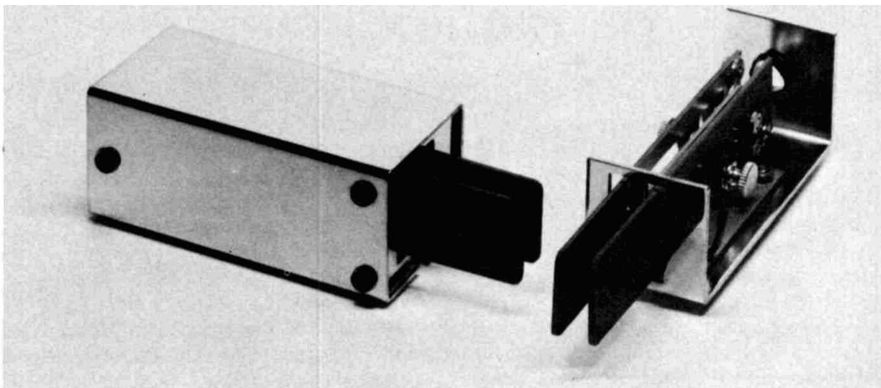
If you have ever used a conventional bug with a solid-state keyer, particularly integrated-circuit units, you can appreciate the problems caused by corroded silver keying contacts. Problems don't show up with vacuum-tube keyers because there's enough current drawn through the contacts to keep thrm clean. However, with solid-state devices, it's a completely different story. The current through the contacts is infinitesimal. and if you don't clean the keying contacts fairly often you'll be plagued with erratic keying. I've tried a number of solid-state keyers, but problems with erratic keying had always forced my back to the 'old reliable' vacuum-tube unit. I wonder how many other amateurs have had the same problem?

Another problem with conventional keys is contact bounce. Transistor and integrated-circuit keyers are a lot faster than most vacuum tube units; if your key has any contact bounce (and most of them do), you'll experience further trouble with badly made CW characters. Any corrosion that the contacts may pick up further aggravates this problem. If you have any doubts about contact bounce, just run a little dc through your keying contacts and take a look at the waveform on an oscilloscope. You'll be amazed at the 'fuzzy' leading edges caused by bounce.

The James Research Company took a very close look at these problems, and then designed a unit that would solve them. For the corrosion problem, they used gold-diffused silver; the contacts will take up to 8 amperes and will not corrode. The pivotless two-paddle design using fiberglass epoxy with gold-plated conductors results in a low-mass paddle with equal performance at both low and high currents. Tension can be adjusted from 5 to 50 grams. None of the adjusting screws carries any current, so there are no problems with high-resistance paths.

Contact bounce with the Permaflex key is practically non-existent. In oscilloscope tests that I've made, I've been unable to detect any contact bounce at all. This shows up when using solid-state keyers too. With a conventional key, I was plagued by extraneous dots when the paddle was pushed to the dash side. With the Permaflex any keying errors I get are completely my own.

The Permaflex key is small—only 1-9/16-inches square by 3-3134-inches long—but the silicone rubber feet do a remarkable job of preventing key "walking" during operation—as long as you keep them clean. Actually, there are two sets of feet on the key—one set is used when the key is operated as a sideswiper, the other set when it's used as a hand key. This is one of the more interesting applications of the Permaflex key; if you're working a station where a hand key would be more suitable than an electronic bug, you simply turn the key on its side and use it as a hand key! Three additional silicone-rubber feet are provided for this purpose.



The biggest problem with the Permaflex key is getting used to it. After using a regular "bug" for a number of years, it's difficult to adapt to a keyer that has 5 grams armature pressure. If you operate an electronic key, you probably bat the keyer around a lot—I know I do. Keying with the Permaflex doesn't take any effort at all, so it takes a little getting used to. But after you've mastered it, it's pretty hard to go back to the high-mass paddles used in conventional designs.

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propagation predictions for december

Old sol exposed for a cold winter's column

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

Winter has arrived in the northern-hemisphere ionosphere. Unlike the earth-atmosphere weather system, which has a thermal time constant of about a month and a half, the ionosphere responds almost directly to changing solar declination. The ionization time constant is measured in minutes for the E- and F1-layers, and probably tens of minutes for the F2-layer.

Little seasonal change in ionospheric parameters is expected during December or most of January since the solar declination is close to 22 to 23 degrees south during most of these months. Accordingly, the range predictions presented this month are sufficiently accurate for both months.

The seasonal changes from earlier months are these:

1. Lower noise levels

2. Higher daytime F2-layer muf's in north temperate latitudes

3. Lower nightime muf's in north temperate latitudes

4. Shorter duration (and maximum path length) of Northern Hemisphere daytime band openings.

The shorter hours of daylight will reduce somewhat the path muf's for Northern Hemisphere paths greater than 5000 miles in length —even though the path muf's for single-hop paths may be increased slightly over those of November. December and January are also the months of a secondary peak in occurrence of sporadic-E in northern-hemisphere temperate latitudes.

the quiet sun

Last month I indicated dissatisfaction with the use of running average smoothed sunspot numbers for prediction of median ionospheric conditions (critical frequencies). This month I will describe the quiet sun so that you will have a background for the discussions of solar-terrestrial relations that will follow in coming months.

The sun is powered by self-sustaining thermonuclear reactions confined by gravity. The solar diameter is about 864,000 miles, almost four times the distance from the earth to the moon. Composed primarily of hydrogen and helium, average density is less than that of the earth; however, the forces of gravity compress the deep interior of the sun to about 5 times the density of gold.

The temperature at the center of the sun is believed to be about 73 million degrees Kelvin and almost all (about 88%) of the atoms are ionized. Most of the nuclear energy of the sun, from the transmutation of hydrogen to helium, is produced near the center of the sun where the temperature is sufficient to permit fusion.

Nuclear energy, gravitational energy and thermal energy form a feedback system that keeps the interior of the sun in thermal equilibrium. Energy is transported through most of the solar interior by radiation; however, there is a surface layer about 60,000 miles thick where the principal energy transport mechanism is convection. The solar temperature falls from a hot million degress at the base of the convection layer to about 4400°K just outside the visible disk (photosphere).

The radiation from the center of the visible disk is, to first order, that of a black body at a temperature of 6100°K. Thus the peak intensity of radiation occurs in the visible part of the spectrum. The amount of visible light emitted by the sun is essentially constant with time. However, the day-to-day variations in the quiet ionosphere suggest that the ultraviolet and x-ray emissions (not black-body) do vary appreciably from day to day and throughout the solar cycle.

Superimposed on the black-body spectrum are absorption and emission resonance lines and bands from ionization of hydrogen, helium, oxygen, nitrogen, calcium, carbon, silicon, magnesium and iron in the sun's atmosphere. Only one particle in 6000 is ionized at the photosphere, but the temperature and ionization percentage increase with height in the sun's atmosphere.

The inner atmosphere, called the chromosphere, extends to about 12,000 miles from the photosphere. The temperature of the photosphere rises from 4400°K near the photosphere to about 300,000°K near the corona, which is the outer atmosphere of the sun. The corona is at a temperature of 1 to 2 million degrees *and* extends to as great as 20 solar radii (8.6 million miles).

A small fraction of the solar atmosphere is accelerated to supersonic velocities in the corona and continually escapes to form the solar wind. The earth is constantly bombarded by the solar wind, which has a quiescent density in the vicinity of the earth of about 10 particles per cubic centimeter and a velocity of 200 miles per second. The solar wind distorts, and in fact terminates, the earth's

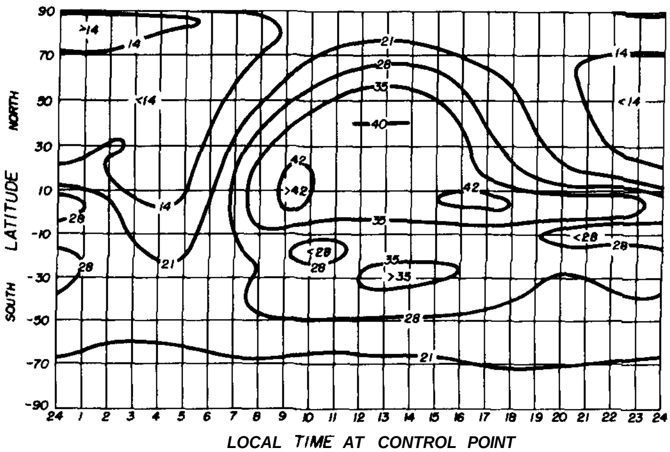

fig. 1. Time chart of predicted median MUF for December 1968 for the American continent.

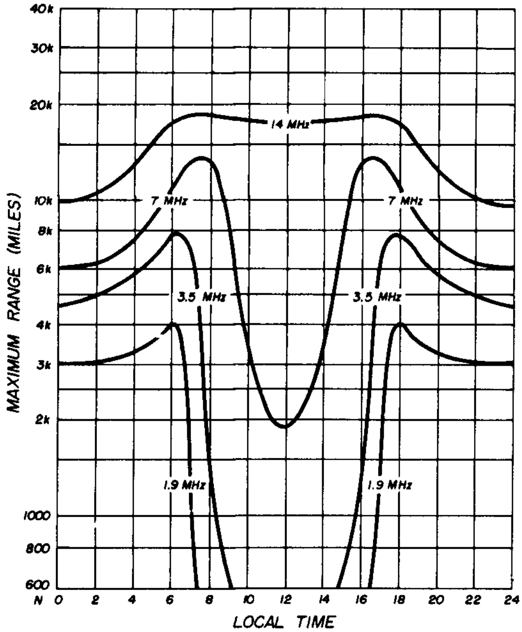

fig. 2 Maximum range to tho north from 38° N latitude due to absorption.

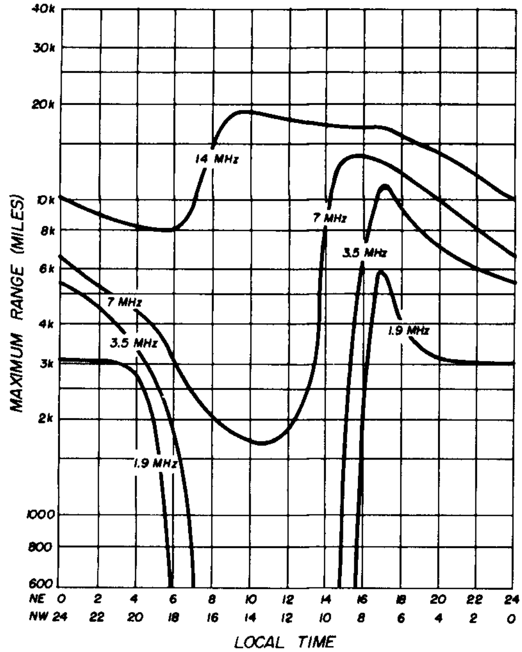

fig. 3. Maximum range to the north-east (top time scale) and to the north-west (lower time scale) from 38° N latitude due to absorption.

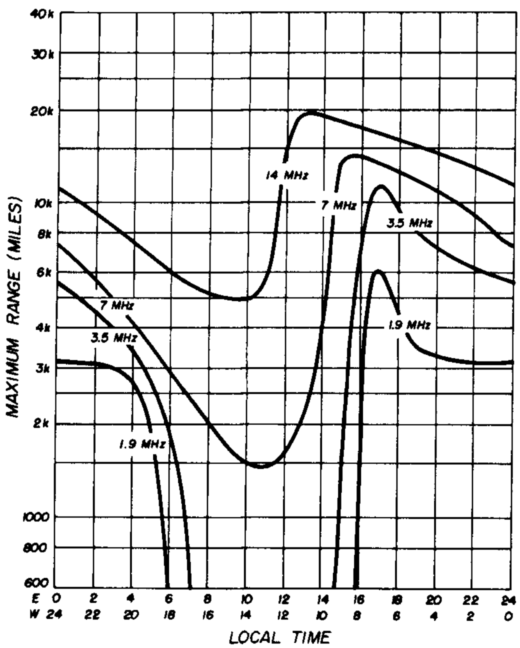

fig. 4 Maximum range to tho **east** (top **time** scale) and to tho **west (lower time scale)** from 38° N **lati**tudo duo to **absorption.**

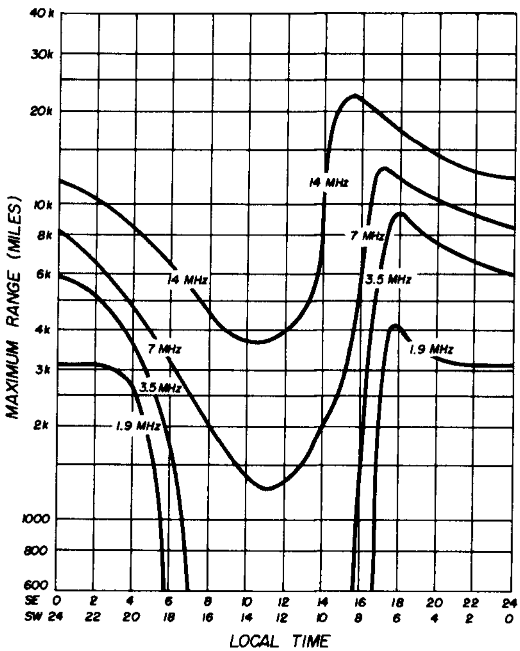

fig. 5. Maximum **range** to tho south-erst (top time scale) and to the **south-west** (lower **time scale)** from 38° N latitude duo to absorption.

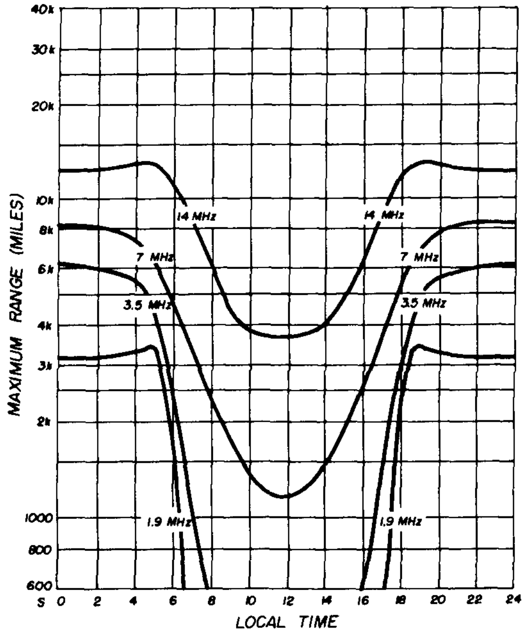

fig. 6. **Maximum range to the south from 38° N latitude due to absorption.**

magnetic field in space and is responsible for the aurora. The solar wind is believed to extend past the orbit of Mars.

propagation summary for december

160 meters. During December and January, low night-time muf's will spread over large areas of the Northern Hemisphere, resulting in reflection from higher altitudes (and thus longer hops and less path loss for long paths) than possible during earlier and later months. Morning and evening horizontal gradients of ionization are also greatest during these two months. Communication over long paths to the Southern Hemisphere is not expected due to large summertime atmospheric noise levels at the other end of the path. Night-time 160-meter ranges will be almost equal to 80-meter ranges for a 10-dB increase in power or antenna gain. Daytime absorption can be used to advantage to reduce interference from LORAN stations, and East-Coast stations can work west just before sunrise when there is no propagation to the East-Coast LORAN stations; East-Coast stations can work east just after sunset when there is no propagation to the West-Coast LORAN stations.

80 meters. Some of the same comments made about 160 meters apply. During disturbed conditions and after about 10 PM local time, skip zones may occur to a couple of hundred miles. Local operations may be troubled at night by interference from stations 1000 to 2500 miles away.

40 meters. Forty will be the best band for nightime DX but will be crowded. Try operating near sunrise and sunset (for west and east, respectively) and between 10 PM and 4 AM for maximum skip distances.

20 meters. Twenty will be good for daylight openings to the Northern Hemisphere and twilight openings to the south. Don't count on twenty being open to anywhere before 5 AM or after 10 PM local time, except for an occasional sporadic-E opening.

15 meters. Fifteen meters will offer somewhat wider coverage (especially for polar paths) than ten meters, but signals may be slightly weaker. Fifteen should open about 6 AM to the southeast and close about 8 PM to the southwest.

10 meters. Ten meters will open about 1 hour later and close about 1 hour earlier than 15. Two-hop polar paths are most likely near noon, path-midpoint time.

6 meters. Some single-hop F2-layer openings may occur between coasts (1800 to 2000 GMT) and to the south (9 AM to 4 PM local time). Some combination sporadic-E and F2-layer openings may offer extreme range contacts to the south of east and west.

Sporadic-E. A minor sporadic-E peak will occur during December and January allowing 20, 15, 10, and 6 meter openings to distances up to 1400 miles by single hop and occasionally 2500 miles plus by multiple hops. During the peak of the previous sunspot cycle 144 MHz sporadic-E was worked during the winter E_s season, and so 144.1 MHz or whatever frequency you watched this summer would bear watching again.

Meteors. The Geminids meteor shower is expected December 10 through 14 and the Quadrantids meteor shower is expected January 1 through 4.

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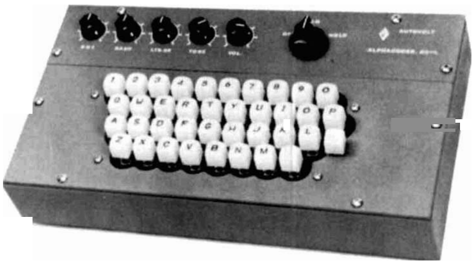
with the NEW AUTOVOLT ALPHACODER

Model AC-1



The new **Autovolt Alphacoder,** AC-1, makes any letter or digit with one stroke of your **finger.** You can produce perfectly **uniform** code, **including** preset letter-to-letter **spacing using** the standard **electric typewriter** keyboard of the AC-1.

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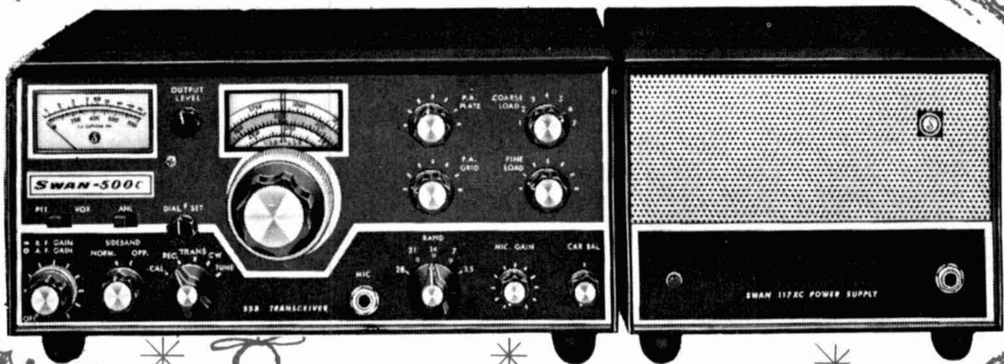

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This year we can drop those subtle hints for you. Simply fill out the blank and mail it directly to **Raytrack** Company; within a few days, your Santa (he or she) will receive a letter which contains all the information needed to purchase your gift. Please don't wait too long, though, because Santa's Helper is going to rely on the U.S. Mail . . .

Best Wishes for a Merry Christmas!

Raytrack Company

TALK POWER

The AutoLevel is the **ultimate** in volume compressors. This unique device provides all the talk power your transmitter can use. The AutoLevel was designed for use with SSB or AM transmitters, with or without ALC capabilities.

The AutoLevel is not an audio or RF clipper - all compression is attained by a photo-optical regulator which provides **14** dB's of compression with a minimum of wave form distortion

The AutoLevel is easily installed in the mike line, and it contains its own power supply; (there's no need to bother with batteries). It can also be used with your phone patch for the utmost in ease of operation

When **you're** ready for the finest, ask your local dealer for the AutoLevel.

SPECIFICATIONS

dB's compression - **14** dB minimum
Wove form distortion - **neglible**
Input impedance - suitable for dynamic or crystal microphone
Output impedance - **50K** (nominal)
Power supply - 115 **volts AC**

Dimensions - **2-3/4"** x 4-1**1/16"** x **6-3/8"**
H W D
Weight- 32 ounces
Color - Bone White with Black front panel
Price - $87.50

(Please print or type)

YOUR Name: .... Call: ....
Address: ....
City: .... State: .... Zip: ....

SANTA'S Name: ....
Address: ....
City: .... **State:** .... Zip: ....

TELL SANTA I WANT:

☐ The **AutoLevel** Electronic Volume Compressor ☐ The Horizon VI **6-meter** Receiving Converter

VALUE ENGINEERING
. . . to 2 meters

SWAN 250C
SIX METER TRANSCEIVER

We are proud to announce production of the deluxe model 250C, new contender for top honors on the 6 meter band. The 250C is built to performance standards of the highest order, including a receiver front end employing dual Nuvistors in cascode with a noise figure of less than 3 db. Deluxe features include a built in 250 KC crystal calibrator; 'S' meter; selectable sideband; vernier control of the megacycle tuning range; and accessory sockets on the rear apron to receive the model 210 external VFO, the new Swan Noise Silencer and the VX-2.

The famous model 250 is a consistent winner in VHF competition —The 250C is designed to set new records! We are confident that the 250C will satisfy the most critical requirements of the serious VHF operator.

$420

SPECIFICATIONS:

- Frequency Range: 50-54 MC
- Power Rating: 240 watts PEP Input in SSB mode, 180 watts CW Input, 75 watts AM input
- Two 6146 B Power output tubes
  Distortion Products: down approx. 30 db
- Unwanted Sideband: down more than 40 db
- Carrier Suppression: better than 50 db
- Receiver Noise Figure: Better than 3 db, with two 6CW4 nuvistors in Cascode
- Selectivity: 2.8KC at 6 db down, with crystal lattice filter at 10.9 MC.
- Antenna Matching: Wide range Pi network.
- Metering circuits: S-meter on Receive mode, P.A. Cathode Current and relative output in transmit mode.
- 250 KC Crystal calibrator
- Selectable upper and lower sideband
- Receiver Mode switch provides for AM reception
- Accessory sockets for noise silencer, external VFO and VOX unit

ACCESSORIES:

MATCHING AC POWER SUPPLY
Model 117XC $105

12 VOLT DC SUPPLY
Model 14-117 .... $130

SWAN NOISE SILENCER
Model NS-1 .... $ 36

EXTERNAL VFO
Model 210 .... $120

PLUG-IN VOX UNIT
Model VK-2 .... $ 35

PHONE PATCH
Model FP-1 .... $ 48

2KW LINEAR AMPLIFIER
Model 6B. With power supply $660

AND FOR 2 METERS
SWAN TV2 TRANSVERTER

A receiving and transmitting converter, which may be used with the 250C or 500C to operate in the 144-148mc band. Provides 240 watts P.E.P. transmit power, and low noise receiver front end with Nuvistors in Cascode.

$295

One of the main reasons Swan maintains its position as the leading manufacturer of amateur radio equipment is our dedication to the principles of Value Analysis, and Value Engineering.

We continually examine our products to find ways of improving their performance, reducing costs, and making them better values. As new components are developed in the electronic field, we analyze them to see how they can be used to improve the performance of our transceivers.

The most recent results of our Value Analysis and Value Engineering are shown on these pages: The 500C 5 band SSB Transceiver, our new External VFO, the 510X Mars Oscillator, and introducing our new 6 Meter Transceiver, the 250C.

Yet with all the improvements, the increased reliability, and proven performance, Swan's Value Engineering results in substantially lower prices than competitive equipment. And every piece of Swan equipment is backed up by service second to none.

Value Analysis, Value Engineering keep Swan in the lead. Visit your Swan dealer soon.

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OCEANSIDE, CALIFORNIA
A Subsidiary of Cubic Corp

VALUE ANALYSIS

From *80* meters. . .

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FULL COVERAGE
EXTERN[illegible]

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Unit is designed fo[illegible]
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It provides for tr[illegible]
receiving on separate frequenc[illegible]
and plugs directly into the bac[illegible]
the 500C. A separate Dual-[illegible]
adaptor is no longer required, si[illegible]
the relay control circuitry is b[illegible]
into the 508. A panel control [illegible]
mits selection of VFO's so that [illegible]
eration may be transceive m[illegible]
with the 500C VFO, transceive [illegible]
the 508 VFO, or transmit on [illegible]
500C and receive o[illegible]
Model 508 features [illegible]
500 kc each, with [illegible]
It may also be use[illegible]
transceiver.

MARS OSCI[illegible]

Ten crystal contr[illegible] els
with [illegible]
Plugs directly into [illegible]
may also be used w[illegible]
and other Swan tra[illegible]

[illegible]DEL 510[illegible]
[illegible] crysta[illegible]

SWAN 500C
FIVE BAND
TRANSCEIVER

80 through 10 meters • 520 watts • Home station, mobile, portable operation • SSB-CW-AM.

The new model 500C is the latest evolutionary development of a basic well proven design philosophy. It offers greater power and additional features for even more operator enjoyment. Using a pair of the new heavy duty RCA 6LQ6 tetrodes, the final amplifier operates with increased efficiency and power output on all bands. PEP input rating of the 500C is conservatively 520 watts. Actually an average pair of 6LQ6's reach a peak input of over 570 watts before flattopping!

The 500C retains the same superior selectivity for which Swan transceivers are noted. The filter is made especially for us by C-F Networks, and with a shape factor of 1.7 and ultimate rejection of more than 100 db, it is the finest filter being offered in any transceiver today.

For the CW operator the 500C includes a built-in sidetone monitor, and by installing the Swan VOX Accessory (VX-2) you will have break in CW operation.

Voice quality, performance and reliability are in the Swan tradition of being second to none.

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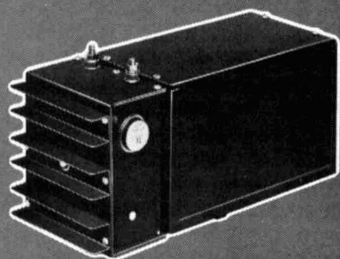

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12 VOLT DC SUPPLY

Complete D.C. supply for 12 volt mobile or portable operation. Includes cables, plugs, and fuses. Will also operate from 117 volt A.C. by detaching the D.C. module & plugging in 117 volt line cord. Negative ground standard. Positive ground available on special order.

$130

PHONE PATCH,
Model FP-1 . . . .$48

ASK THE HAM . . .
WHO OWNS ONE

new products

radio communication handbook

The new Radio Communication Handbook published by the Radio Society of Great Britain is one of the most exciting handbooks I have seen in a long time. This new book, which weighs in at six pounds, covers every aspect of amateur radio communications and is filled with solid-state projects and circuits.

The topics that are covered include electronic principles, vacuum tubes, semiconductors, hf and vhf/uhf transmitters and receivers, keying and break-in, modulation systems, single sideband, rtty, propagation, hf and vhf antennas, noise, mobile equipment, power supplies, interference, measurements and station layout.

To give you an idea of the contents of the book, lets take a look at the chapter on single sideband, 108 pages that is almost a handbook itself. This chapter leads off with a comparison with a-m, delves into a lengthy discussion of balanced modulators with many good circuits and as well as circuits that should be avoided.

The section on ssb filters covers both mechanical and crystal types, shows response curves and how to measure them and circuits that are designed for the various types. Following filters is an excellent discussion of frequency conversion, along with charts for determining frequencies that will result in minimum spurious signals.

The section on linears discusses various circuits with the advantages and disadvantages of each. Measurements of operating linear-amplifier parameters are covered in detail. The construction section features four complete ssb transmitters and transceivers as well as several linears and power supplies. Perhaps most interesting is the solid-state six-band transceiver that covers all the bands from 160 to 10 meters: the only vacuum tubes are in the driver and final power-amplifier stages.

Nearly all of the transistors and tubes used in the projects in the **Radio Communication Handbook** have American equivalents, so there is no problem in duplicating them. The staff of the RSGB has thoughtfully provided many of these equivalents in the text; the device equivalents that are not listed in the text are listed on a small sheet that is supplied with all copies in the United States and Canada.

If you're troubled by the lack of semiconductor circuits and modern technology in the other handbooks, the **Radio Communication Handbook** is the answer; it is the most up-to-date radio handbook that has crossed my desk. $11.95 postpaid in the United States and Canada from Comtec, Book Division, Box 592, Amherst, New Hampshire 03031.

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BOTH STORES CLOSED THURSDAYS

fet applications handbook

Field-effect transistors are becoming more and more popular as prices start to drop. This book, edited by Jerome Eimbinder, is based on current practical material prepared by some of the most capable men in industry. It contains a wealth of data on field-effect transistors and their applications in practical circuit designs. FET types, parameters and characteristics are covered in the early chapters, followed by many practical circuit designs covering all phases of electronics.

To provide the reader with a good solid base to work from, current-voltage relationships and biasing considerations, complete with transfer curves, are thoroughly discussed. Considerable attention is devoted to the FET as a constant-current source with detailed information on the constant-current element and cascaded operations. Other topics include oscillators, noise measurements, audio and vhf amplifier design and applications for the photo-FET. The appendix contains most often-needed FET design data and charts arranged to serve as a quick-reference source of useful field-effect transistor information. $12.95 from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

new vhf fet from motorola

You vhf and uhf enthusiasts will be happy to know that Motorola has just introduced two budget priced plastic versions of the popular 2N4416 JFET. The MPF106 and MPF107 feature unusually low noise figures —2.0 dB maximum at 100 MHz and 4.0 dB maximum at 400 MHz. In addition to being ideal for rf front ends, they will work equally well in any low-noise, high-gain amplifier from dc to 400 MHz.

These N-channel devices provide a power gain of 18 dB minimum at 100 MHz and 10 dB minimum at 400 MHz. Input and output capacitances are 5.0 and 2.0 pF, respectively. For more information on the MPF106 or MPF107, or Motorola's complete line of new FET's, write on your company letterhead to Motorola Semiconductor Products, Inc., Department TIC, P.O. Box 20924, Phoenix, Arizona 85036.

easy way to **service** radio receivers

This brand new book by Leo G. Sands shows you how to diagnose and correct troubles in just about every kind of a-m broadcast receiver—tube or transistor. The book starts out with descriptions of typical receiver circuits and how they work. Following chapters are devoted to various troubles with step-by-step procedures for efficient trouble shooting. To help speed the diagnosis, numerous charts are presented that show trouble symptoms and the most probable causes.

Other chapters provide complete details on alignment, quick-check techniques, circuit modifications, test equipment and how to use it and special worthwhile test procedures. With this book, you should be able to repair any a-m receiver quickly. $6.95 hardbound; $3.95 paperback from TAB Books, Blue Ridge Summit, Pennsylvania 77214.

new ways to diagnose electronic troubles

With this new book, Jack Darr makes even the most complicated electronic circuit seem simple. He draws on his many years of experience to explain how almost any trouble-shooting job can be boiled down to a logical sequence. By analyzing typical schematics, the reader will be able to find his way through even the most complicated looking power supplies and associated circuits—regulators, filters and voltage-distribution networks. Author Darr also covers quick ways to check color sweep circuits, including high-voltage supplies, as well as vertical oscillator and output stages and video circuits. Other chapters cover tuners, troublesome agc and sync circuits, trouble symptoms and signal tracing. If you ever have to put a radio or television set on your workbench, this book should save you a lot of time. $3.95 paperback, $6.95 hardbound from TAB Books, Blue Ridge Summit, Pennsylvania 17214.

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comments

Dear HR:

Congratulations on your new magazine. The April issue, which is the only copy I've seen, intrigues me. Particularly, Larry Allen's "Repair Bench" column fills a long-standing need in ham publications . . .

Louis G. Madsen, WAGQYN

Dear HR:

I happened across your first edition here in W9/K9 land, and all I can say is wow! Even my somewhat non-technically oriented xyl was impressed by the appearance of the magazine.

Elliott Kanter, K4YOC/9

Dear HR:

Thought I would drop a line commending your fine magazine, ham radio. Being interested in keeping up with the state of the art, especially in vhf and antennas, I so far find your periodical well worth adding to my library of magazines which consists of QST (December 1916 to present) and CQ (January 1945 to present).

Bill Wakefield, WB6WBC

Dear HR:

I received your magazine and I couldn't help sending my congratulations on a beautiful job and well done. I have been in ham radio 39 years and a pressman printer 53 years, so I know a good job when I see it.

Bill Stone, WGFAV

Dear HR:

. . . I have the first three issues—but can't seem to find June––so, the only alternative is to subscribe. You have a new, fresh and constructive approach which should do much to upgrade the state of the art. Good luck!

John Koehler, W9DGB

Dear HR:

As an individual amateur radio operator for the past ten—almost eleven—years. I have read and enjoyed many electronics-oriented publications. Having received my formal education in the field of English, I can obviously claim that any knowledge that I have gained of electronics has been sweatfully developed through "trial-and-error" cramming, or thank goodness, through another and infinitely more enjoyable means. This other means of which I speak is, of course, such publications as the new **ham radio.** I feel that your new venture captures much of what is good about our hobby itself, encompassing the joys of experimentation, discovery and learning. Consequently one of the prime products of the unique amateur radio experience has always been, and continues to be, young men (and, yes, occasionally ladies) who are intellectually and scientifically awake.

I am most confident that your publication will be very successsful, since you have already presented several excellent features which really represent "communications technology" of **today.** While confessing that some of the material has not been within my grasp thus far, I commend the manner in which you present these difficult phases of modern electronic technology. Please make no mistake—I also enjoy the "operating" aspect of our hobby, and subscribe to other amateur publications which tend to accent these areas of our hobby. But I trust that our hobby will long be the better off by the addition of your magazine to the electronics publishing scene.

With cordial 73 for vour continued success,

Will Smith, K4SAY

Dear HR:

Received my latest magazine three days ago and I've already gone through it from cover to cover. Let me say I honestly believe you've a truly superior magazine; better than any other so-called amateur radio magazine put out today.

Your articles are fine, well written in not too difficult-to-understand fashion, well illustrated, etc. It is my type magazine —one who is primarily interested in how-to-build-it articles and some theory review. The ssb articles have been fine.

Joseph H. Smith, WSZDN

Dear HR:

I received a free copy of your magazine a few days ago. I was very pleased with the magazine. Almost all of the articles were the practical, useful type that I like to have in my library. I am studying for an amateur license and I appreciate the careful explanation of the why of the circuits. The magazine is remarkable in the high quality and quantity of useful information. I was as pleased with the absence of sections devoted to special interest groups. . .

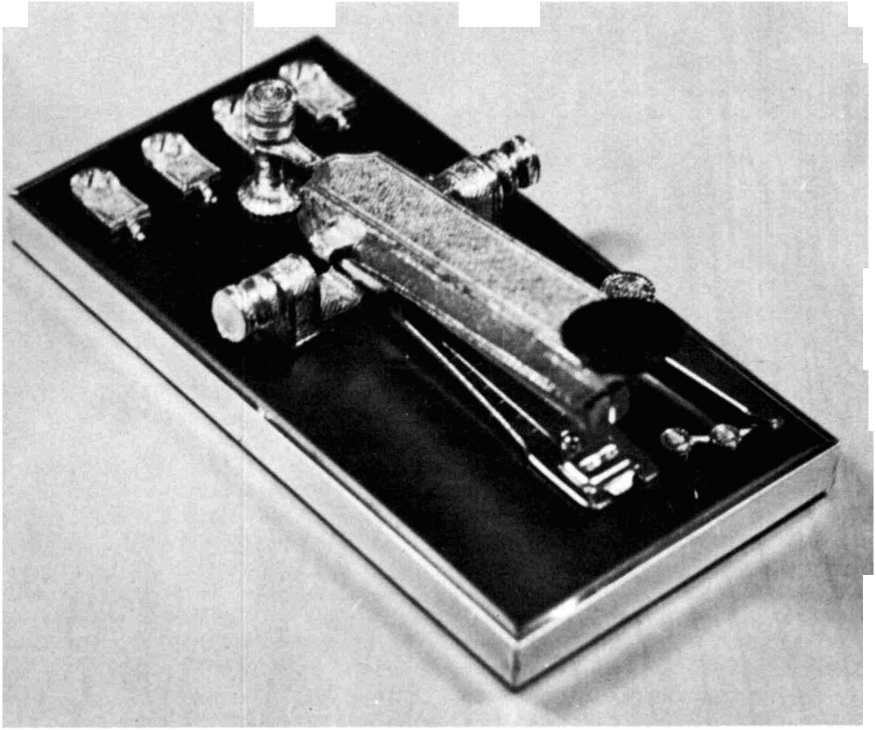
David Robertson
St. Louis, Missouri

Dear HR:

. . . I like the timely publication, title arrangement, presentation, size, and the type of paper used and the price seems right. Each of your feature articles was well prepared, informative and easy to read, and with few exceptions, appeared to me to be accurate. Of course, advertisements help to defray expenses and are generally necessary to balance the books, in addition to bringing one up to date and knowing what is on the market.

Professionally I am generally involved in a systems aspect. However, personally I enjoy technical magazines, predominately electronics, and prefer articles in circuitry using solid-state devices for receivers, transmitters (especially ssb), control devices, and not to forget, antennas. I am not interested in operating news or amusing stories, although I suppose it does have some reader appeal.

J. Roy Wolfskill, W1YLQ



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CHRISTMAS GIFT subscriptions to HAM RADIO. See **Special** offer this Issue. Order today.

73, FEBRUARY **1961** thru 1964. $14; QST, 1958 thru 1963. $27 or $5 yr.; Heath HD-10, $30; Vibroplex Zephyr, $11; **Simpson I-176** VOM, $12. All postpaid. **K6SRM**, 484 Fourth Street East, **Sonoma**, California 95476.

SWAP: Have 26-ampere, 117 volt variable transformer (similar to a Variac) with adjustable output from zero to 135 volts. Wanted: 10-amp, 220 volt variable transformer with or without **over-voltage.** Also wanted. January and March 1961 issues of 73 magazine. Jim Fisk, **W1DTY**, Box 25, Rindge, N. H. 03461.

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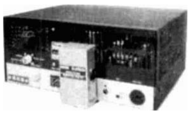

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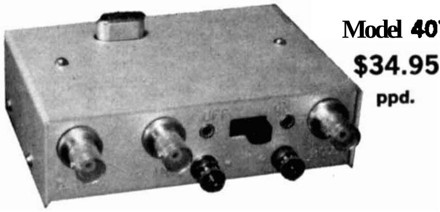

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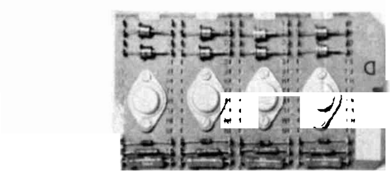

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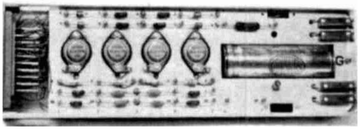

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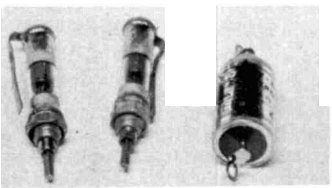

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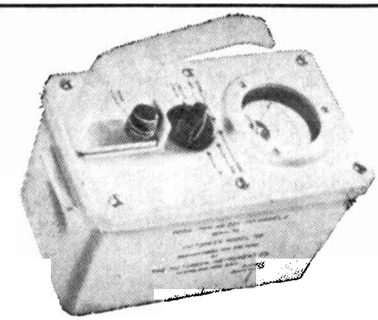

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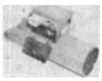

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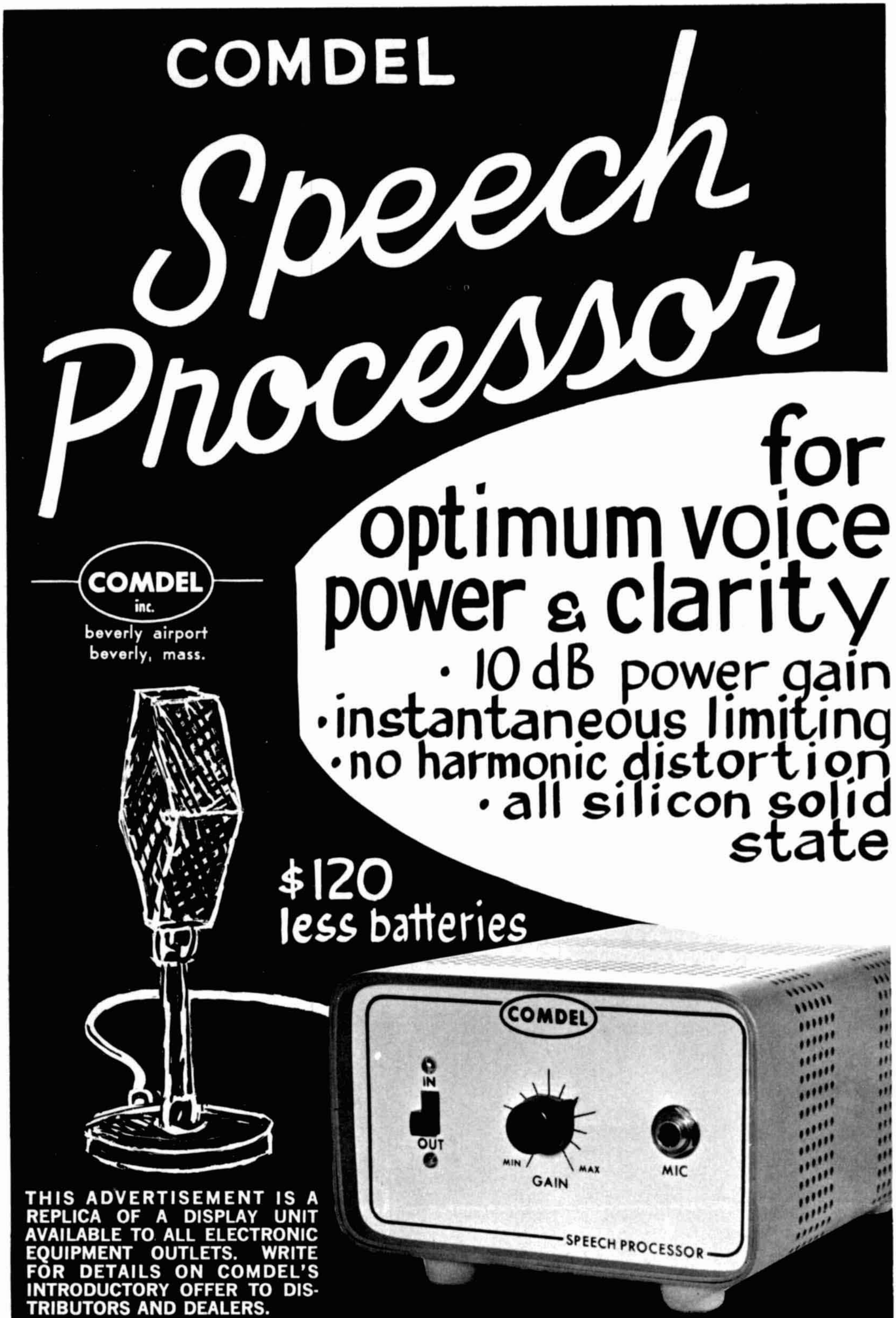
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index to ham radio magazine volume 1–1968

This index covers all articles published in **ham radio** during 1968. The articles are listed alphabetically under each category along with the author, page number and month. Categories are: antennas and transmission lines; audio-frequency equipment; commercial equipment; construction techniques; features and fiction; fm and repeaters; integrated circuits; keying and control; measurements and test equipment; miscellaneous technical; power supplies; propagation; receivers and converters; RTTY; semiconductors; single sideband; transmitters and power amplifiers; and vhf, uhf and microwave. Articles followed by (HN) appeared in the **ham notebook.**

antennas and transmission lines

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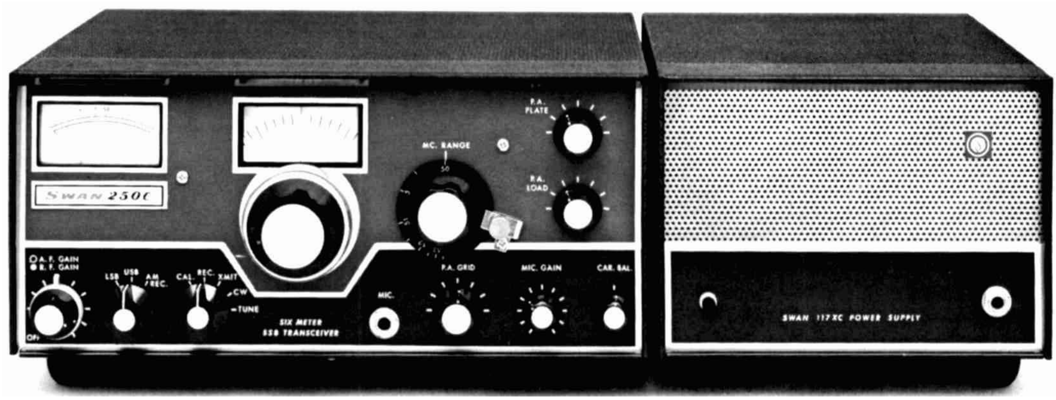
vhf, uhf and microwave

SWAN AND VHF



The new Swan 250C features a noise figure under 3dB with 1/2 microvolt sensitivity at 50ohms for 10dB signal plus noise-to-noise ratio.

Not many hams realize that for almost fifteen years Herb Johnson, **W6QKI**, founder of Swan, has been a major contributor to VHF activities, especially antenna designs. You will see this when you purchase Radio Publication's VHF Handbook by Bill Orr and **W6QKI.** Herb's love for the VHF challenge led to the immensely popular Swan 250, thousands of which have been sold.

Now Swan has vastly bettered the 250 with their C model, available concurrently with this ad. Although the price has gone to $420.00, many additions or refinements have been made. To begin with, the megacycle control knob has been recalibrated and fitted with a vernier. This, together with a built-in 250 KC marker calibrator. enables full band coverage with an accuracy of 2 KC.

The panel meter now functions as an S meter on receive and additionally serves as an RF voltmeter for tune-up as well as cathode current monitoring of the final. The **6146B's** continue to be used with 240 watts PEP input on either upper or lower sideband. On CW its rating is 180 watts while on AM the input runs 75 watts.

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Optionally available accessories include the VX-2 VOX at $35.00, the external 210C VFO at $120.00. the external crystal oscillater type 510X at $45.00, and the newly announced mobile noise blanker which Swan is coming out with shortly. All of these items are provided for with accessory sockets and built-in wiring.

Swan offers its familiar AC console supply. model **117XC** at $105.00 and for a **DC** supply on 12 volts, the model 14-117 at $130.00.

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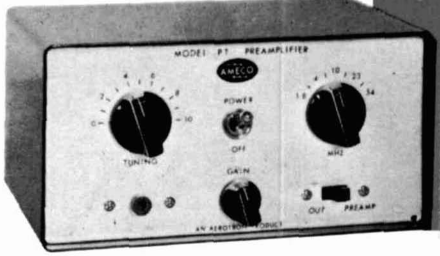

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