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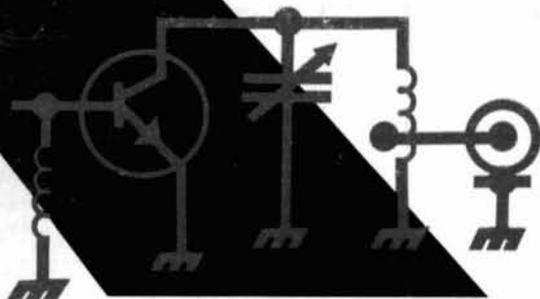
focus
on
communications
technology . . .

ham radio

OCTOBER 1968

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**solid-state
transmitter
for 6 meters**

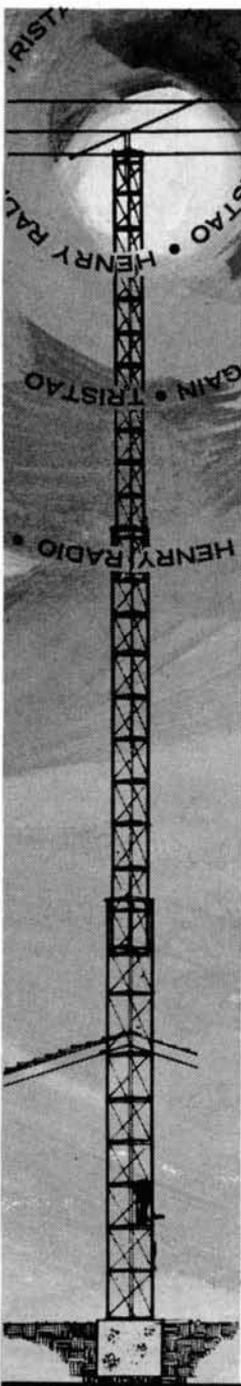


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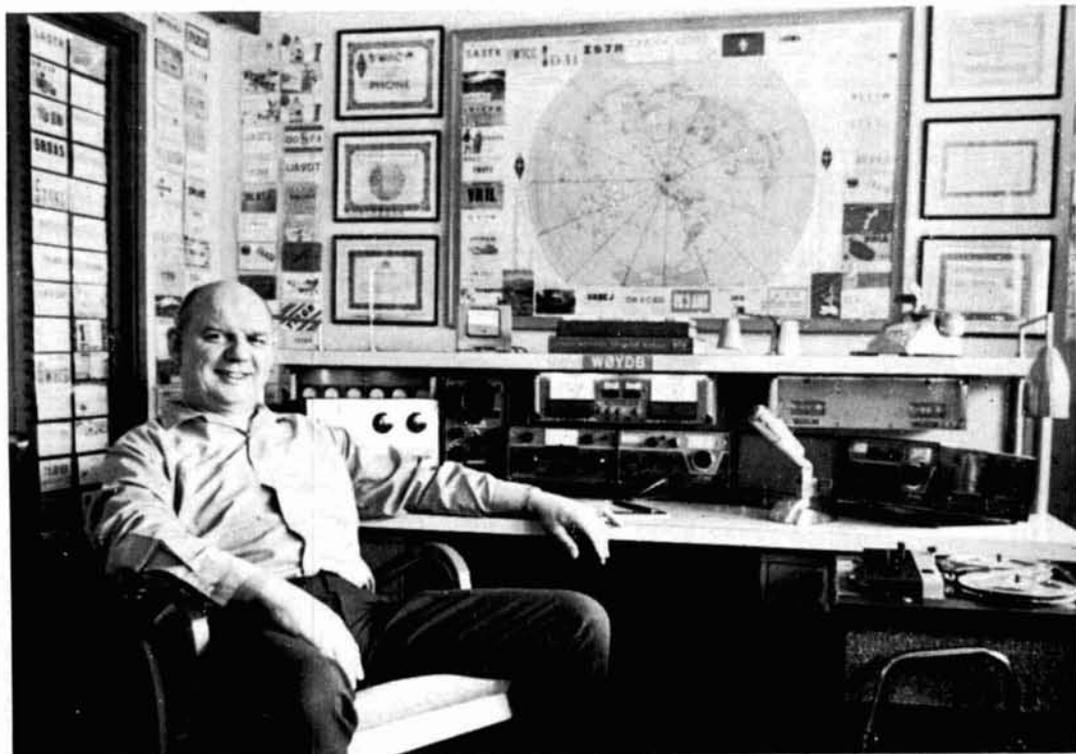
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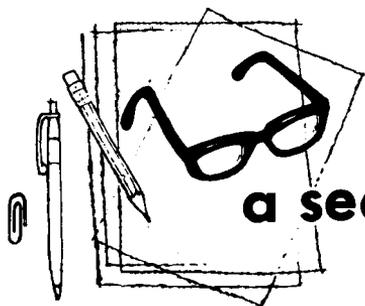
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a second look

by jim
fisk

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 state-side 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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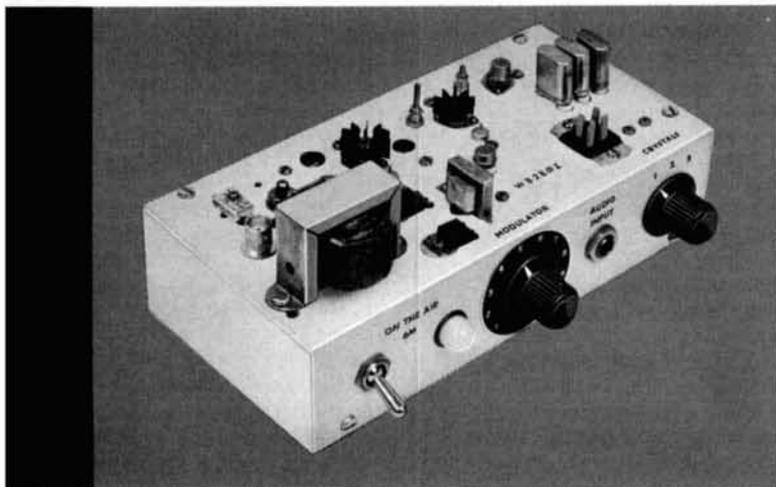
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higher power from transistors on six meters

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-meters³ prompted me to try higher power—this time on the six-meter band. Whether you choose this ap-

proach or something simpler, here are a few useful hints to help you along the road to success.

the circuit

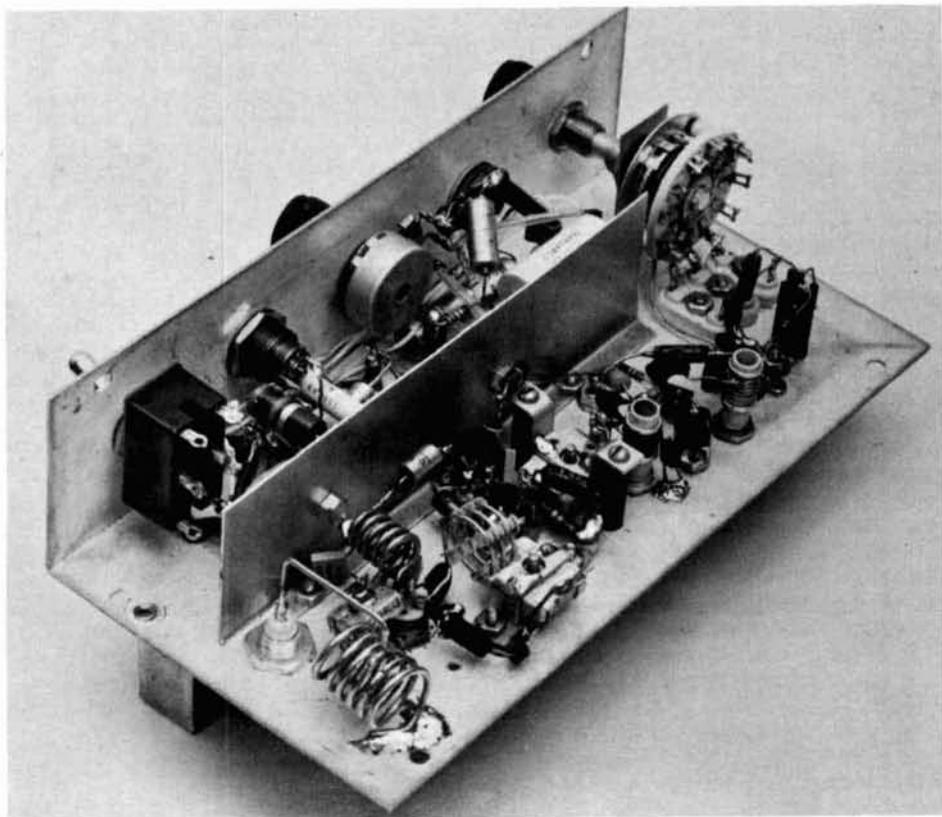
On a dollar-per-watt basis, transistors are still not competitive with tubes. In order to compensate, low-cost parts were used where possible without sacrificing performance. The oscillator stage exemplifies this philosophy. An RCA 40081 is used in a 50-MHz overtone crystal oscillator circuit. Low cost 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

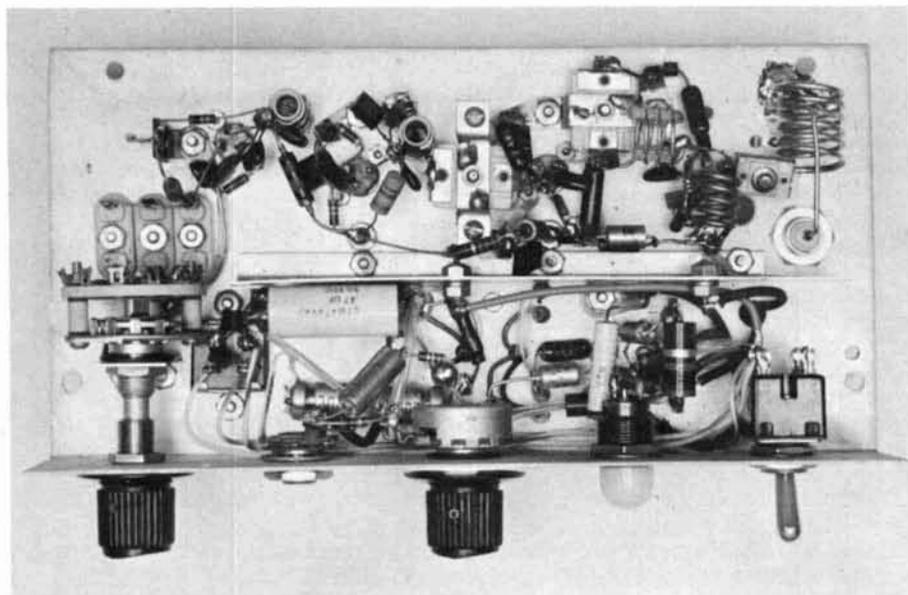


may be used in this stage will be discussed later. A 1/2-ohm resistor in the emitter limits current peaks. While this is degenerative and costs a little power output, it will prevent burning out the 2N2876 by excessive rf drive or certain regenerative phases which occur during tune-up. Unlike the base return, this

capacitor and the 0.05- μ F 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 μ F). 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 by-passing 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

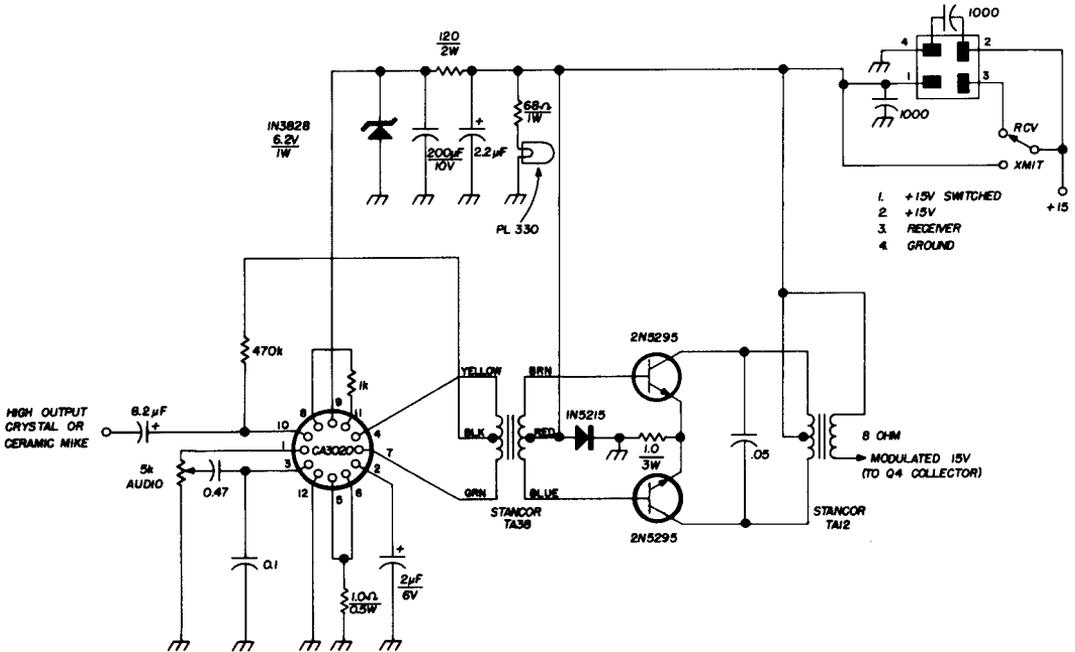


fig. 2. Five-watt modulator for the six-meter transmitter.

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 motor-boating 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.

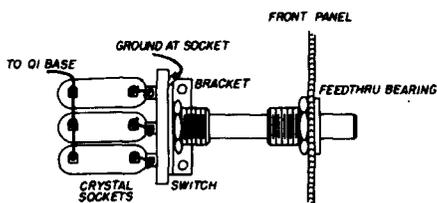
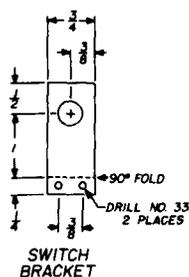


fig. 4. The crystal-switch bracket and rf shield are made from 0.040 aluminum.



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

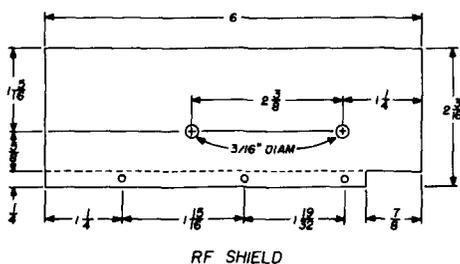
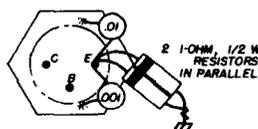


fig. 5. Wiring the rf power transistor. The emitter bypass capacitors are grounded directly to the transmitter case.



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

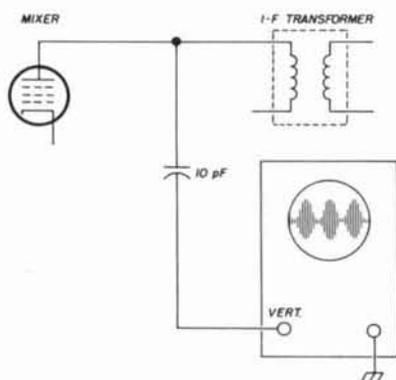
2. The 2N3375 is the most efficient, but not significantly different than the 2N2876. On a dollar-per-watt basis, the 2N3375 is also preferable. Why then, did I choose a 2N2876 for the transmitter? The primary reason is that the transistor is a lower frequency device and inherently more resistant to 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 persistent proponents of that device.

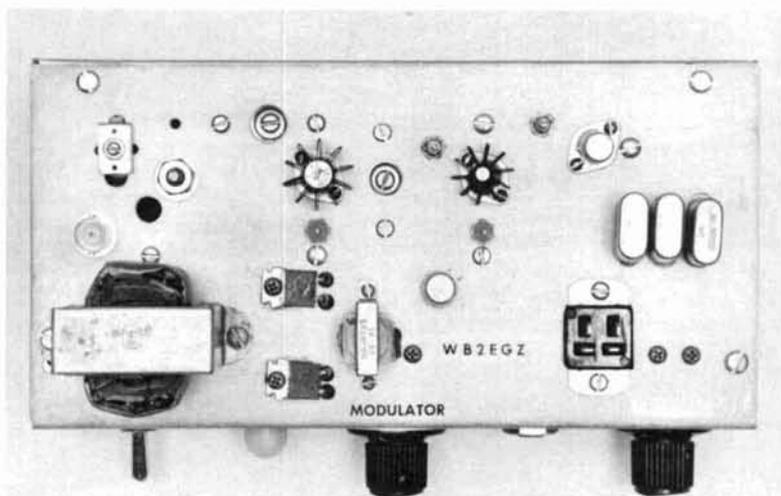
Some experiments were performed using different coupling techniques with efficiency and purity of output as objectives. Results showed a preference for tapped tank circuits over L's or π 's for rejecting unwanted har-

monics. 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.



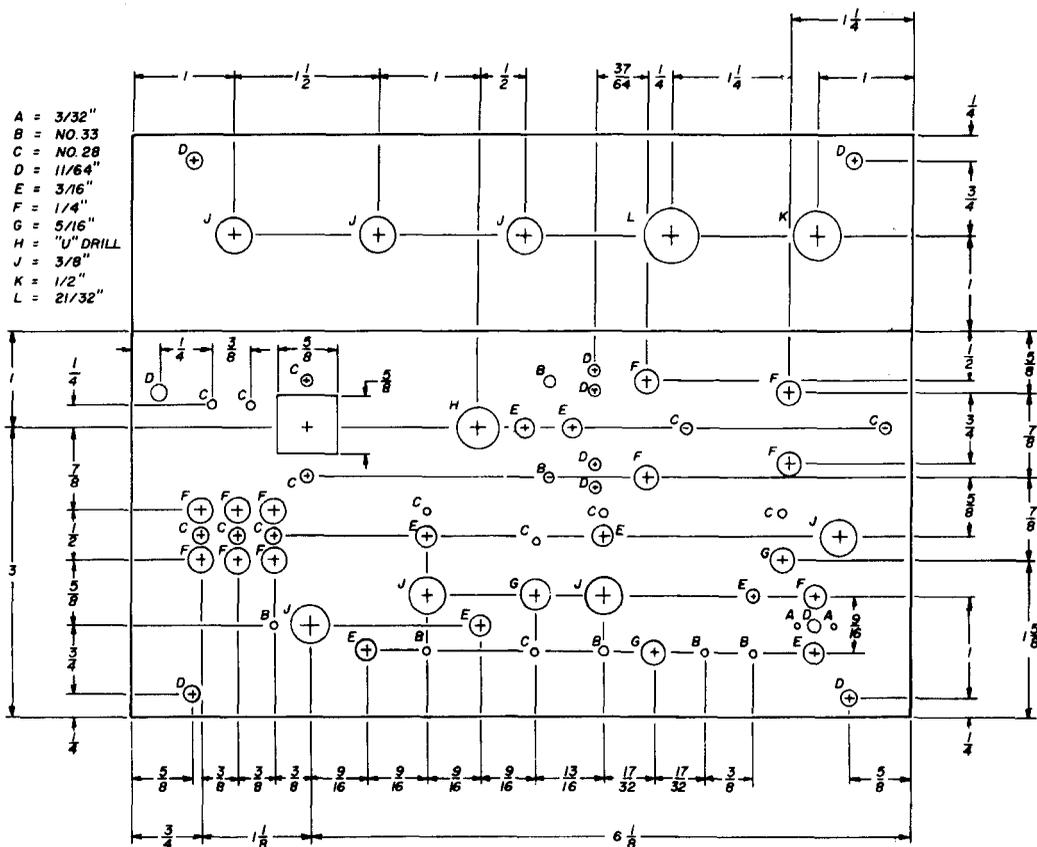


fig. 7. Chassis layout for the six-meter a-m transmitter. Make from 0.040 brass and silver plate. This chassis is designed for a Bud CU592 Converta box.

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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2. Edward P. Tilton, W1HDQ, "50-Mc Transistor Transceiver, Mark II," *QST*, February, 1967, p. 11.
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

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.

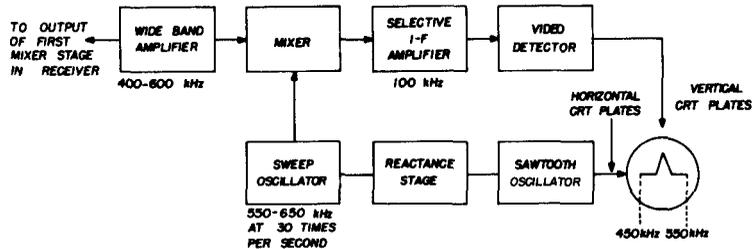
The same effect can be obtained if you have a separate receiver which is **manually** tuned through the i-f of the main receiver. The analogy to a manually-tuned receiver is worth-while because with this in mind it is easy to appreciate both the values and limitations of panoramic reception. As with 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).

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

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-swept, 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.



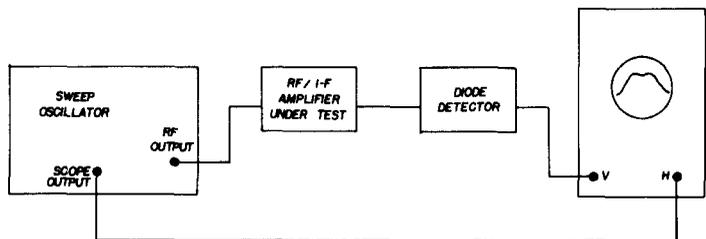
scan rate. Extreme selectivity and a slow scan rate (1 second, for instance) may be desirable if you are scanning a small frequency range—say ± 3 kHz—either side of the i-f. If you are scanning a larger frequency band (500 kHz, for instance) resolution must be sacrificed for using a scan rate of up to 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 simi-

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 pre-amplifier) 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

fig. 2. Method of using a sweep oscillator for checking or i-f amplifier response curves.



larity 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 $\pm 5\%$ 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

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.

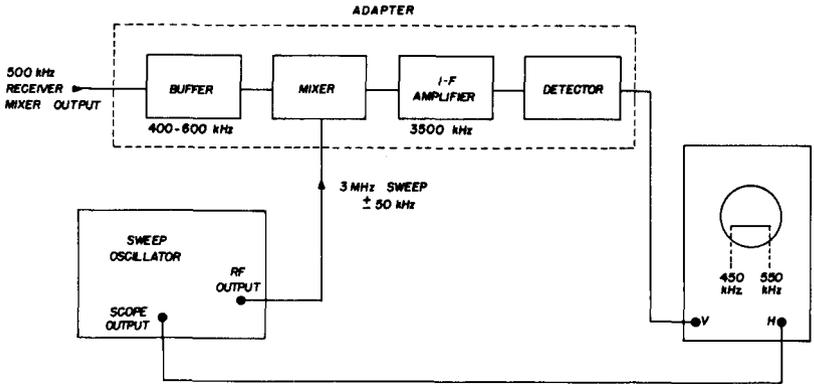
If you use care when selecting the injection frequency from the sweep oscillator, you'll find that a wide variety of auxiliary receivers

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.



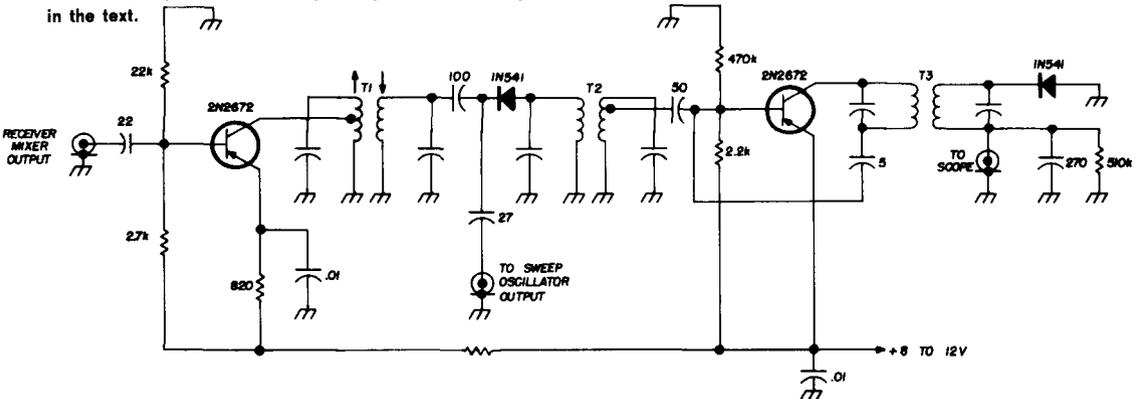
unit encompasses all the stage functions shown in fig. 3 and can be adapted over a wide frequency range. The 2N2672 input stage is coupled to the first mixer stage in the receiver. This is the signal that will be panoramically scanned. The input is untuned and presents an insignificant load-detuning 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.

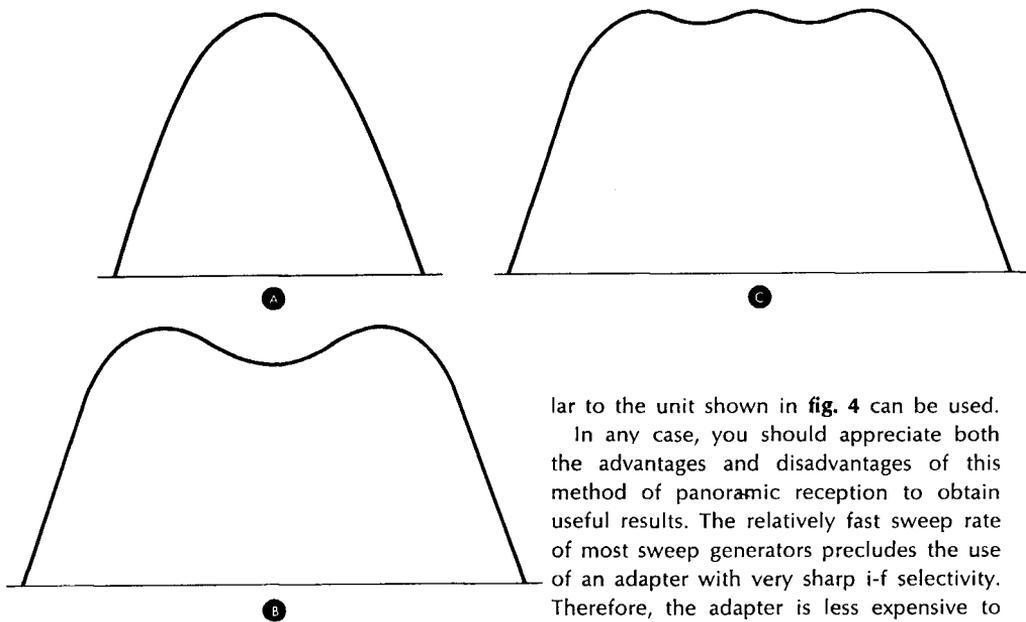


approximated in **fig. 5A**. The response is fairly broad since the selectivity is a result of the first stages of the i-f strip. The panoramic adapter must be connected at this point—not later in the i-f strip. Otherwise it will be impossible to scan anything but a frequency range restricted by i-f selectivity. If T1 is

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 simi-

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** does 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 alternately to the low and high limits of the scanned frequency range.

lar 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

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,

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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- μ F 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 mi-

crosecond, 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: 6.23×10^{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

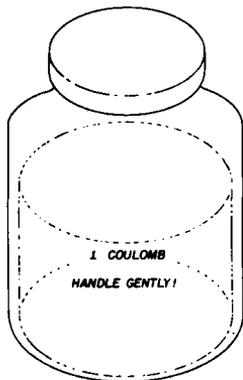
table 1. Electrical quantities and how they are built.

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: $R = E/I$. Di-

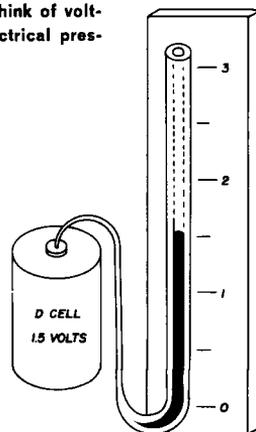
mensional analysis tells us ohms equals volts per amp, which we can put to work immediately. See **fig. 4**.

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 milliamp 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

fig. 2. Let's think of voltage as an electrical pressure.



joule sounds like a basic unit, but we can make it up as a volt-coulomb, volt-amp-second, amps²-ohm, etc. A joule will heat 4.186 grams of water 1-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

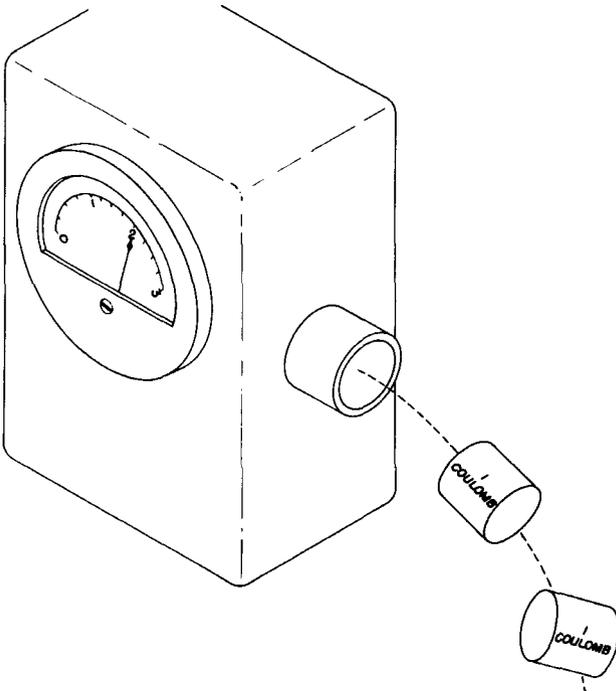
amount of energy, a standard unit of heat or of work. Stored energy, as in capacitors, is calculated in joules.

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.

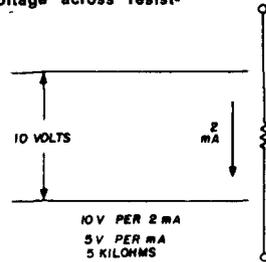
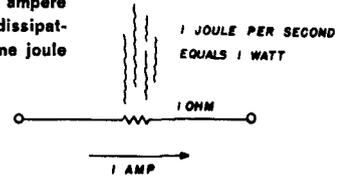


fig. 5. Here is an ampere through an ohm dissipating one watt, or one joule per second.



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.

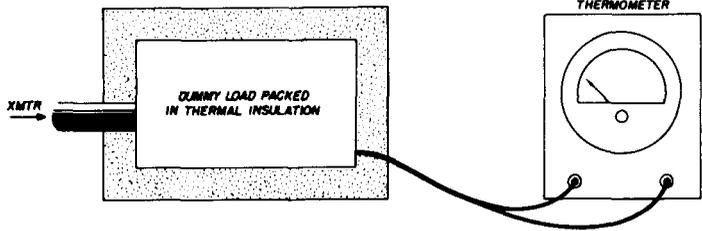
I've only touched on calorimetry here. There's a lot to it, which you can find in any basic physics book. A good calorimetric test setup will completely eliminate all uncertainties about rf voltage, current and power.

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 in-

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

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.



ductors, 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

a capacitor in a photoflash lamp project, once the lamp specs are known.

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 1-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.

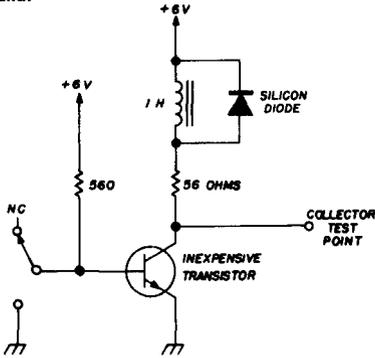
Let's think about inductive circuits that

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	$J = 1/2LI^2$ 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	$J = 1/2CE^2$ 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.

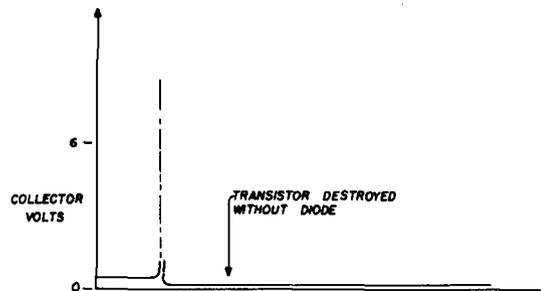
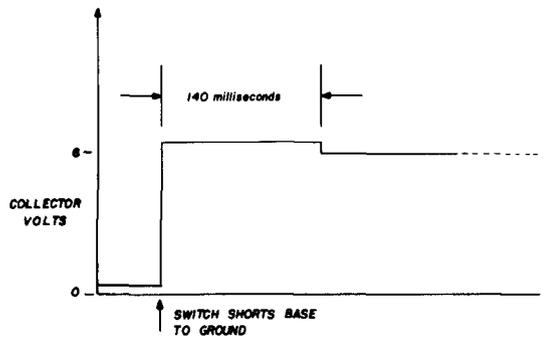
Now, we know the turn-on current, which was established by the resistor. But circuit-limited current means we don't have a volt-seconds figure to work with, as we might in a pulse circuit problem. So we restate the inductance in henries as volt-seconds per amp, and we know the circuit will somehow have 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 what 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—
vox and mox;
here's how
these systems work

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

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

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

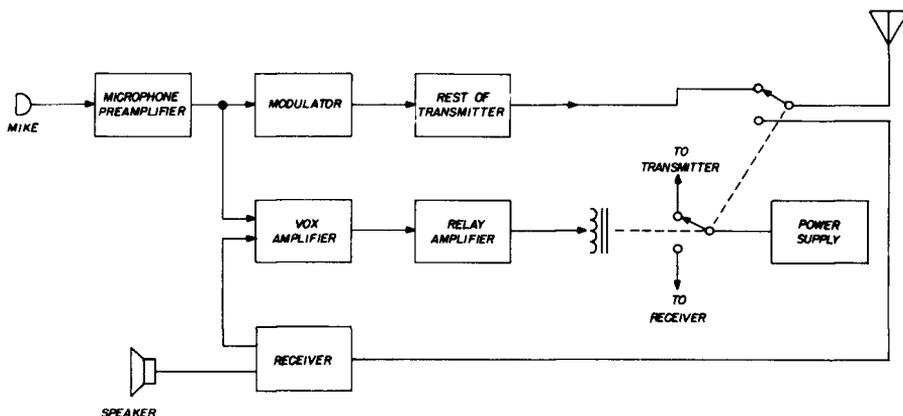
In the block diagram of **fig. 1**, you'll see how the typical vox circuit operates. Incoming microphone audio is taken from the out-

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**.

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 V1, the vox amplifier. (C4 bypasses any rf which

fig. 1. Basic vox circuit used in ssb equipment. The anti-trip circuitry that keeps receiver audio from triggering a switchover is also shown.



put 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

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.

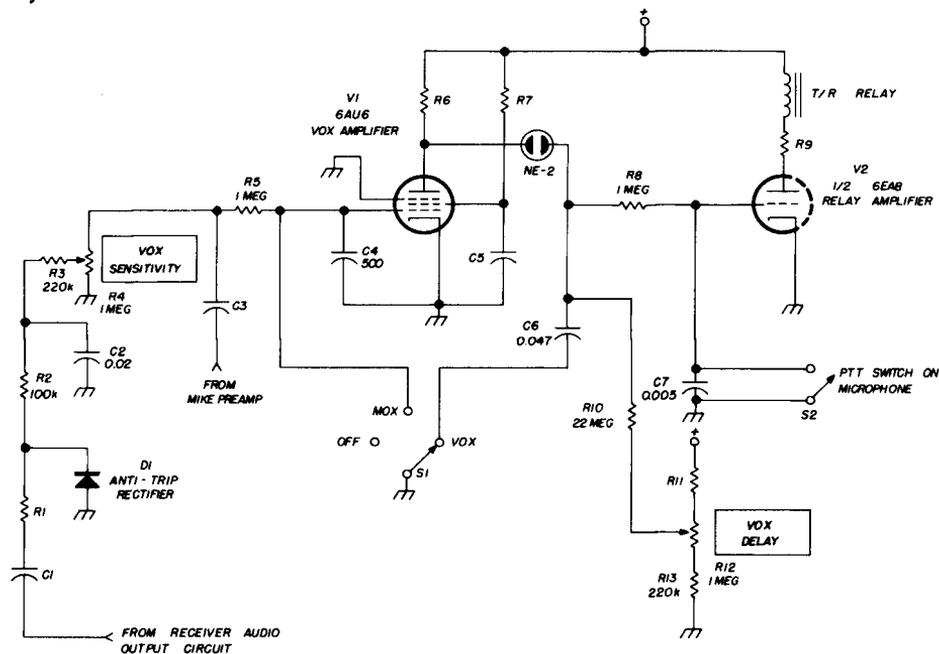
Negative half-cycles, however, drive down the too-positive grid bias of VI, reducing plate current and causing plate voltage to increase during each half-cycle. When the plate voltage gets high enough, the neon lamp fires, applying a dc voltage across R10, control R12, R13, and capacitor C6. C6 charges, 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 V1 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 a 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, V1 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 V1. 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 V1 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 **vox** position, grounding the grid of V1 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 key-

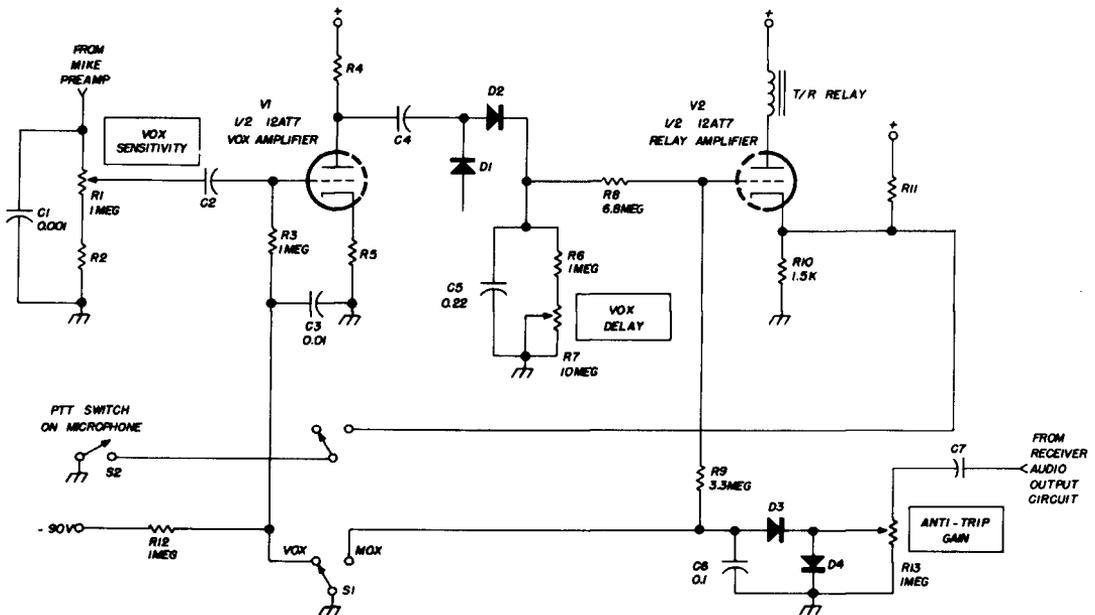
age doubler D1-D2 rectifies the amplified voice signals to produce a positive dc voltage that is fed to the grid of V2, the relay amplifier.

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

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.



ing 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 V1. In the V1 plate circuit, volt-

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

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 V1. When function switch S1 is in the **vox** position, the bottom of R3 is grounded; the -90 volts does not affect V1. But when S1 is thrown to the **mox** position, the -90 volts biases V1 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.

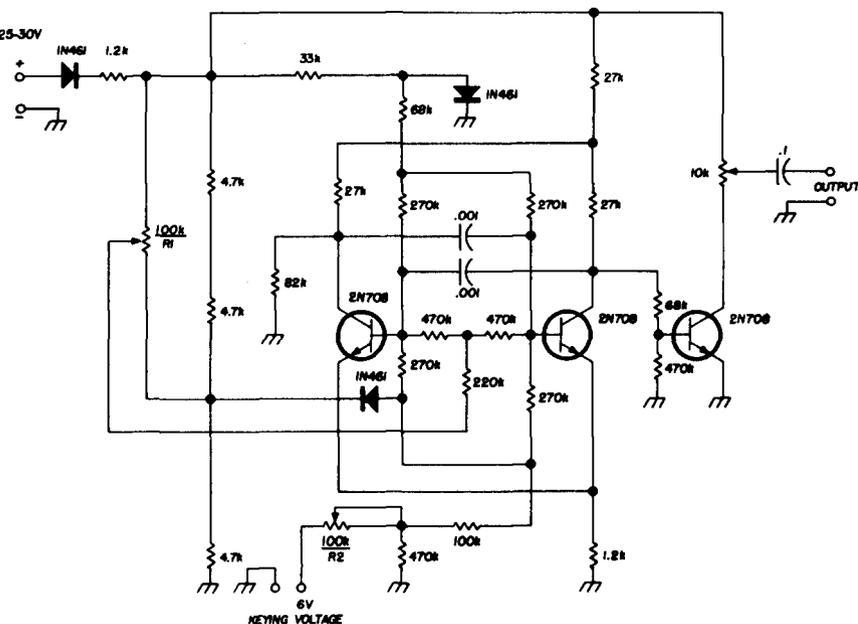
ham radio

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

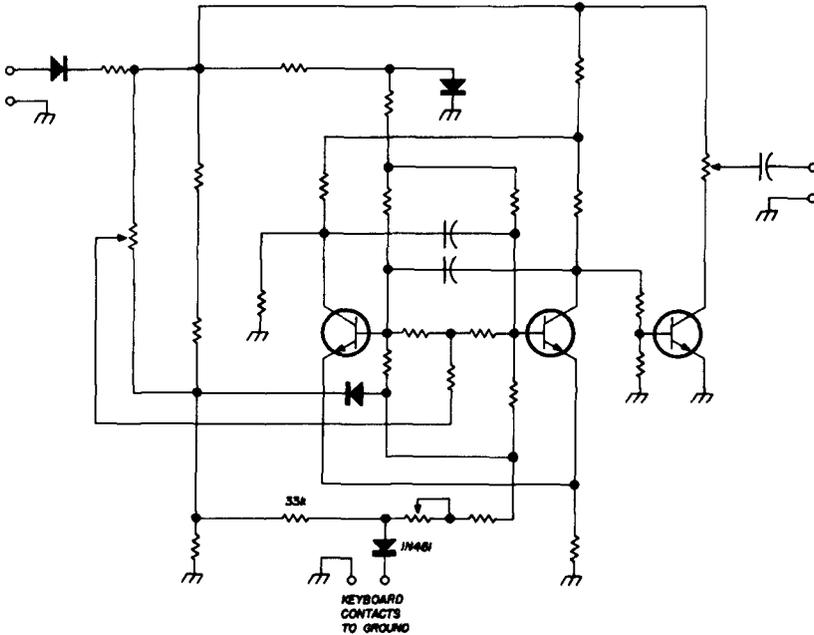
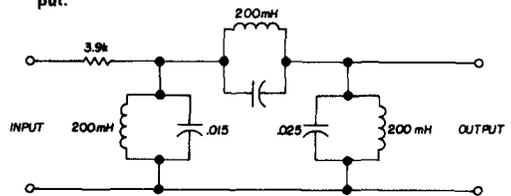


fig. 2. Afsk oscillator designed for keying to ground. Except for the parts that have values, this circuit is identical to the one in **fig. 1**.

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.

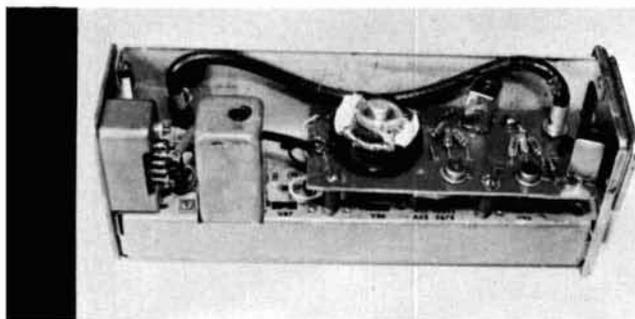
For easy adjustment, use ten-turn potentiometers for R1 and R2. The output is a square wave. If you want a sine wave output, use a band-pass filter as shown in **fig. 3**. The center of this filter is set at 2424 Hz, and is 4 1/2-dB down at 2125 and 2975 Hz. At the 20-dB points, the pass band is 1500 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

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 I 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

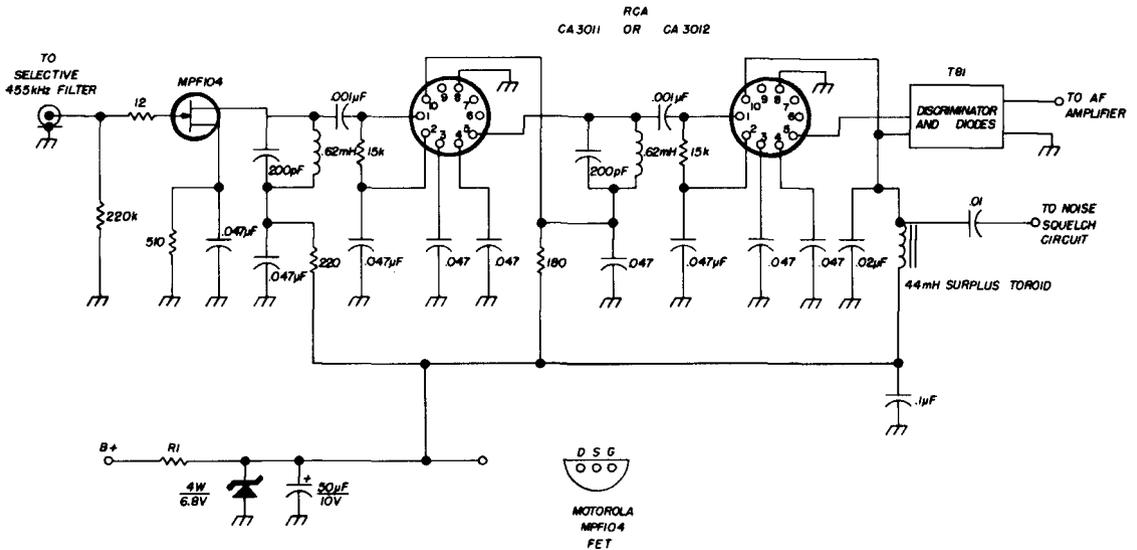
■ Frank Jones, W6AJF, 850 Donner Avenue, Sonoma, California

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.

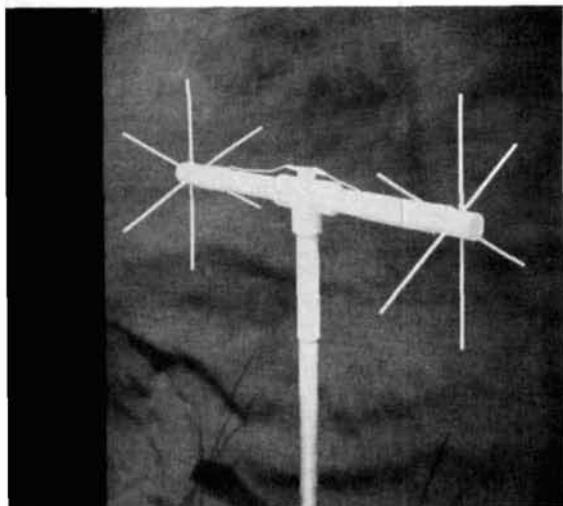
construction

The two IC's and some of the other parts were mounted on a piece of bakelite board about 4 x 1 1/2 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 narrow-band 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.

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 using fewer 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

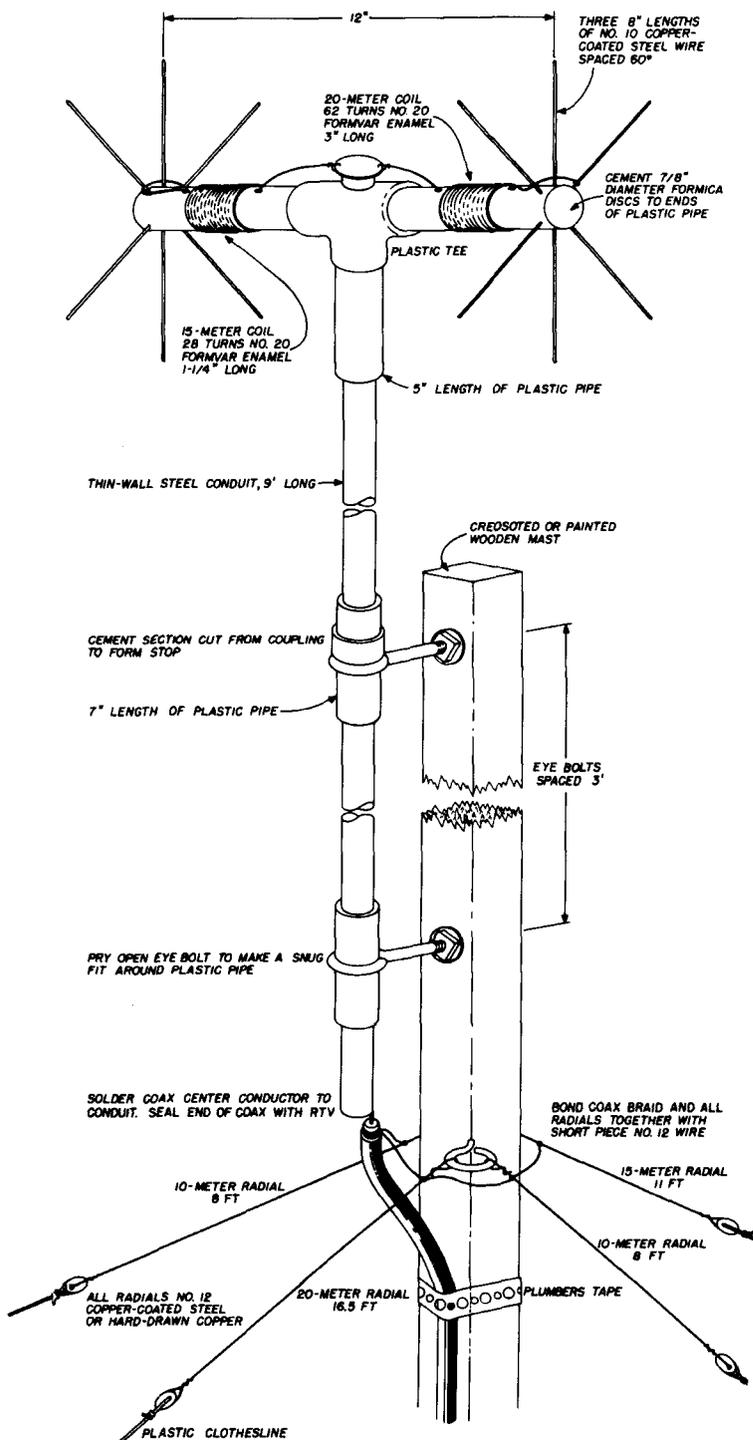
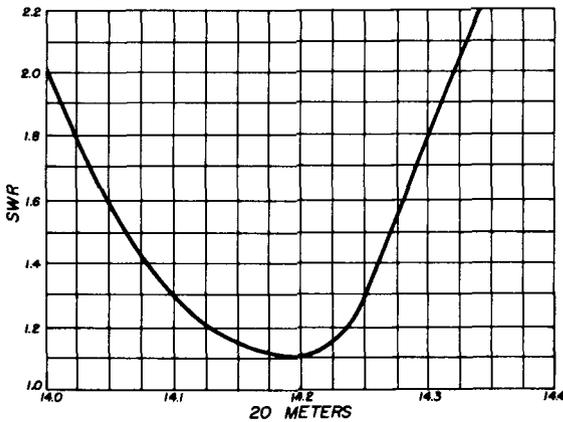


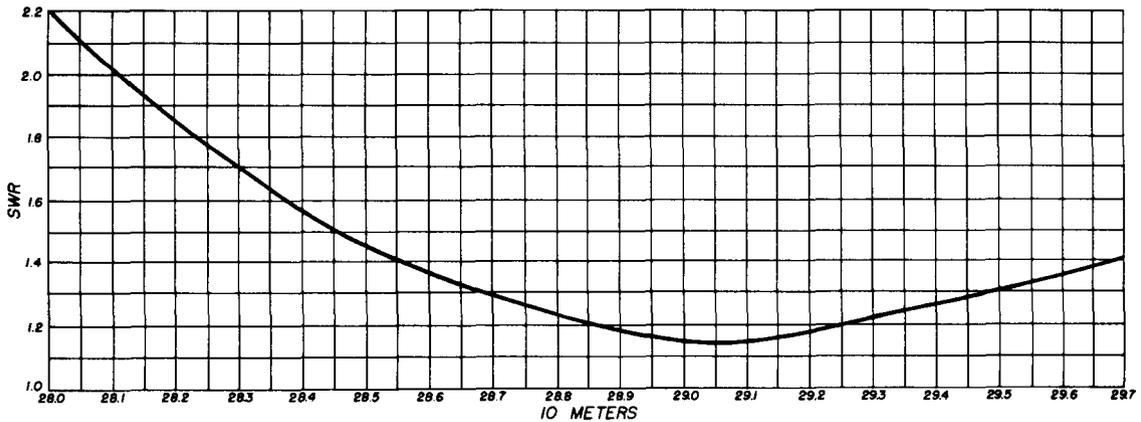
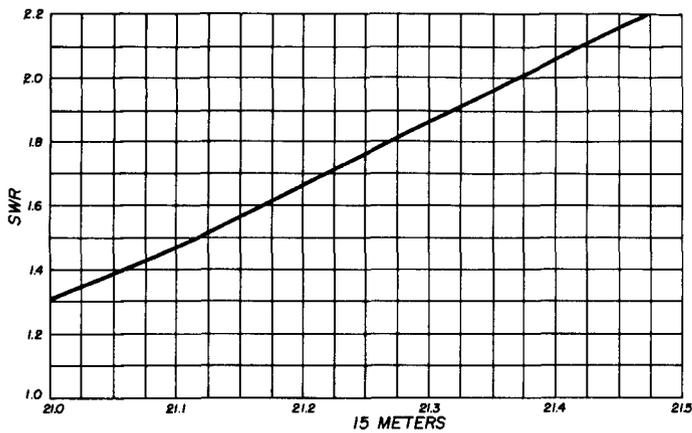
fig. 1. Construction details of the 3-band ground plane. The top tee is bored out so the end of the conduit sticks through; then a copper plate is soldered to the conduit for connecting the loading coils. Coil leads must be soldered to all three spikes.

fig. 2. Swr performance of the 3-band ground plane.



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



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

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



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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

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**.

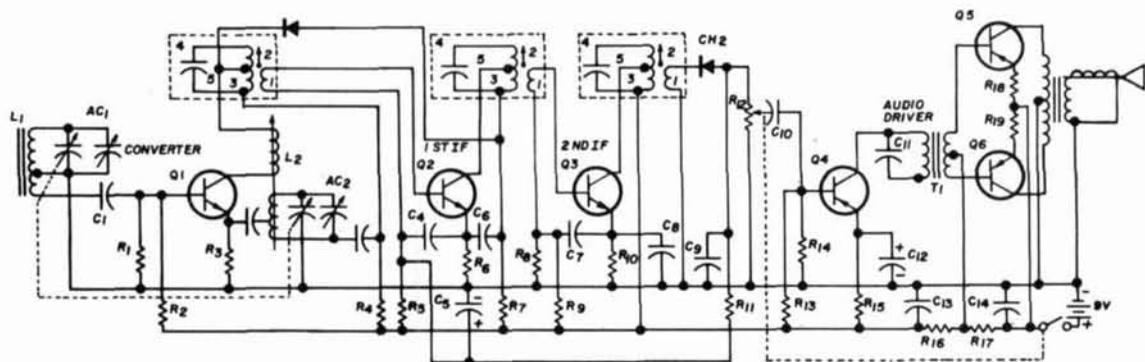
Scheduled weather broadcasts are made twice each hour: at 15 and 45 minutes past the hour. At 15 minutes past the hour, current weather reports for eight to twelve locations within 150 miles of the broadcasting station are transmitted. At 45 minutes past the hour, similar reports are broadcast for important 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

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

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 re-

ports 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

AC2 90-pF variable capacitor with trimmer ganged with AC1

C1 .02 μ F

C2, .01 μ F

C3

R1 6.8k, 1/4 watt

R2 27k, 1/4 watt

R3 1.5k, 1/4 watt

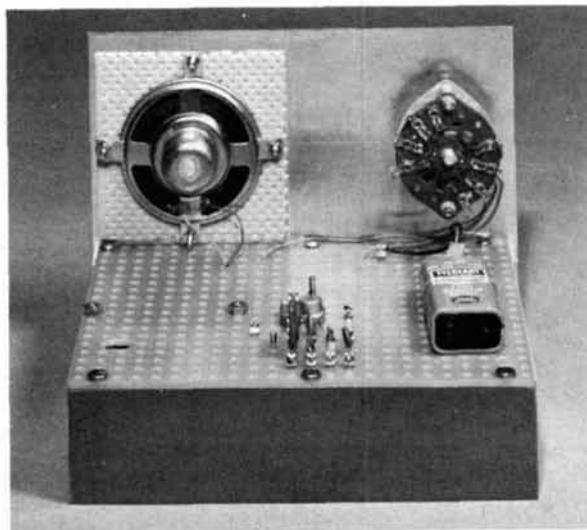
R4 470

L1 435 μ H loopstick antenna

Q1 2N194A

fig. 1. Typical six-transistor broadcast receiver.

Top view of the modified broadcast receiver 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 radios 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 receiver to cover 362-kHz FAA weather broadcasts.

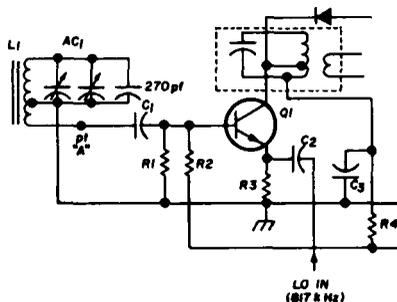
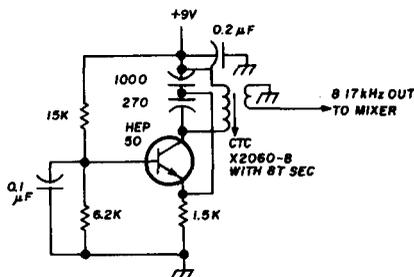


fig. 3. New oscillator stage for the modified broadcast receiver.



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.

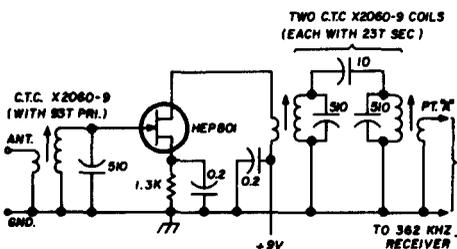
By removing the oscillator coil (L2) from the circuit, the autodyne can easily be converted into a simple grounded-emitter mixer as shown in fig. 2. Local-oscillator injection must then be furnished by a separate oscillator. The local-oscillator drive should preferably come from a small link on the oscillator 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.

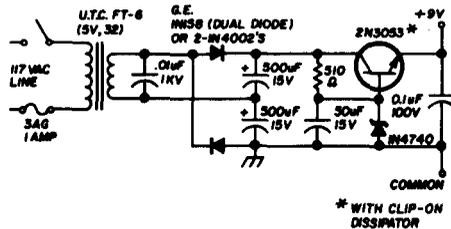
A word about old transistor BC radios is in order. The majority are discarded (and available inexpensively) because of one of two electrical faults: autodyne trouble, or electrolytic capacitor trouble, or both. Since this modification calls for destruction of autodyne action with a separate local oscillator, 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 sta-

tion (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
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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

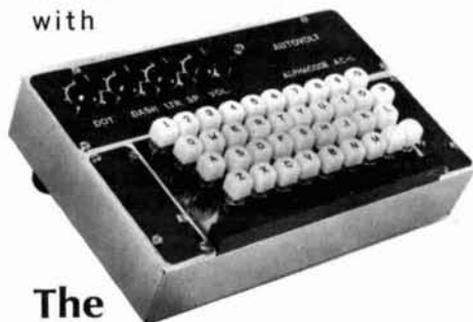
Davenport	224	S
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Harrisburg	338	S
Pittsburgh	254	C
South Carolina		
Charleston	329	C
Spartanburg	248	C
South Dakota		
Huron	391	S
Pierre	347	S
Rapid City	254	C
Sioux Falls	245	S
Tennessee		
Knoxville	257	C
Memphis	371	S
Nashville	304	C
Texas		
Amarillo	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

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C Continuous broadcasts

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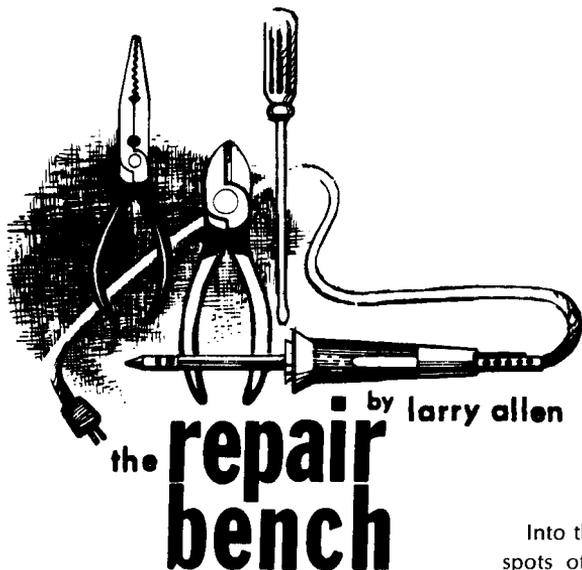
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ham radio



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.

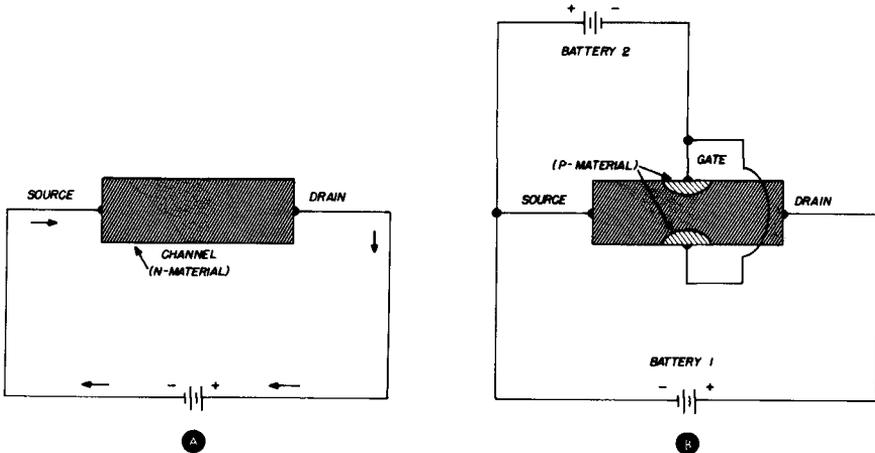
In a JFET, there may be slight leakage

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

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

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.



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.

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).

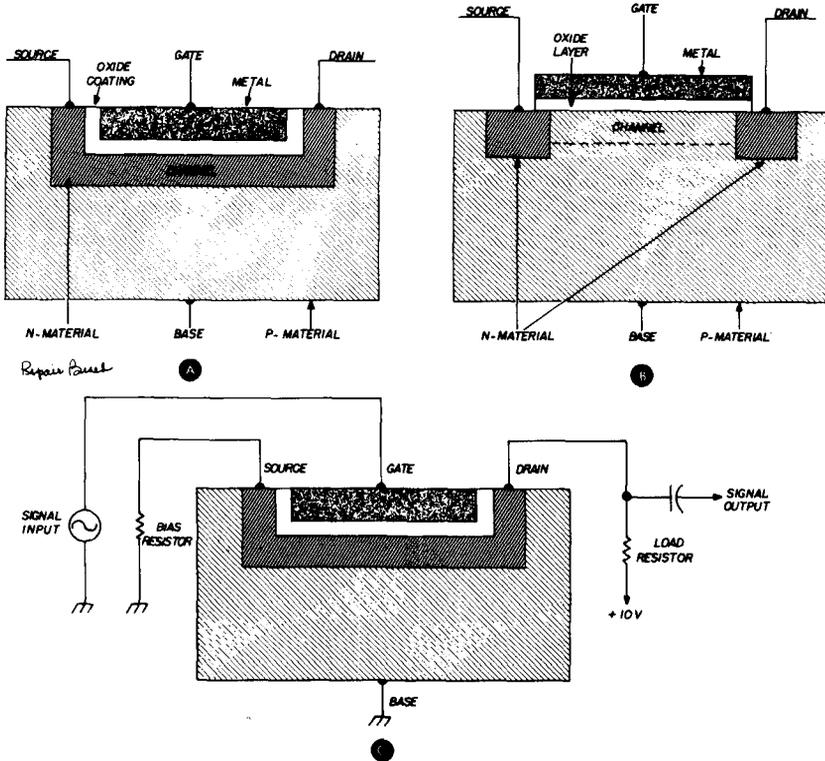
bipolar—unipolar

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

the carriers are holes. FET's are therefore said to be **unipolar**.

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

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.



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 FET's, 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 FET's, the carriers are electrons; in P-channel types,

reference—that bipolar transistors are the ordinary kind and that unipolar transistors are a family to which FET's belong.

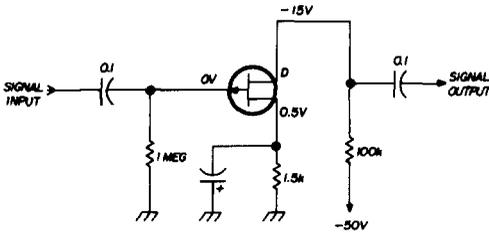
what happens to fet's?

Since FET's 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 re-

sult. 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 experiment-

ing with FET's, pay attention to the voltage ratings. Next most detrimental is overcurrent. Even slight overcurrent may eventually overheat a JFET junction, resulting in permanent damage.

FET's 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 FET's 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

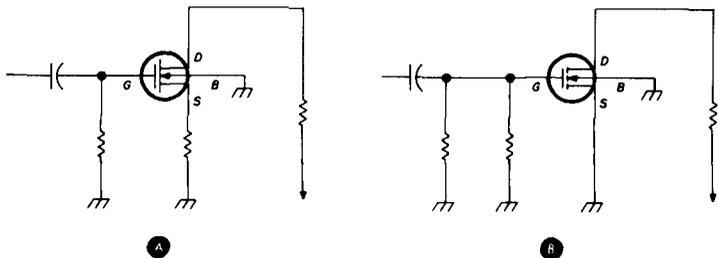
There are three ways to track down a FET that has given up. One is by signal injection or signal tracing—methods I discussed in the April and May 1968 issues. You simply pin down the faulty stage and then test it to see if the FET in that stage is at fault.

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 FET's 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; FET's 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 fed from regular dc supply.



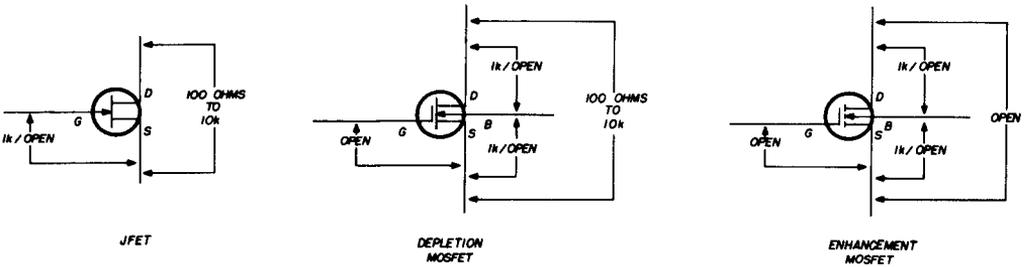
channel JFET (you can tell it's P-channel because the arrow points outward). Gate bias is developed in the source circuit, across the 1.5k resistor.

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

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 current 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.

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.



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.

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.

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.

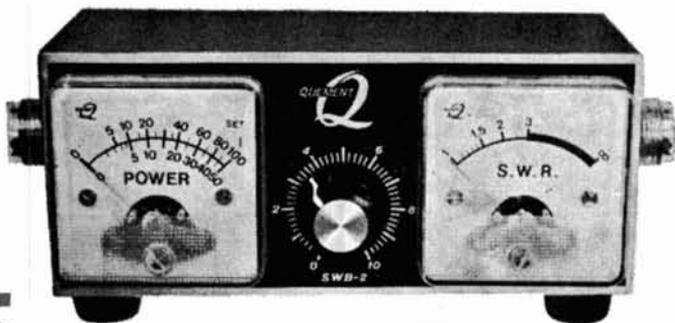
Diagrams that will help you check different FET types are given in fig. 5. The readings you should get are indicated clearly. Each type has its own peculiarities.

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

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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 1k 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 1k, 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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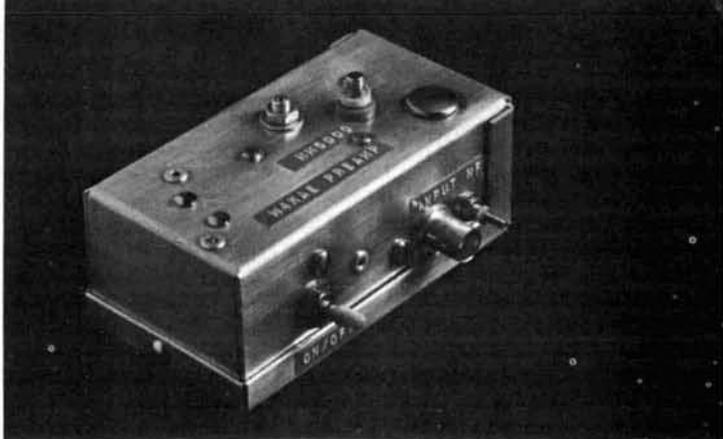


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MM5000 2-meter preamp

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 transistors 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 burn-out 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.

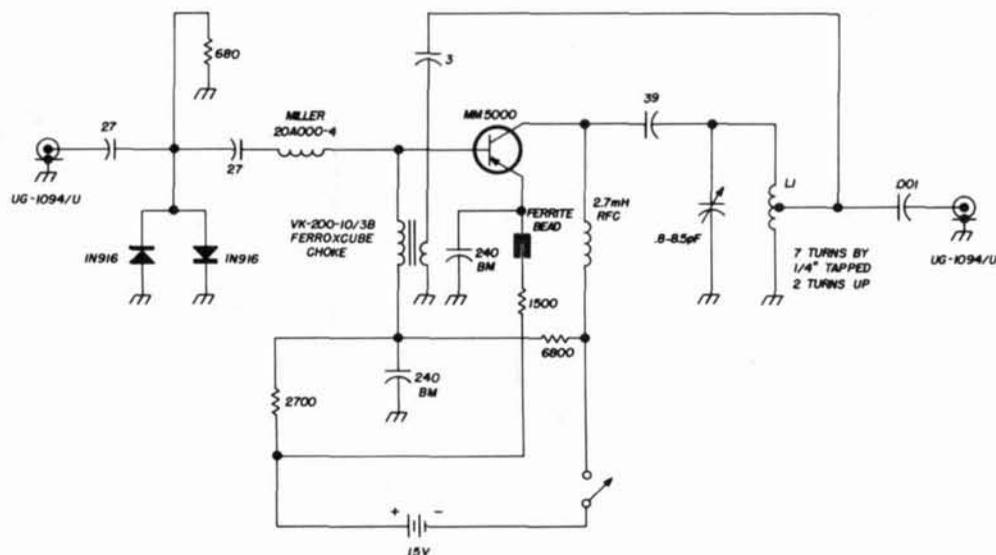


fig. 1. Schematic diagram of the low-noise two-meter preamp. Capacitors marked "BM" are button micas.

construction

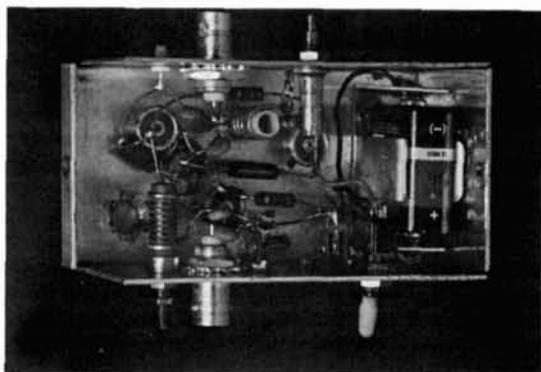
The MM5000 two-meter preamp is constructed in an LMB-421 flange-lock box chassis. It has what might be considered to be a floating ground although it's heavily bypassed to the chassis with button capacitors. A glass piston capacitor is used to tune the output tank. A series NF adjustment coil is provided 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

Internal construction of the preamp.
The series NF coil is in the lower left next to the input coaxial connector.



estimated 1.2 dB for the MM5000—both maximum values at 150 MHz.

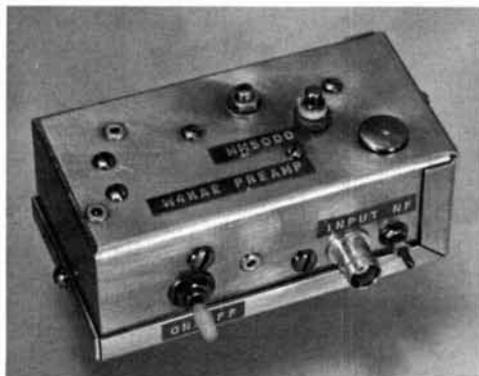
I found that noise figure is best adjusted through the use of series inductance tuned for a dip in the input signal or noise. Of course, the output tank circuit was previously set for maximum gain. I observed that

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 an-

The complete preamp is housed in a small LMB flange-lock chassis.



tenna can provide perfectly satisfactory reception. Burn-out has been a myth with 150 watts output to a flat line.

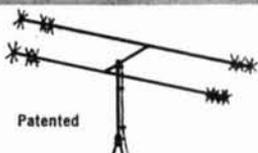
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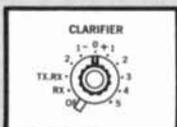
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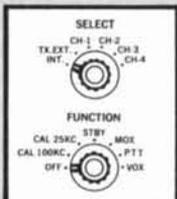
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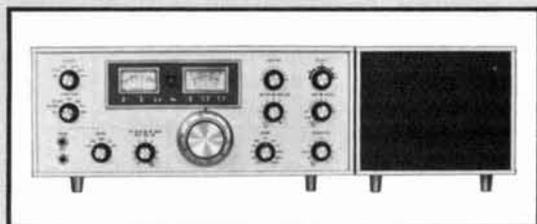


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During recent years, amateur equipment has improved to the point where 1-kHz frequency accuracy is expected, and may well be essential for many reasons. Several receivers, exciters and transceivers are available which have a specified dial accuracy 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 1-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

E. H. Conklin, K6KA, Box 1, La Canada, California 91011

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 1000-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 1-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

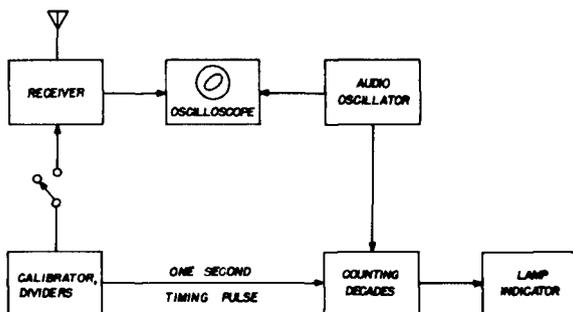
you avoid 3.5-kHz signals and concentrate on the higher bands where the percentages are in your favor.

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

fig. 1. Basic circuit used by K6KA for making amateur frequency measurements.



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.

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.



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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 1-kHz decade which is its 30,000th harmonic! So, no problems here either.

Note that I haven't used the word "multi-vibrator." 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 multi-vibrator. 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 $2^{1/2}$ 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 Fremont 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.

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designing single-tuned interstage networks

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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. C₁ provides dc isolation between the collector of Q1 and base of Q2. The coil carries bias current 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.

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 Rb1 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:

$$N_1^2 R_1 = N_2^2 R_2 \quad (1)$$

Two Q's should also be defined for future use. Qo is the natural unloaded Q of the coil itself:

$$Q_o = R_o / 6.28 f_o L \quad (2)$$

Where f_o is the resonant frequency in hertz. The other Q is the operating or loaded Q which is called Q_L:

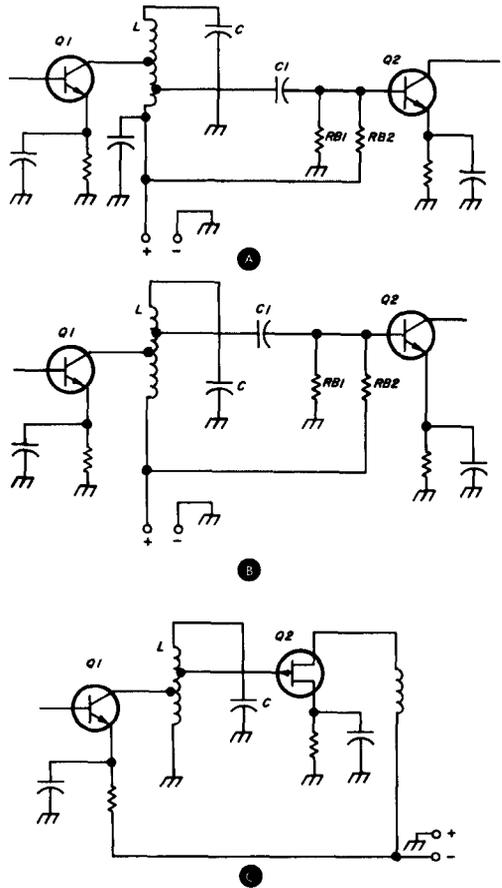
$$Q_L = \frac{\text{parallel combination of } R_o, N_1^2 R_1 \text{ and } N_2^2 R_2}{6.28 f_o L} \quad (3)$$

$$Q = 1 / (6.28 f_o L [1/R_o + 1/N_1^2 R_1 + 1/N_2^2 R_2]) \quad (3a)$$

the design problem

When starting an interstage network design problem, you'll know six quantities: R1, C1, R2, C2, f_o and P_{av}. The first five quantities were defined before; P_{av} is the power available from the driving stage. The other eight quantities, N1, N2, L, Q_o, Q_L, C_x, C_a and P_t have to be determined. P_t is the power transferred to the driven stage.

fig. 1. Three examples of the tapped-coil interstage network.



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-

cally in **fig. 4**. Mathematically:

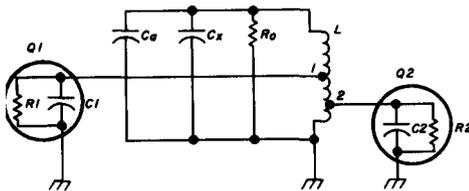
$$P_t/P_{av} = (1 - Q_L/Q_0)^2 \quad (4)$$

From **fig. 4** it can be seen that the loaded Q (Q_L) is always less than the unloaded Q (Q_0). 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

fig. 2. Equivalent circuit of the three circuits shown in fig. 1. The positions of taps 1 and 2 determine N_1 and N_2 , respectively.



use slug-tuned coils, or, the median capacitance of the tuning capacitor.

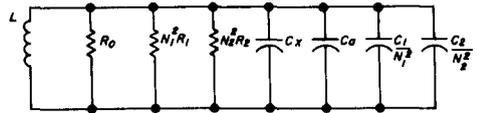
With these four quantities in mind, you can use **fig. 4** to find the required unloaded Q of the coil. Inductor manufacturers usually provide this information in their catalogs or you can measure it with a Q meter. Now the inductance of the coil can be calculated from the following formula:

$$L = \frac{2Q}{78.88} \frac{Q_0 - R_1 6.28 f_0 (Q_0 - Q_L) (C_1 + C_2 R_2/R_1)}{Q_0 f_0^2 (C_x + C_a)} \quad (5)$$

Don't be scared by the arithmetic—it is only simple addition, 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 f_0 in hertz. The values of Q and N are unitless.

If the inductance calculated from **eq. 5** results in an unwieldy value, you may want

fig. 3. Simplified equivalent circuit.



to adjust Q_L , P_t or C_a slightly.

The next step is to calculate the turns ratio, N_1 :

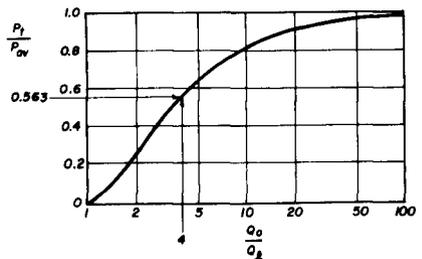
$$N_1 = \sqrt{\frac{12.56 Q_L Q_0 f_0 L}{R_1 (Q_0 - Q_L)}} \quad (6)$$

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 N_2 from:

$$N_2 = N_1 \sqrt{\frac{R_1}{R_2}} \quad (7)$$

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 C_a . P_t , Q_L and C_x are known and Q_o is obtained from **fig. 4** as before. The inductance L is found from:

$$L = \frac{N1^2 R1 (Q_o - Q_L)}{12.56 f_o Q_L Q_o} \quad (8)$$

Then C_a is calculated from:

$$C_a = \frac{1}{39.44 f_o^2 L} - \left\{ C_x + \frac{1}{N1^2} \left[C1 + C2 (R2/R1) \right] \right\} \quad (9)$$

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 Q_o along with C_a and an estimated C_x . Now you can determine Q_L and P_t from **fig. 4**. This will require a compromise since you can't have high P_t 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 C_a . The loaded Q and P_t are determined as in method three; L , C_a 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, $C1 = 2 \times 10^{-12}$ F, $R2 = 2000$ ohms, $C2 = 5 \times 10^{-12}$ F, $P_{av} = 2 \times 10^{-3}$ W and $f_o = 144 \times 10^6$ 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 C_x 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 P_t 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 $P_t/P_{av} = 0.563$. Therefore, P_t , the input power to $Q2$, is 0.563×2 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. C_a 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

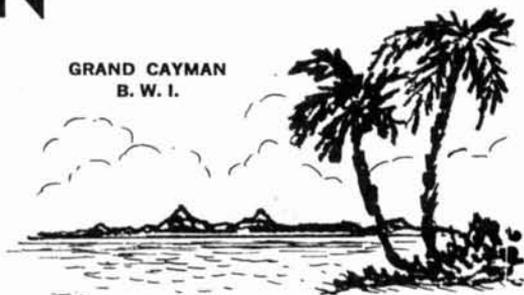
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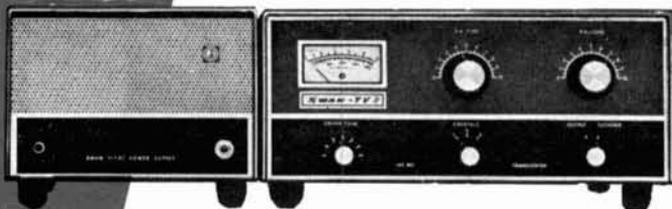
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Cliff Watson, K6WC, pounding the key at the Dewey Mine Station near Grangerville, Idaho in 1906.



early wireless stations

Here is
a brief history
of early wireless
and some
of the amateurs
who lived it

Ed Marriner, W6BLZ, 528 Colima Street, La Jolla, California

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 de-

tector 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

Marconi 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 when 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 telephones.

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.

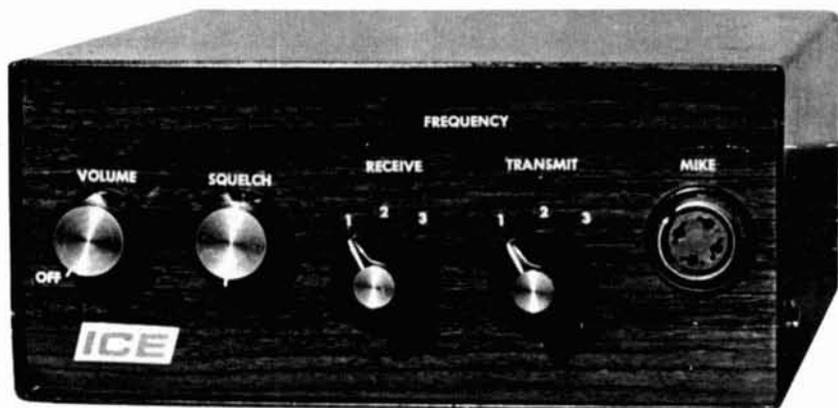
The first International Wireless Conference was held in 1903. At this conference, CQD was added to the operators' signals for distress; however, the Germans continued to use SOE. The New York Navy Yard had a wireless school established with 13 students. DeForest went to England to demonstrate high speed 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.



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ICE

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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 O-2 in Portland, S-2 in Seattle, or take a

Syd Fass, W6NZ, on the left, and W7QY, with an unknown operator on board the SS General Lee near KPH about 1912.



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.

ham radio

next month in ham radio magazine:

Transistor Transmitter

Six-meter transverter

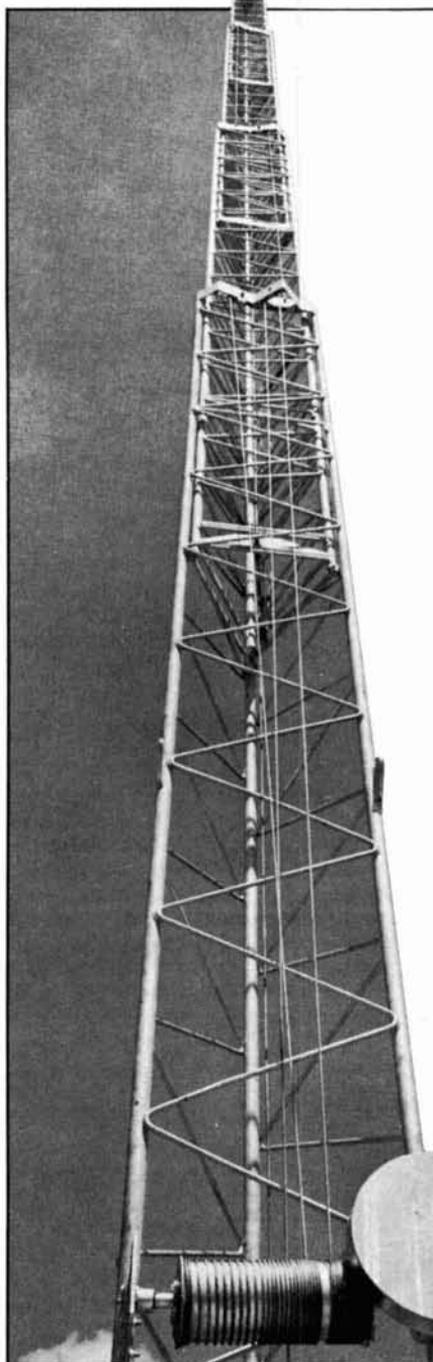
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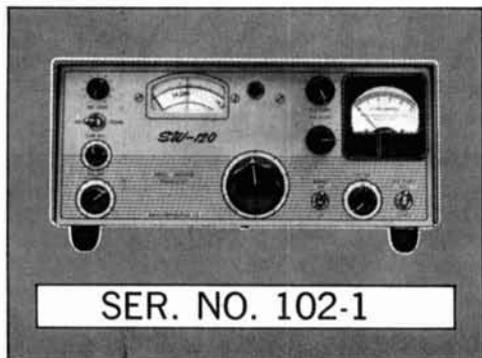
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REWARD: One new Swan 500C Transceiver with 117-XC power supply.

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.



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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

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

fig. 2. Time chart of predicted muf for October 1968 for 90° W longitude.

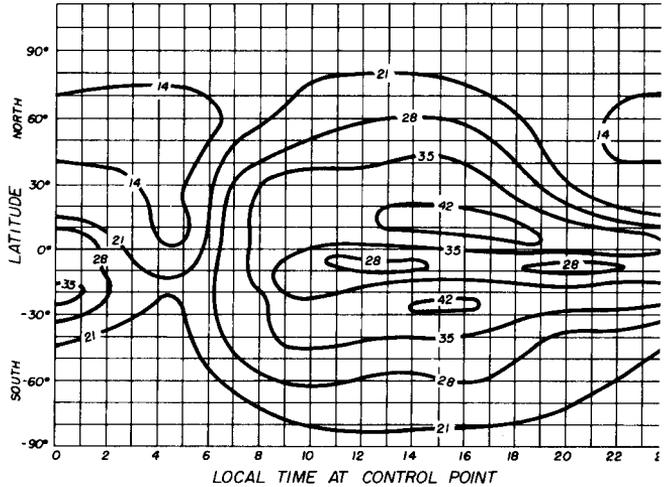
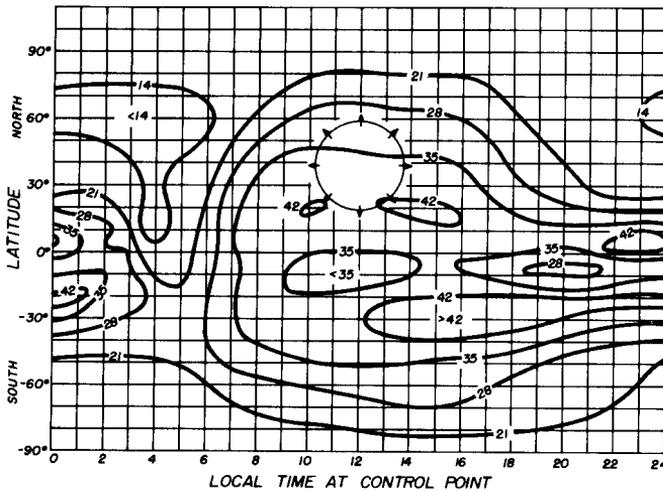


fig. 1. Time chart of predicted median muf for October 1968 for 45° W longitude.

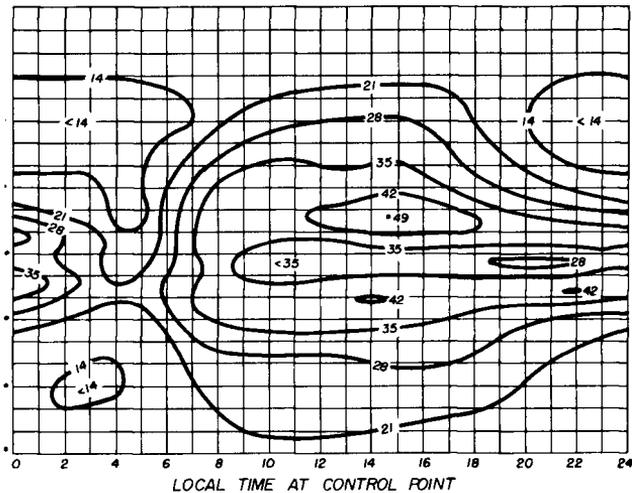


MUF charts to show the variation of MUF with longitude. **Fig. 1** is a time chart of the median MUF derived from ITS predictions for October 1968 for 45° W longitude; **fig. 2** is for 90° W and **fig. 3** is for 135° W. **Fig. 1** is the more accurate chart for the East Coast

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

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,

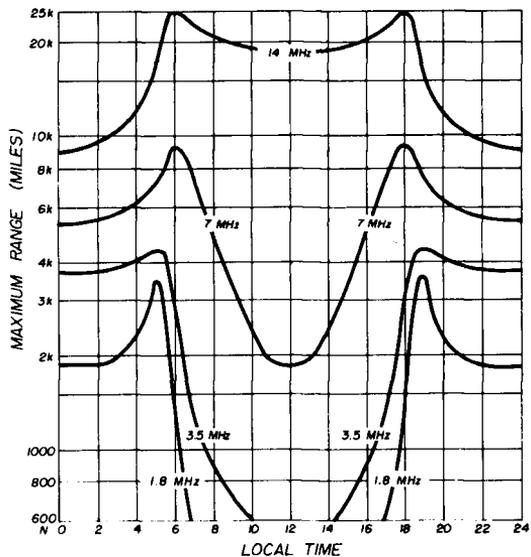


fig. 4. Maximum range vs local time for propagation to the North from midlatitude United States (38° N).

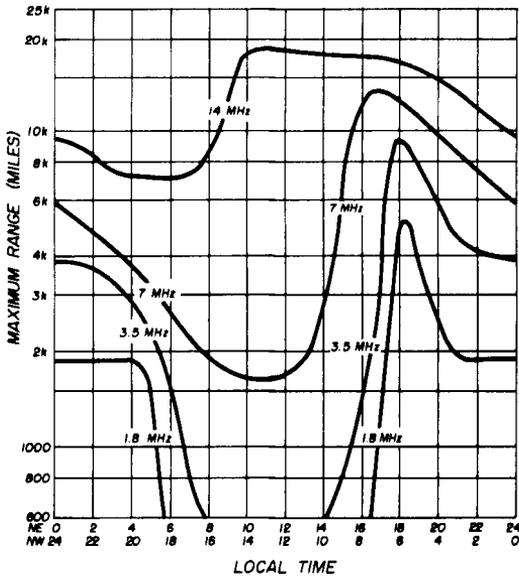


fig. 5. Maximum range vs local time for propagation to the Northeast (top time scale) and the Northwest (bottom time scale).

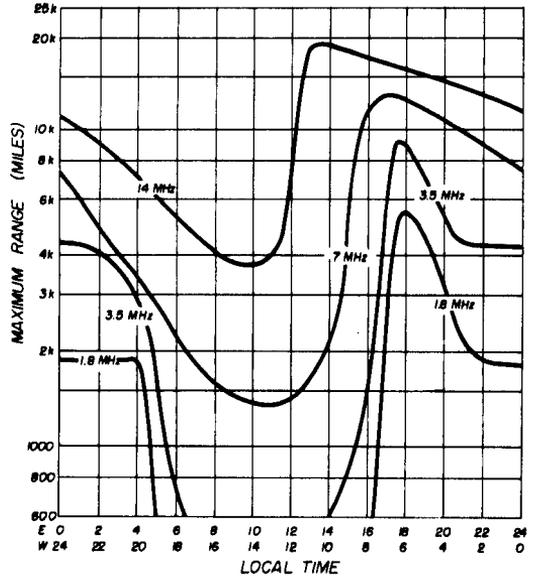


fig. 6. Maximum range vs local time for propagation to the East (top time scale) and the West (bottom time scale).

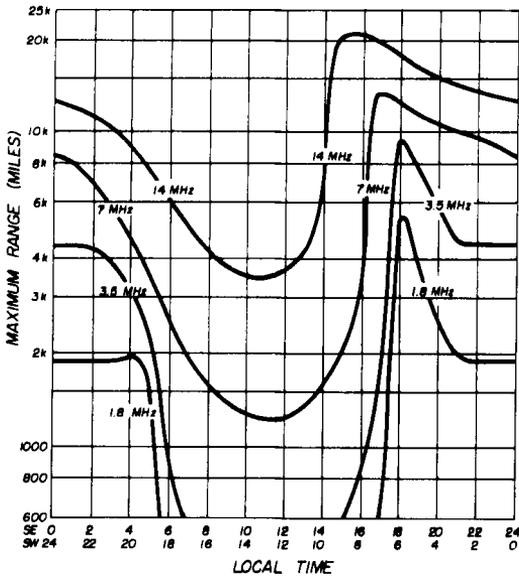


fig. 7. Maximum range vs local time for propagation to the Southeast (top time scale) and the Southwest (bottom time scale).

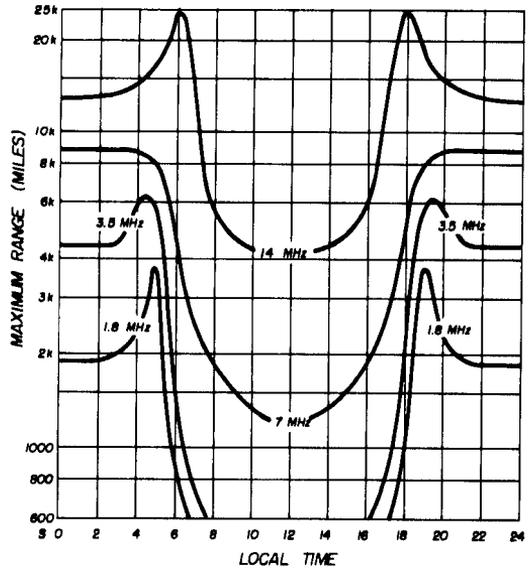


fig. 8. Maximum range vs local time for propagation to the South from 38° North.

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 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 that 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 0600 and 2030 hours control-point time. The band 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, 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 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.

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 latitudes (degrees N).

your latitude (degrees N)	direction				
	N	NE/NW	E/W	SE/SW	S
24	42	38	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	59	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.

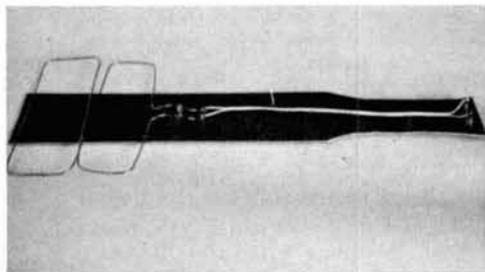
ham radio



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

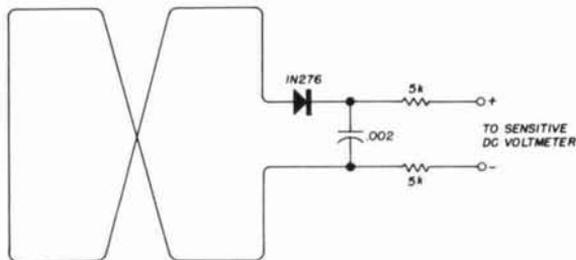
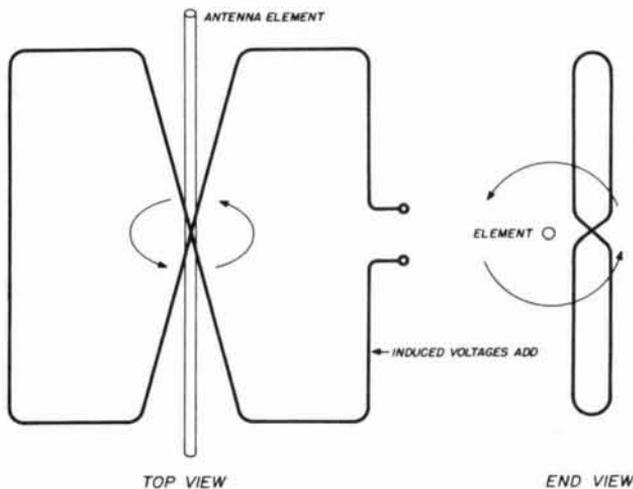


element. It is essentially a loop connected to an rf voltmeter which permits inductive coupling to the antenna element. The voltmeter gives a reading which is proportional to the current in the conductor.

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 masonite. No dimensions are critical except that the two loops should have approximately the same area.

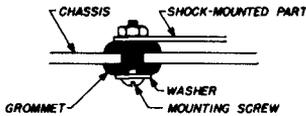


This device proved very useful in checking out the performance of the three-band ground plane described on page 32.

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

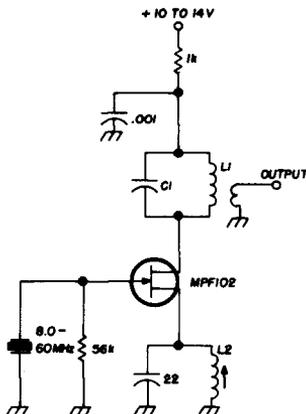


drilled for the grommet; then the relay, etc. is mounted. The grommet acts as a shock-absorbing washer. The larger the grommet, the greater the shock absorption. If the inside diameter becomes too large to hold the head of the mounting screw, a flat washer should be added.

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

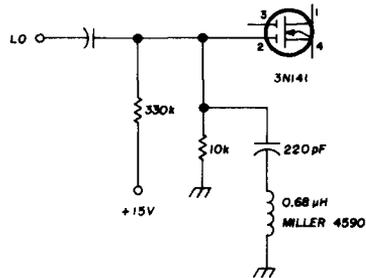


to the third, fifth or seventh overtone of the crystal. Although the schematic shows a link-coupled output, a small capacitor connected to the drain may be used to couple rf out.

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.

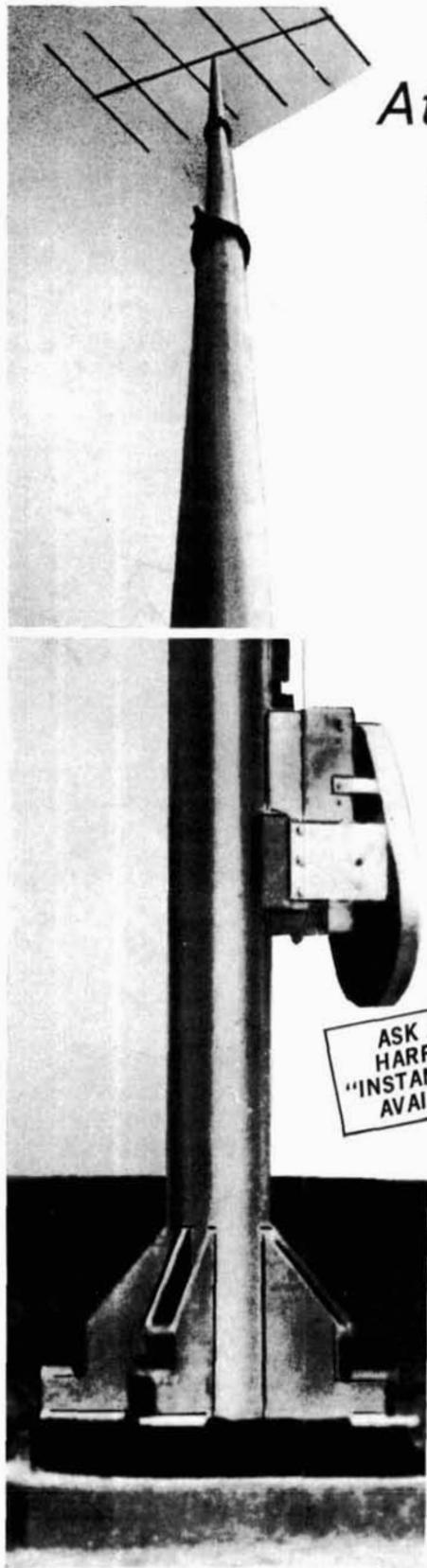
"NASTAR Receives Teleprinter," *CQ*, May, 1968, p. 46.

"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.

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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.

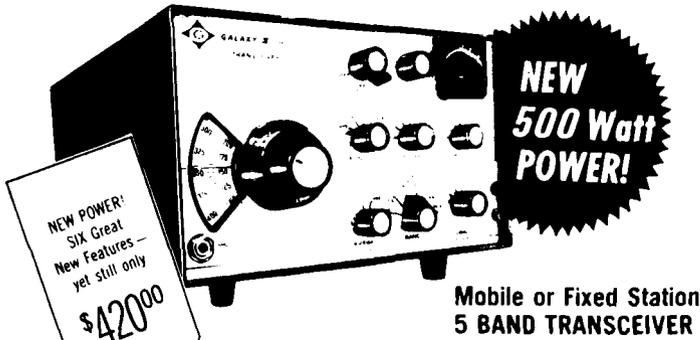
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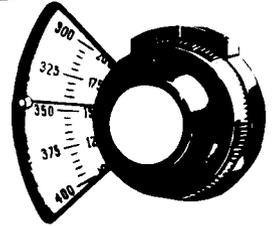


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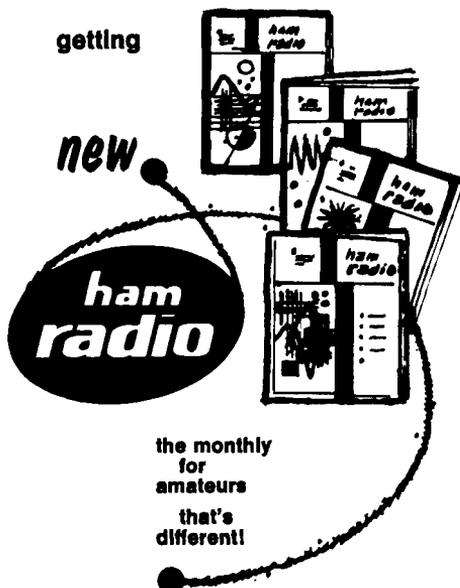
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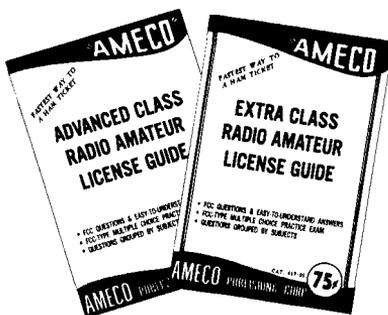
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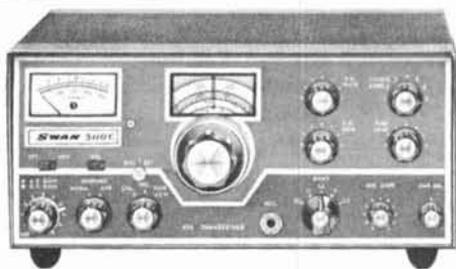
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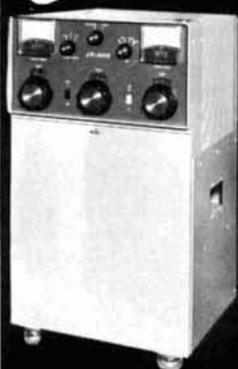
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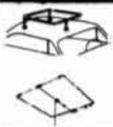
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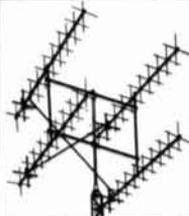
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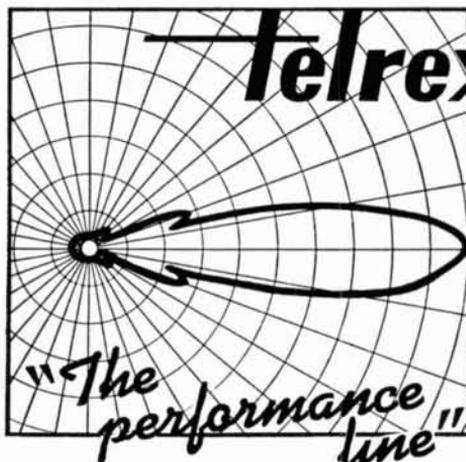
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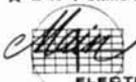
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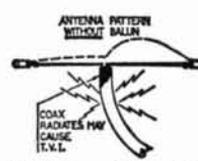
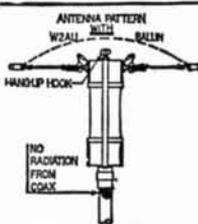
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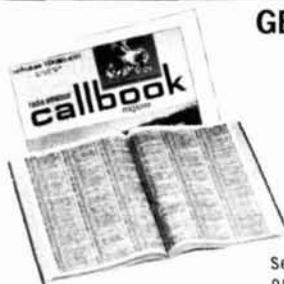
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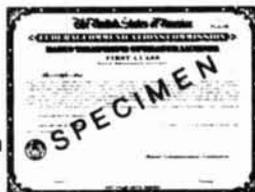
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