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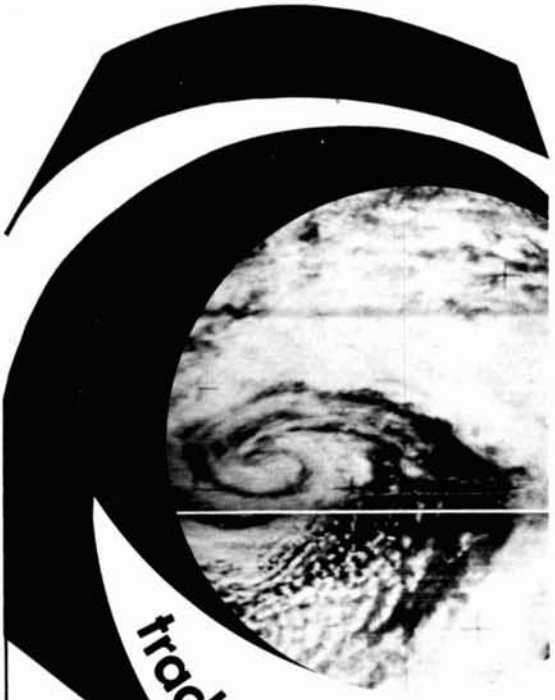

tracking and recording satellite transmissions

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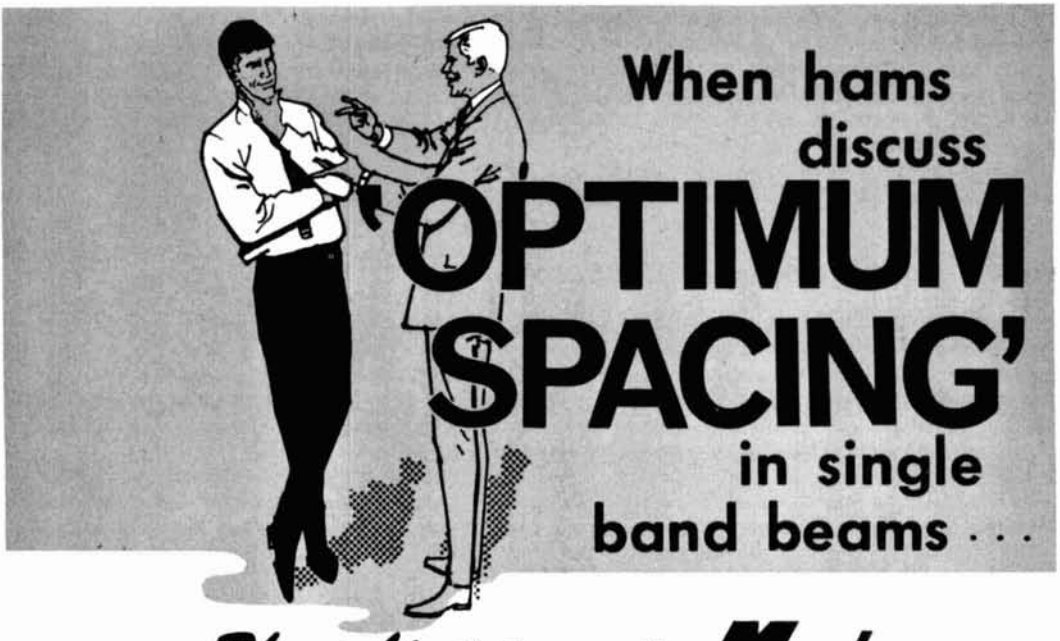

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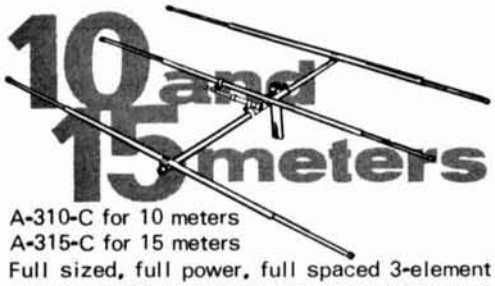

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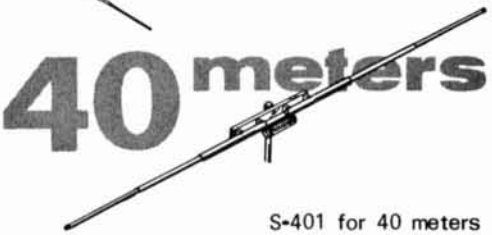

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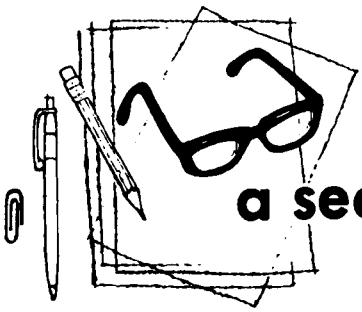

a second look

by jim fisk

If you asked the man on the street what area of electronics had the most impact during the past decade, he would probably mention color television or computers—they've certainly affected his life the most. To the amateur it might be the widespread use of single sideband or high power on uhf—depending on his interests and point of view. The engineer would no doubt mention lasers and computers, not necessarily in that order. All of these things are important of course, but in the future, the laser will probably have more impact on all our lives than any of the others.

Since the first working laser was put together by Dr. Maiman in 1960, it has captured the attention of scientists and the imagination of the public. To the general public, the laser is a zap gun that will cut through metal like it was a cube of butter. It is much more to the engineer: ultra-fast computers that use optics in place of electronics, three-dimensional TV pictures, and relief to our crowded spectrum by permitting millions of messages to be transmitted on a single light beam. This is just part of its potential; it has already been used in medicine, precision machining and welding, optical gyroscopes, optical recording that is 100 times faster than magnetic tape and optical memories for computers.

This doesn't mean that the laser has taken industry by storm—far from it. But advances are being made every day and it has been predicted that in the future, the lines between electronics, optics and quantum mechanics will be blurred through laser technology. Research is progressing slowly and it will be a good many years before your telephone calls will be transmitted over a laser beam, but advances are being made. Consider the number of materials that have been made to lase—over 2000; more than 1000 of these are semiconductors, with solids, liquids and gases making up the remainder. Solid crystals, carbon dioxide, neon, argon, helium-neon, organic and inorganic liquids and doped glass are just a few that have been made to work.

You're probably wondering what all this has to do with you. Just this: here's an area for the basement experimenter who has run out of worlds to conquer. So far as I know, the only amateur laser communications were conducted in 1963 by members of the Electrical-Optical Systems Amateur Radio Club. Although the output power of their laser was only 125 μW, modulated at 28.62 MHz with a Viking II, they managed to transmit over a 118-mile line-of-sight path in Southern California. Possibly other amateurs have been working with lasers during the ensuing five years, but I haven't heard about them.

In any event, lasers have come a long way since this early experiment in the San Gabriel mountains. For one thing, costs are down. If you're interested, you can buy a precision helium-neon laser for under $200 from University Laboratories in Berkeley, California. If you're interested in a semiconductor laser, try Allied Radio—they list a 15-watt unit (pulsed) in their 1968 industrial catalog for $95. If you want to build one yourself, consult the **Scientific American**—they had a complete construction article a couple of years ago.

If you do decide to try the laser, use caution and do a lot of reading first. They can be dangerous if you don't know what you're doing. Relatively small lasers have been used to bore holes in diamonds, so you can imagine what one would do to you if you get in the way of the beam. Protect your eyes particularly; just looking at a laser beam can damage the retina.

If you have a working laser system or are working on one, I'd like to hear about it; I'm sure some of our other readers would too.

Jim Fisk, W1DTY
Editor

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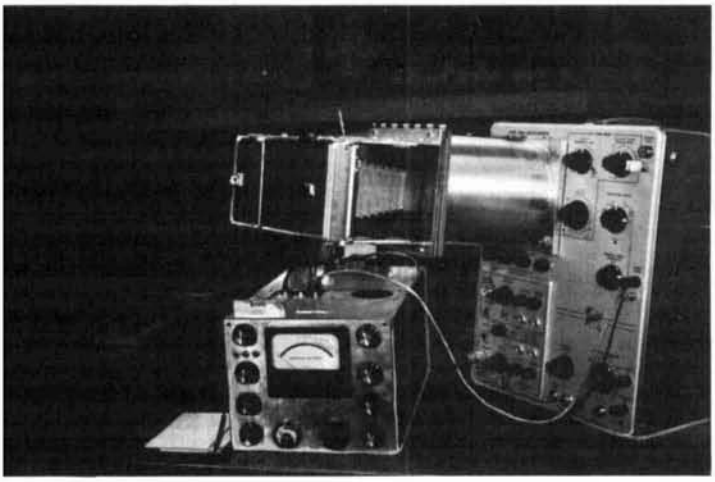


tracking and recording satellite transmissions

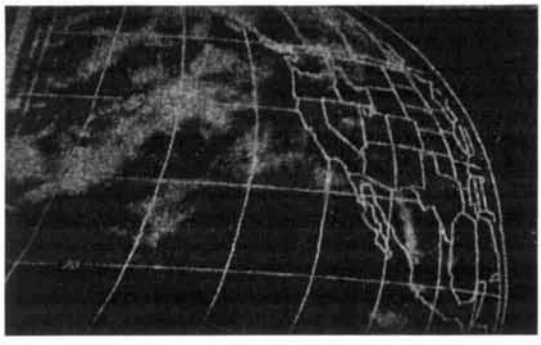
A project that will appeal to the experimenter, builder, vhfer and camera enthusiast

Greg Toben, W6CCN, 1336 Marilyn Drive, Mountain View, California

Automatic picture transmission (APT) is moving into its third year of continuous coverage. Prospects for the increased use of these versatile and extremely useful weather satellites look good.[1]

Two satellites, Essa IV and VI, are now transmitting overlapping APT pictures daily covering nearly the whole earth. An applications technology satellite (ATS-1) in a stationary orbit over the

Photograph of cloud cover over the Northern Hemisphere from the ATS satellite. This picture was gridded and re-broadcast from Mojave, California.



mid Pacific has broadcast pictures of the whole earth and rebroadcast pictures from other satellites, both in the APT mode (**table 1**).

Experiments will be continued by ATS-3, which is in stationary orbit over South America. A third in the Nimbus series will probably also transmit night-time infrared pictures early next year. A panorama of good pictures is available to users daily from Greenland, the North Pole and Siberia, to Florida, Central America and Hawaii. Coverage includes most U.S. latitudes. NASA encourages you to use this service, and participation in the many experiments is stimulating and rewarding.

This article describes an all-electronic system (**fig. 1**). It is based on a system used by WSM-TV, but good pictures can also be produced by other means.[2,3,4]

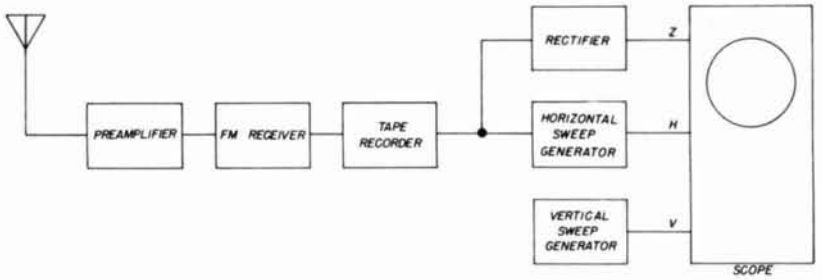
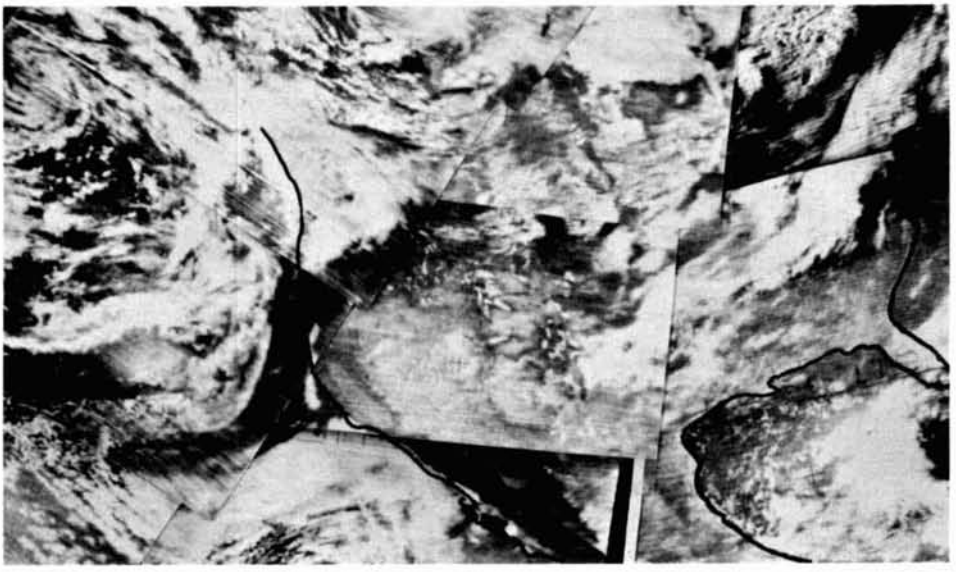


fig. 1. The complete receiving, display and recording system used by W6CCN for automatic picture transmissions. The z-axis rectifier is part of fig. 2; the horizontal sweep logic diagram is shown in fig. 3.

the antenna

Anything from an fm antenna to a stack of multi-element yagis can be used. A four-element yagi from the Handbook probably gives the best results for the least effort, but the picture steadily improves as antenna gain and signal-to-noise ratio go up. An S-meter is adequate

Panoramic view of North America as transmitted by ESSA VI June, 1968. It's difficult to make out the outline of the United States because of cloud cover, but the Gulf of Mexico, Baja, California, and the West Coast are in the clear. Note the many storm systems that are active.



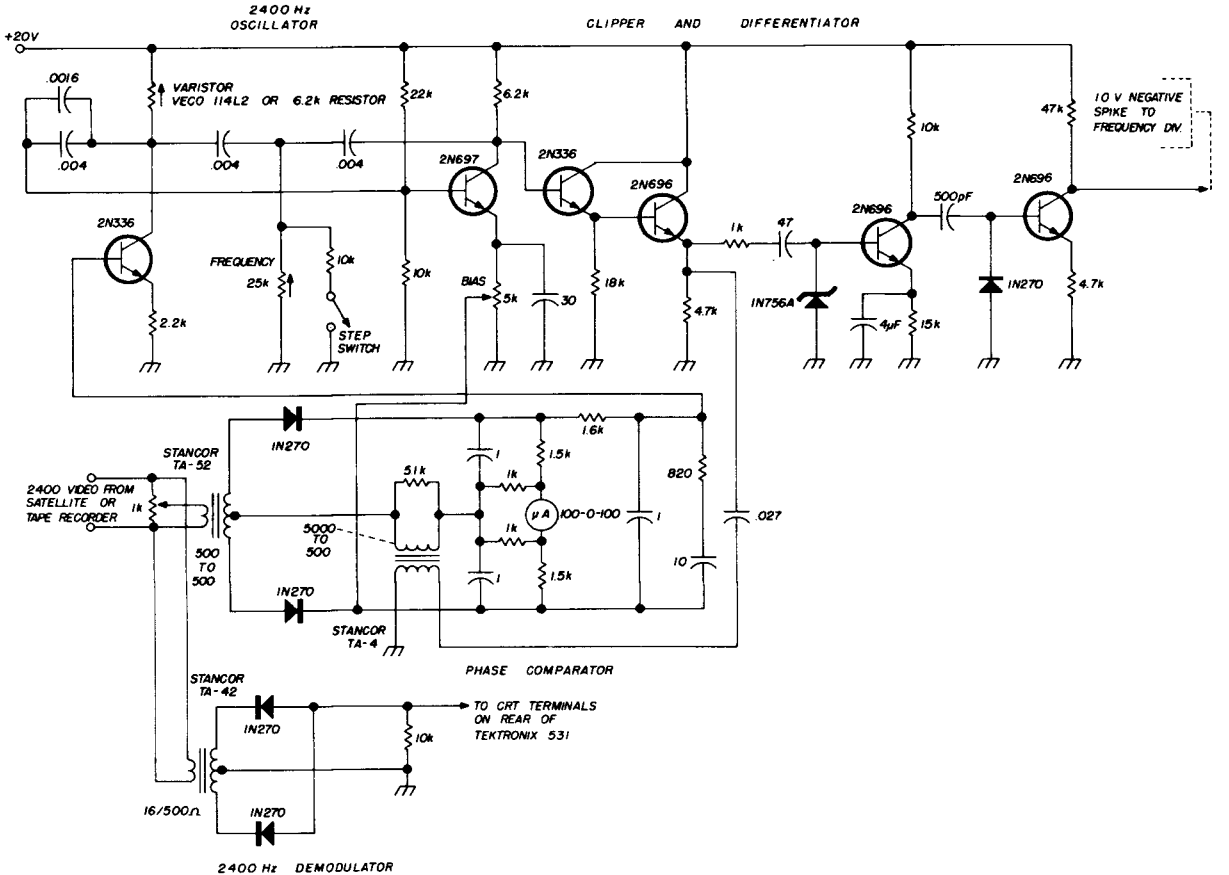


fig. 2. Phase-lock oscillator, clipper/differentiator and phase comparator used in the solid-state sweep generator used to display satellite weather pictures.

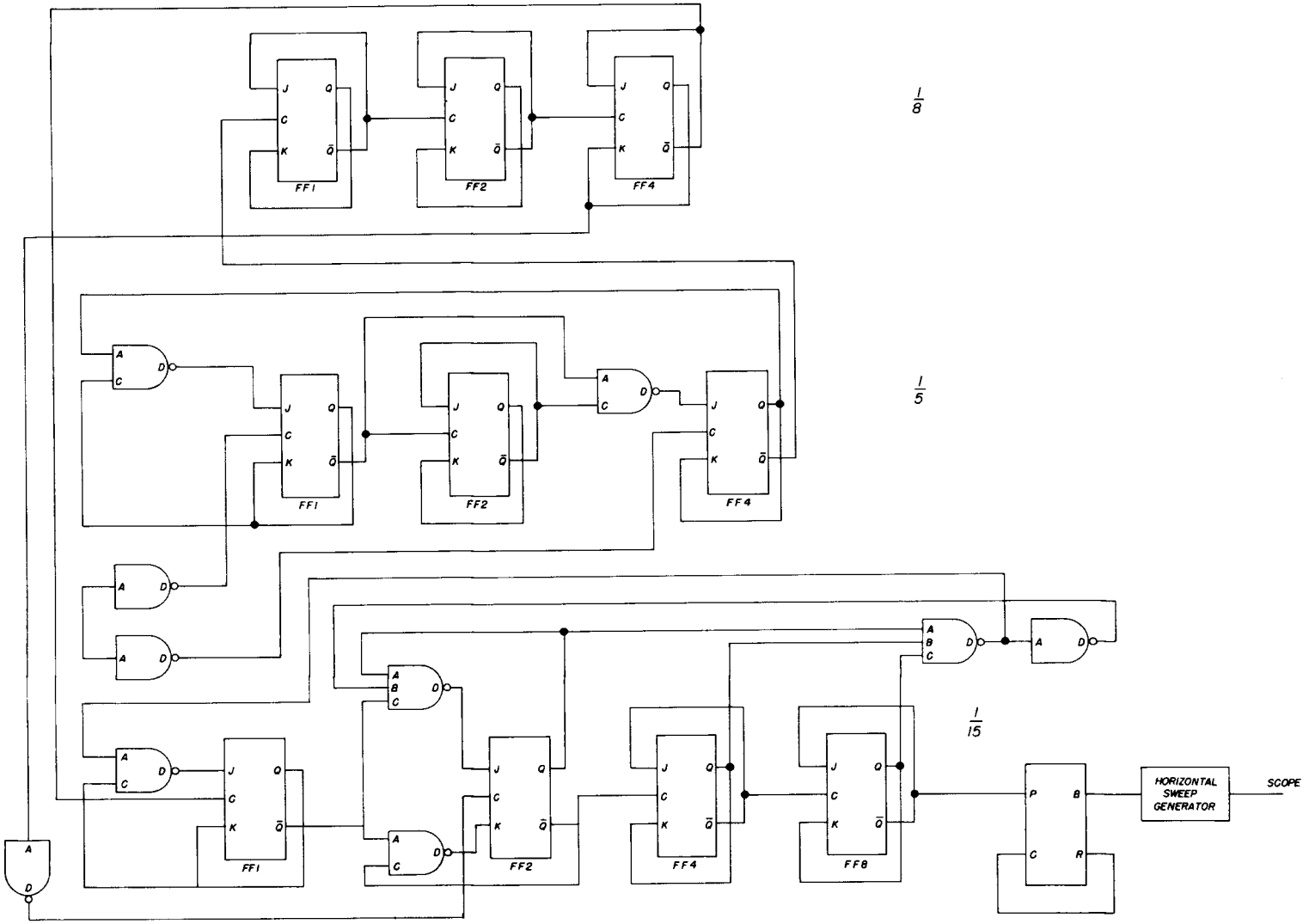


fig. 3. Logic diagram for the horizontal-sweep generator. Although the transistor circuits shown in fig. 4, 5, 6, 7 and 8 were used here, integrated circuits would be ideal for this job.

for monitoring, but a panoramic adapter is even better.

The ear is the most sensitive monitor and can follow the signal right down into the noise. Motor-driven antennas are necessary for remote control, but they are noisy, and need a wide speed range as well as some means for rapid search. Use a manual control if at all possible. So far, all transmissions have been within 1 MHz in the 136-138 MHz satellite band.

The antenna transmission line should have a flat response, and everything should be peaked for maximum gain. Rotation around the horizon (azimuth) is necessary on all but the overhead passes. Nimbus satellites come up in the south and disappear almost due north, while Essa satellites come up from the north and disappear to the south. Tilting up (elevation), is less critical, but rotation around the axis of the antenna (polarization) is important.

So far, the satellite signals have been linearly polarized, and reception on a circularly-polarized antenna has meant some loss of signal. There is a choice, then, of using a helix or crossed yagis and losing some signal, or rotating the antenna on its axis with the attendant mechanical problems.

The plane of polarization rotates as the satellite passes over. It may rotate several complete turns during a pass. The resulting fading is both sharp and deep (20dB). It is hard to adjust crossed yagis for truly circular polarization, and some fading always results.

the receiver

A receiver with 1-μV sensitivity can get a good picture on the overhead passes, but a good preamp is necessary to get clear pictures on the early and late passes. Overloading of the input stage is a problem in metropolitan areas where fm, TV, police and commercial fm stations bathe the antenna in a mish-mash of strong signals night and day. A coaxial cavity between the antenna and the preamp is some help.[5]

fig. 6. Single-shot multivibrator used in the APT sweep generator shown in the logic diagram in fig. 9.

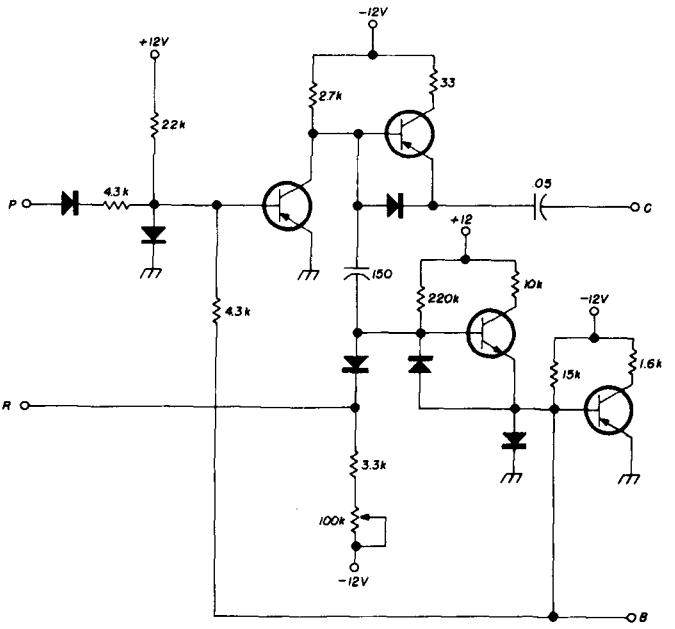


fig. 4. Three-way NAND gates used in the sweep system.

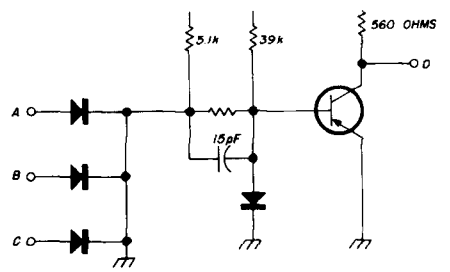
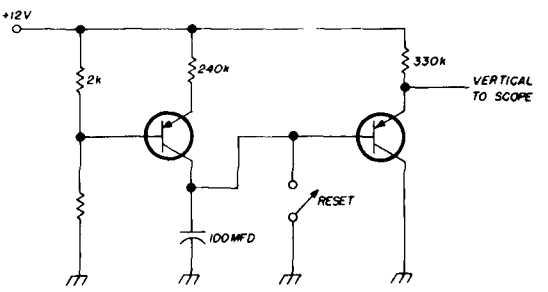


fig. 5. Vertical circuit.



Several receivers are possible, but the very excellent Motorola 148-174 MHz Sensicon "A" receivers are so good and reasonably priced that there isn't much point in discussing alternatives. These are receiver strips from commercial two-way radio equipment such as that used on trucks, taxicabs, etc. They are double conversion, single-frequency superhets and have five cavities in the front end, sharp filters in the i-f strip, good squelch and good over-all stability. The rf stages can be tuned to 137.5 MHz, and a type RM16 crystal (26.4-137.5 MHz) completes the conversion. The output is taken from the discriminator through a volume control to avoid overloading the input stage of the tape recorder. The Perma-kay i-f filter can be removed unless interference is a severe problem.

fig. 7. Horizontal sweep.

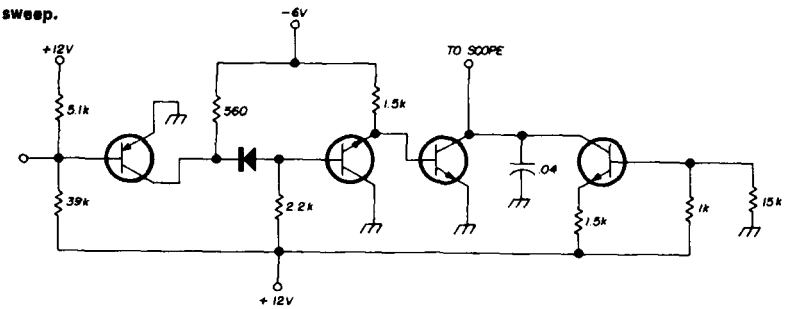


fig. 8. Flip-flop circuit used in the horizontal-sweep generator.

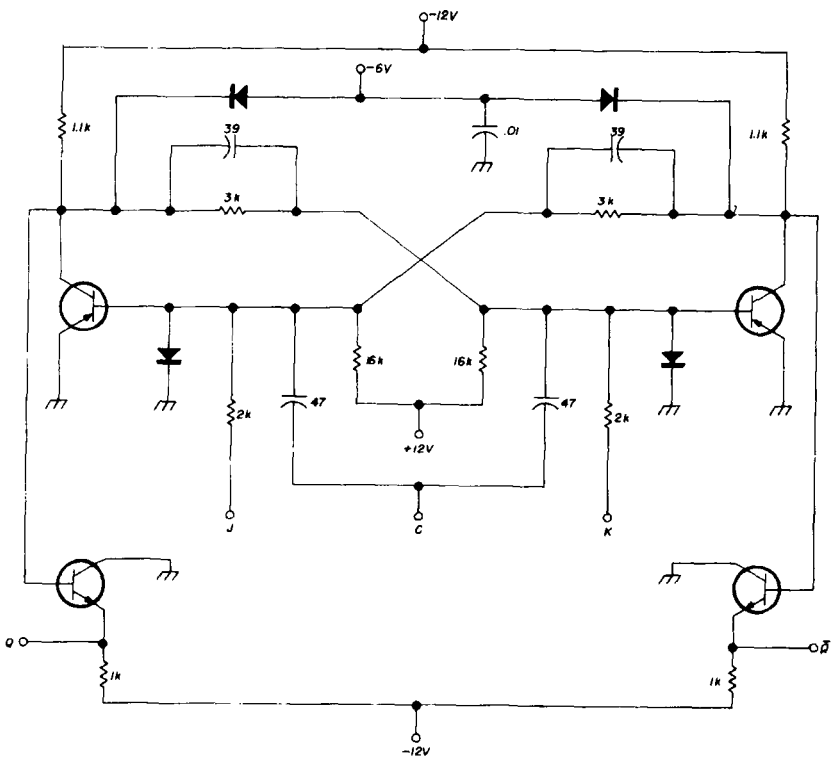


table 1. satellite automatic picture transmission systems

ATS	135.6 MHZ
Nimbus	136.950 MHZ
ESSA	137.500 MHZ
modulation	fm
deviation	10 kHz
subcarrier	2.4 kHz
modulation	a-m
maximum amplitude*	white level
minimum amplitude	black level
black-to-white ratio**	26 dB
video	dc to 1.6 kHz
lines per picture	800
line rate	four lines per second
frame time	200 seconds

Frames repeat every 208 seconds from Nimbus and about every 360 seconds from ESSA.

*** A white level calibration signal is transmitted between frames.**

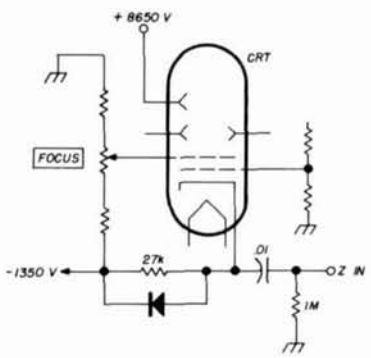
**** Equal voltage steps for equal gray-scale stops.**

tape recorder and sync system

The tape recorder can be almost anything with 3-kHz response and reasonable speed control. This eliminates only the small battery-powered jobs. A 7-inch reel of 4-track tape will hold pictures gathered in four two-day weekends (1 month). The pictures can be viewed directly, but this takes the operator's attention away from the antenna, and some pictures will be missed while changing film. The high cost of the film dictates that only choice shots be preserved, and the tape is used for previews and replays.

The output of the tape recorder will vary from the recorded speed if the line voltage or the mechanical friction has changed since the recording was made.

fig. 9. Dc-restore circuit for the Tektronix 531 oscilloscope; circuits for other oscilloscopes would be similar.



Storm pattern transmitted by the ATS satellite, gridded and re-broadcast from Mojave, California.

A phase locked voltage controlled oscillator (VCO) keeps the scope synchronized despite these changes. The circuitry for this "black box" is the only homemade electronic equipment necessary.

Again, there are many ways to do the job.[1,2] In my system, the shaped 2400-Hz pulses from the VCO, which is driven by the 2400-Hz subcarrier from the discriminator, are fed to a count-down chain. The first three triggers count up to 5, then reset and pulse on to the next three. These reset after 15 pulses and drive a single shot. The output of the single shot, through an inverter, shorts out the charge on the horizontal sweep capacitor in a few milliseconds; thus a new sweep is initiated after every 600 pulses. The time constant of the VCO is such that it will supply pulses and maintain sync for a few seconds if the signal fades out.

the scope

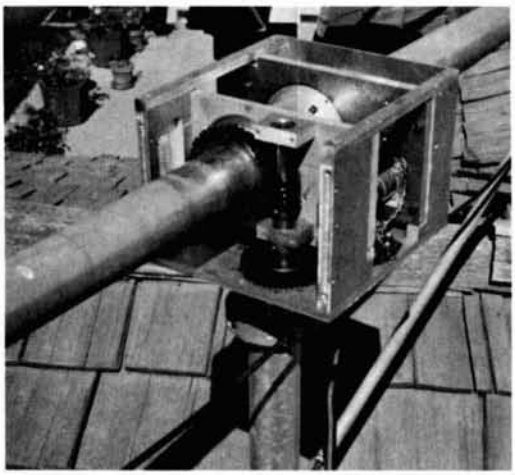
The spot size and focus pretty much determine ultimate picture quality. I use a Tecktronix 531, but I have seen good pictures on a 5-inch Paeco and on a TV tube. The intensity is set so the trace is just barely visible before the Z modulation is applied. At this low intensity, spot size is small and focus is better.

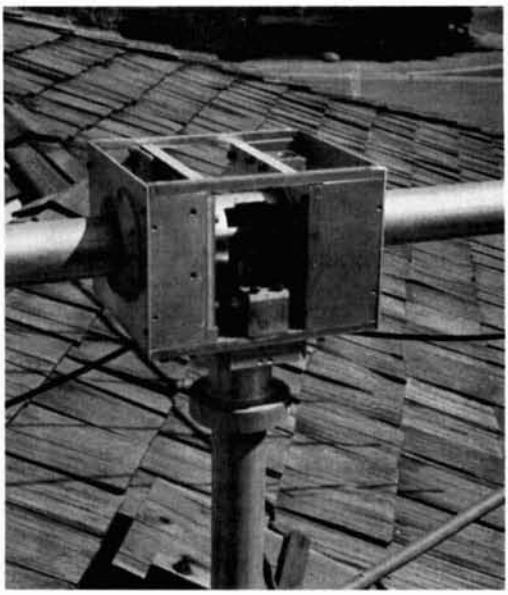
The vertical sweep is started manually and allowed to run off the bottom of the picture (or top).[4] Reversing switches on the deflection plates allow the picture to be painted from the top to bottom on n-s (Essa) passes and from bottom to top on s-n (Nimbus) passes. A diode restores picture reference dc to the CR-tube grid. P2 phosphor is about the right persistence, and the green gives a realistic tint to the pictures which will last for about three minutes in complete darkness.

the camera

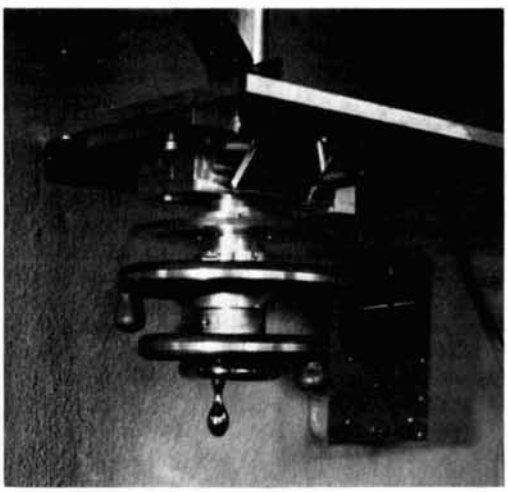
The camera was made by machining a piece of 6-inch aluminum pipe. The tube is lined with black felt and used as a hood when viewing in daylight. A used Polaroid No. 800 3 x 4-inch camera was used for its developing box and bellows. The lens, iris and remote-operated shutter are from an old Kodak. The sliding mount, as well as horizontal- and vertical-gain controls on the scope, give a complete freedom of picture sizes. Polar-

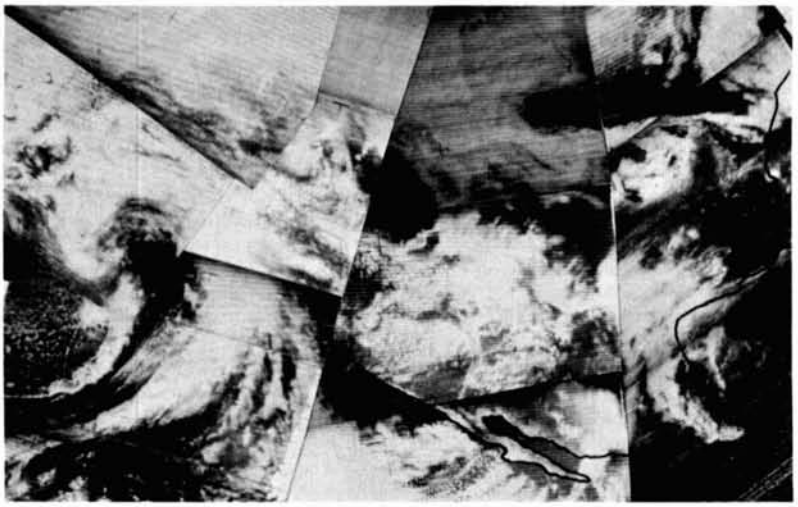
Some of the constructional details of the azimuth/elevation mechanism used at the base of the antenna. This mechanism is driven through the roof of the house by the controls shown in the photograph just below.





Antenna controls inside the house. From the bottom up is the antenna polarization control, elevation control, azimuth control, azimuth indicator dial, mirror readout and clamping mechanism. The selsyn elevation indicator is on the wall to the right.





Panorama of North America transmitted by the ESSA satellite in April, 1968. If you look closely, you can see some familiar landmarks in the lower-righthand corner—particularly the Florida peninsula and Baja, California.

oid type film 42 at f24 has been used for the past two years, but the newer rapid-processing films look promising for the future.

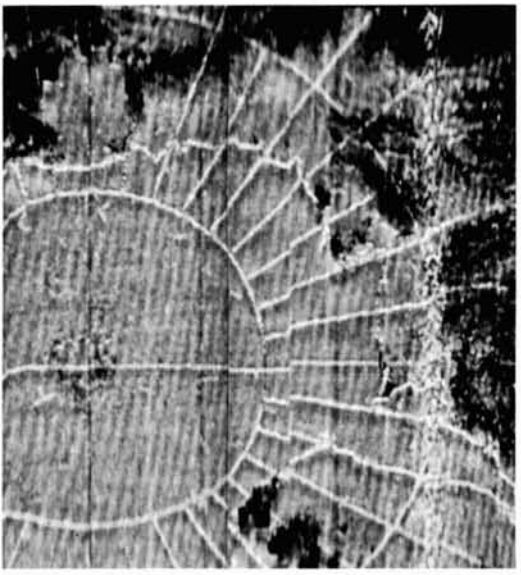
transit times

It takes about 1 hour and 50 minutes for these satellites to make a complete trip around the world. This works out so that Essa VI arrives about an hour earlier each day. If you hear it at 2 o'clock today, you could expect it at 1 o'clock tomorrow, or at 8 o'clock six days from now. On a given day the other passes can be expected every two hours less ten minutes, just about time enough to run to the store or cut the lawn.

In an APT household everything runs on satellite time. Three passes are available every day, sometimes four, with the middle pass near noon. The longest pass lasts fifteen minutes (three pictures), but the length of each pass depends on where it cuts your receiving range circle. The range depends on the height of the transmitting antenna (600 miles) and receiving antenna, and also on hills, buildings, etc., on the horizon.

W1AW broadcasts accurate equatorial crossing times, but it works out fairly well to just turn up the volume and squelch and go on about your business until the satellite comes over. It won't be long. The Essa signal is a distinctive 2400 Hz with

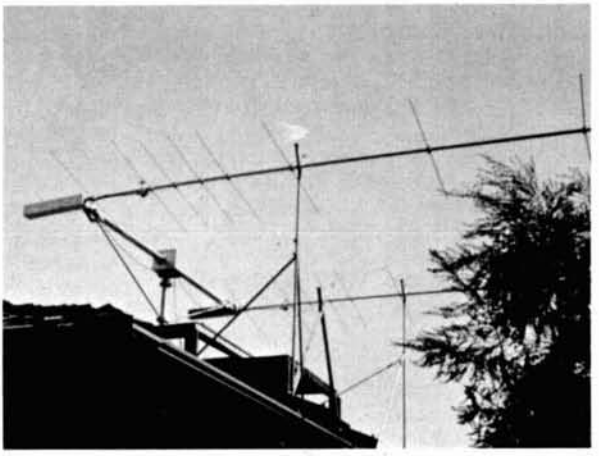
South polar region.



4-Hz modulation for 208 seconds and a steady unmodulated tone for two minutes, and repeat. Nimbus continuously transmits 4-Hz modulated signals with no space between pictures.

The ATS satellites carry a variety of other experiments; however, their power is limited and their time is scheduled to avoid interference with Essa and Nimbus picture-taking operations. They are therefore on at times that are inconvenient

More details of the antenna system used to track ATS satellites. Above is the mast which goes through the roof to the antenna. Above right is a closeup of the antenna mount; two bevel gears are used for polarization control; counterweight at bottom. To the right is the complete antenna system.



for amateurs. A schedule can be had from NASA when your station is ready.

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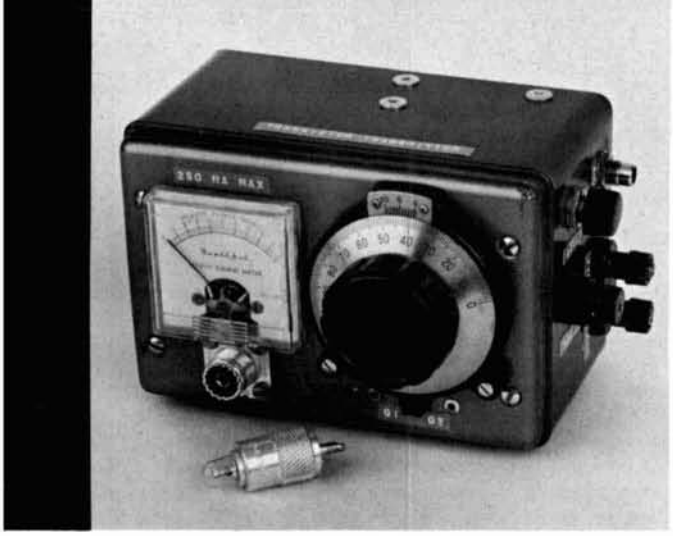
ham radio



broadcast engineer's transistor transmitter

Although the upper frequency limit of this solid-state transmitter is 1600 kHz many of the ideas are directly adaptable to our lower amateur bands

Ralph W. Campbell, W4KAE, 316 Mariemont Drive, Lexington, Kentucky 40505

Notwithstanding the fact that high-power tubes are here to stay, transistors are available at last for a few dollars with a respectable amount of output power. My broadcast engineering friends tell me vacuum tubes are obsolete; this may be true—if you're designing receiving circuits. On the other hand, they may be right on some power applications too.

They wanted me to build a solid-state power oscillator capable of two watts output over a design range of 550 to 1650 kHz. The completed unit had to be absolutely stable with provision for external modulation. The unit was needed to drive a General Radio 916A Radio Frequency Bridge used for testing broadcast antennas after sign off. An audio oscillator is used for modulation.

Several watts of rf power must be developed to override West Coast broadcast signals which are picked up and confuse measurements. The usual method of measuring common-point antenna resistance and reactance is inside the transmitter building where all the phases and loops and feeds converge. However, without a power oscillator, it's impossible to correct mismatches at each tower base which send false indications to the common point.

Power is a problem for vacuum-tube equipment because the ac mains to a tower house

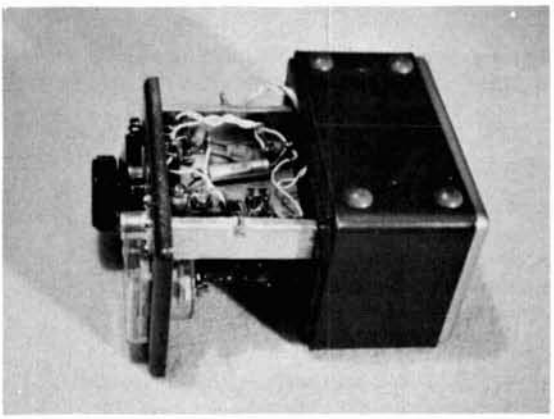
may fail just when you're all set up. With transistors, a car battery is all you need. At 2 o'clock in the morning, reliable battery power is your best friend. The hunt for a small high-power TO-5 transistor was on.

circuit

This power oscillator uses two 2N2631 NPN overlay transistors. The first stage is a series-tuned Colpitts oscillator operated in class A, and the second transistor is operated in class B. The class-A operating point and loadline is shown in **fig. 2.** An emitter-follower output stage, which is normally biased "off," is well into the class-B region when an output of approximately 1.7 or 2 watts is obtained, since there is only current gain with this type of circuitry.

I chose a common-emitter oscillator with an emitter-follower output stage to match the low (and varying) impedances encountered in broadcast engineering work. Oscillator ouput impedance was optimized, mostly by experimentation; frequency compensation over the broad range of the broadcast band was provided (almost accidentally) by using the base bypass value of 0.005 μF with a 200-ohm resistor. This was chosen to equal the measured Colpitt's output-coupled impedance. An "anti-squegg" resistor was added to prevent reflexing in the common-emitter oscillator on its low-frequency range.

A slide switch selects either of two expanded ranges to cover the broadcast band. Two high-Q Carbonyl SF toroidal cores keep frequency stability within tolerable limits, although a trade-off was made in circuit Q because of my inability to find high value silver mica capacitors for the low collector and base swamping reactances. An unloaded Q of 230 is easily achieved in practice.



As a result of W1DTY's article[1] on transistor oscillator design, I undertook the challenge to design this piece of equipment! No one else has devoted much to this topic. Other articles might have been of help, but the average collector characteristics on the 2N2631 data sheet showed the operating area to avoid second-breakdown. A glance did reveal the ac beta to be equal to 8. I don't know what the dc beta is, but I've found a stability factor of twice this (16) to be handy when designing the bias circuit.

A TO-5 type transistor was preferred after I discovered Wakefield 254SI insulated heat sinks were available for $.25.* If you look inside a 2N2631, you'll see why the chip is so easily heat sunk. All of this leads us back to the large amount of power dissipation required for a class-A stage delivering 1.7 watts. Strictly speaking, 1.75 watts was the goal with good frequency stability and isolation from loading effects. Several transistors, one hopeless 2 AM session and a good car battery—later—I succeeded.

development

The front of the completed transistor transmitter is shown in the photographs. Looking at the surplus meter you can see that an arbitrary maximum limit of 250 mA has been placed on this movement. The meter shunt with this instrument is approximately 1/3 ohm. However, it's best to cut and try even with figured data on hand, because even the best ohmmeters are crude instruments with which to measure shunts. Use a series-dropping resistor and a 1½-volt dry cell along with an instrument of known accuracy in series and adjust the full-scale value for correct current.

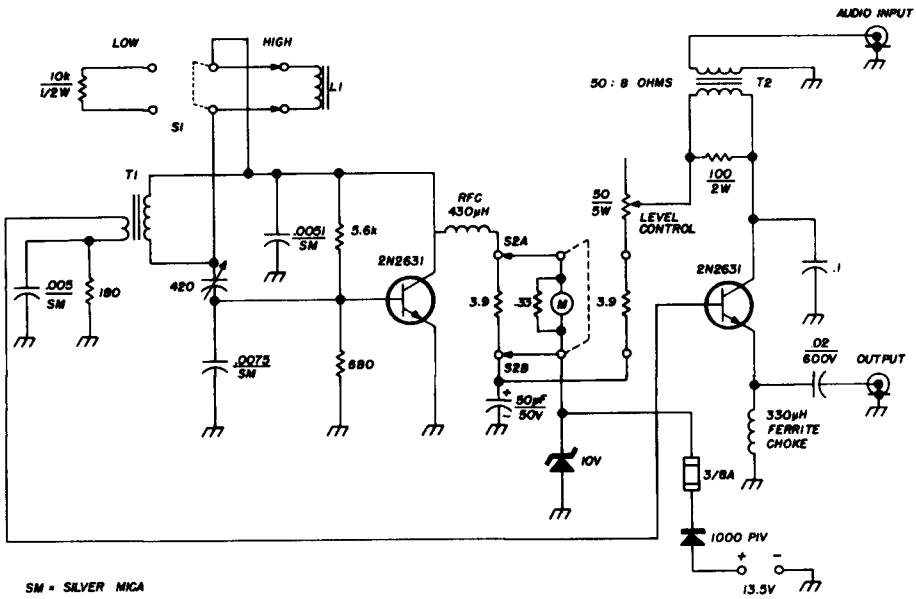
In fact, if the 3/8-amp fuse holds, use a fast AGX 3/8-ampere as opposed to the AGC or 3AG type. This is important since the AGC fuse may not be fast enough to save those

* Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order catalog number 60E6541. $.25 plus postage; shipping weight, 2 ounces.

overlays! The AGX fuse is only one inch in length but will fit the usual 1¼-inch fuseholders.

The 2N2631 overlay transistors have a low collector-to-emitter breakdown voltage. Although the usual rule is to take ¼ BV_{CBO} as a safe value (assuming 100% modulation), in this case it's safer to use 13.5 volts instead of 20 volts as would normally be chosen. First of all, a safety factor of 75% is necessary to prevent excessive rf swing at some points on the dial (such as when squegging); secondly, voltage at 75% of ¼ BV_{CBO} works out to be 15 V. I went above this voltage, and my school-of-knocks observations showed frequent failure of the class-B output emitter follower.

Luckily, with modulation I had no problem. But after seeing how hard equipment can be used by professionals in the consulting field, I slapped a +13.5 V limit on the unit. Also, the discussion on modulation should be added to. You can see in the illustrations that I used a small 3-watt modulation



L1 Close wind number-22 wire on Permacor 57-1516 ferrite core until 85% of the circumference is full.

T1 Primary: close wind number-24 enamel wire to cover 20% of the circumference of a Permacor 57-1516 ferrite core; secondary is close-wound number 26 to fill up remainder of circumference.

T2 50:8-ohm transformer (Knight 54B1493).

fig. 1. Schematic diagram of the broadcast engineer's transmitter. Power output is 2 watts. Both transistors are mounted in Wakefield 254SI heat sinks.

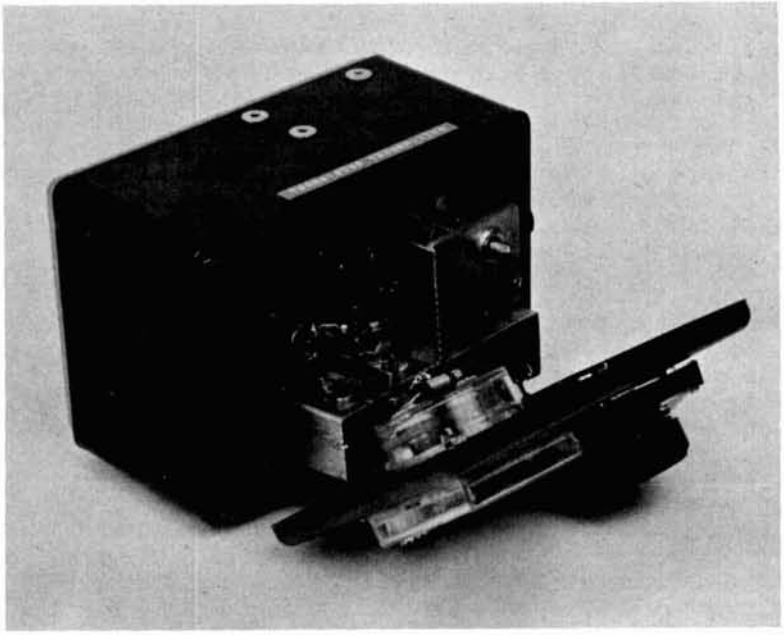
you must allow for peak maximum oscillator base reverse voltage. By this I mean that for the power oscillator stage, a maximum reverse voltage swing of −1.5 volts is all that can be withstood from base to emitter.

With these conditions, even a stable operating point of 10 volts V_{CE} and 55 mA is not held perfectly, because this excursion must exist in any power oscillator. The limiting transformer with a 100-ohm 2-watt composition resistor to swamp hookup transients. This is because transients generated in connecting the unit by moving leads to and from the terminal strip and the battery can cause trouble.

To stabilize the device from such troubles, I installed a 50-µF electrolytic. I got oscillator decoupling from the modulating voltage as

well, since these circuits are in tandem if not truly parallel to the supply battery. Behind the upper terminal is a "goof-proof" 1000-PIV diode. This prevents using the wrong polarity from the battery terminal and vaporizing the transistors.

The internal view shows one of the main features: the Permacor 57-1516 Carbonyl SF toroidal core wound with adhesive tape and dipped in INSL-X™ high-voltage coating. The coating and taping were necessary when too many long-nosed pliers pierced a thinner



undercoat and a solder-splash adhered to the wire insulation and caused a short.

Also, the engineers had an infield short using the old frequency dial which put collector voltage on the base of the oscillator transistor. Problem: rotor to stator plates touching. This all cleared up with the new Midland dial; rigidity is important.

It is important to mention that the stability provided in this toroidal inductor is higher than would be obtained using a ferrite-cored solenoid. The reason is that copper losses, stray coupling and the Q of the variable 420-pF air capacitor used with it offset capacity considerations when figuring from the charts.[1] The reason this is true is that an extra band or range must be provided, even with 420 pF, to cover the entire BC band. The shunting reactance provided by the second toroid permits coverage from 940 to 1550 kHz.

Complete coverage of the BC band was not obtained at first. It was necessary to remove a blocking capacitor originally installed between the variable rotor connection and the underchassis return to the base of the bipolar oscillator. This spread the lower limit to about 535 kHz and moved the upper limit from 960 kHz down to 940 kHz on this lower band. I found no trimming adjustments were needed, but it was necessary to unwind several turns from the shunting core-wound inductance to achieve the upper limit of 1650 kHz on the upper range. Over-all frequency response was finalized from 535 kHz to 940 kHz (lower position, with anti-squegging resistor in-circuit) and 950 kHz to 1650 kHz (upper position).

The subchassis is an aluminum CB-1625 on top of another CB-1625, cut down to fit the Bud CB-463 utility cabinet. Use of two subchassis was for good heat-sinking purposes

mainly, and both Wakefield 254SI insulated sink studs are firmly attached.

The Philmore 1945T variable capacitor shaft is cut very short so that it will fit properly with the Midland dial. Also, space must be allowed for slide switch clearance by cutting a half-circle with an inch-and-1/8th Greenlee punch on the lower lip of the utility cabinet. Spade bolts hold the chassis as shown, and 1/8th-inch steel pop-rivets hold the spade bolts. Speaking of rivets, those you see on top are holding in the modulation/intercom transformer. Although pop-rivets are time savers, screws and bolts are really best.

I looked through catalog after catalog for a 50- or 100-ohm low-impedance and low-dc resistance power transistor transformer without luck, except for the very large Stancor TA-11. It was so large it wouldn't fit the cabinet without drilling into the laminations, removing the crimp-clamp housing and twisting heavy-gauge stainless-steel wire around what was left! This was intolerable and could've fallen apart in use if I hadn't found the identical electrical equivalent from Allied.*

There it was, 3 watts, same ac resistance as the TA-11, and I could mount it. I recommend getting two if you can afford it, because you may want to build a transistor a-m rig sometime after getting your feet wet using 2N2631's. In all fairness to the TA-11, I must admit the extra size may be necessary for low-frequency hi-fi performance in other applications.

The high-range toroid is, as I said above, wound with heavier wire switched in shunt with the low-range one. The output coupling to the emitter-follower is the same as for the first case. Some possibility for squegging occurred, so a 10k half-watt resistor was used to swamp the low-range coil.

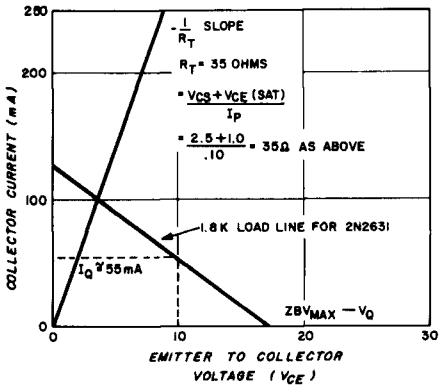
Silicone rubber is the chassis mount for the higher frequency toroid. Both of the cores use this as an encapsulant as well as a mount. With the first coil, two 1½-inch fiberglass circles were cut from 1/16-inch sheet stock. A 9/64th-inch hole was drilled in the center of the inner disc, and 6-32 binder-head machine screws and nuts were tightened down on the pretinned Beldenamel wire from the inside.

* Catalog number 54B1493. $2.03 plus postage; shipping weight 8 ounces.

design

Probably most difficult and most important is finding the correct operating point. Since there is little information in simple form available to the amateur designer, you must rely on load lines and select an operating point so the game fits the rules (to explain operation, if not predict it). **Fig. 2** shows the load line I used. With the manufacturer's data

fig. 2. Load line used in designing the transmitter.



sheet in hand, I took the V_Q, or voltage coordinate point, from the maximum supply voltage of nominally 13.5 V, allowed for a 3.5-volt drop in the combined circuit and saturation resistances (R_t), and established the quiescent current at the measured value of approximately 55 mA.

Fig. 2 shows the collector voltage swing from V_{CS}, the collector voltage at saturation (not neglecting circuit losses) to a peak value of twice the supply voltage (V_{CCmax}) or 27 volts minus the V_Q value, ten volts. Looking at the graph, the 17-volt swing should be the first thing you see. Any convenient point on the load line yields a value of about 1800 ohms for the load resistance.

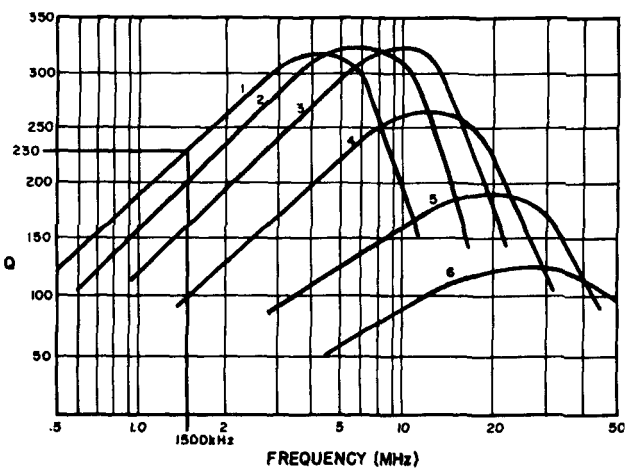
Load lines can be drawn in any part of the operating area, but I suspect that rf base current is so difficult to measure that you should use the cut-and-try approach. Since the overlay transistor is priced below $5.00, this may be the most feasible anyway.

The schematic diagram of the transistor transmitter is shown in **fig. 1.** The biasing resistance was chosen to equal approximately three times the load line value or 5.6k. Cut-and-try methods resulted in about 820 ohms form R_b, and R_a equal to 3.9k. A more reasonable criteria might result in higher dc stability by finding the stability factor. However, I felt that since this was a power oscillator, selection of minimal biasing resistance for R_b was more important.

At three times the load value, only 25% of the oscillator power is consumed in the biasing resistance. Another consideration is the Q with the Permacor 57-1516 core. The specifications for the Permacor core are shown in **fig. 3.** The inductor I used is 14.2 μH; from **fig. 3,** the unloaded Q is 230 at 1500 kHz.

fig. 3. Specifications for the ferrite Permacor cores used in the transmitter.

1. 48 TURNS .020 WIRE = 14.2 μH
2. 33 TURNS .020 WIRE = 6.7 μH
3. 24 TURNS .020 WIRE = 3.9 μH
4. 16 TURNS .020 WIRE = 1.6 μH
5. 9 TURNS .020 WIRE = 0.5 μH
6. 4 TURNS .020 WIRE = 0.18 μH



The output 2N2631 is operated in class B. Input pulses turn the emitter follower on and off with a residual bias voltage stored in C_3, which is also a bypass; the time constant is chosen to compensate over the lower range of oscillator frequencies from 535 to 940 kHz. It turned out that the same values worked well on the higher range from 950 to 1650 kHz.

Before going on, a word about the biasing of the oscillator transistor is necessary. At first R_a and R_b were chosen experimentally to turn-on the transistor. The values were 3.9k and 820 ohms. This wasn't very scientific. So, using these values to explain the results, we computed the stability factor. Results: stability factor of almost exactly 16! Twice the ac beta. The network provided adequate dc stability (something pretty hard to find in germanium bipolars) and no overheating.

conclusion

In this article I have shown how an overlay transistor can be used as a power oscillator for consulting broadcast engineering. I am directing this effort to the ambitious engineer who is also a competent amateur and hope that we'll be getting people interested in transistor uses by showing how a commercial piece of equipment can be made to work reliably and furnish power. Not much has been said in periodicals or outlines, so I assume the empirical method to be as effective as any. The biggest hitch, I think, in rf calculations is the difficulty of measuring rf base currents where needed and deducing class of operation.

Class of operation is sometimes difficult to determine even with tubes, since curves don't hold in every case. With a beam pentode I am currently using, different screen voltages and drive can push an AB_1 linear into AB_2 or B. In my own uses with tubes I have found that by simply listening to the linear output as detected audio, you can learn quite a bit. And, with constant current curves (or presumably constant collector curves), output power and tank-circuit efficiency can be calculated.

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ham radio

single-sideband detectors

How the single-sideband signal is demodulated in your receiver

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

The air is full of single-sideband signals these days. Up and down the ham phone bands, a-m holdouts can hear the donald-duck chatter of their modern-minded cohorts QSOing away—squeezing every last decibel of usefulness from every watt—on sideband.

It doesn't take much to turn that chatter into plain talk. Just a special detector will do it. At least those guys with a-m sets could **listen.**

The sideband operator already has that special demodulator, built right into his ssb receiver. It goes by many names, but the one used most is **single-sideband detector.** Other names come from the method of operation. Product detector, heterodyne detector, carrier-insertion detector, bfo detector—are among the terms that describe typical ssb demodulators.

The basics of a sideband detector are simple. The signal your sideband receiver picks up is nothing but one sideband of some operating frequency. To recover the voice modulation which created that sideband signal, you need a carrier for the sideband to heterodyne with. (That's how

an a-m detector works; the sidebands heterodyne with their carrier in a nonlinear detector—usually a diode.)

A single-sideband signal has no carrier of its own; that was removed at the transmitter. So a carrier has to be added at the receiver. Then the carrier and sideband can be fed together through an ordinary diode detector, and the voice signal recovered.

The i-f amplifier is the best place to mix the carrier and sideband signals. The frequency there is always the same, no matter what band is tuned in up front. A single-sideband detector mixes the i-f sideband signal with a signal at the frequency the i-f carrier would be if there was one. The steady signal is then called the carrier, since its purpose is to supply a signal against which the sideband can beat for demodulation—the purpose of a carrier.

The steady signal in receivers is most often supplied by the bfo that is used for code reception. Adjusting the **bfo pitch** control lets you control the timbre of the demodulated voice. In transceivers, the signal more often comes from the carrier oscillator; it's common practice to generate the initial carrier (before balanced modulation, sideband filtering, and up-translation) at the same frequency as the i-f. Whatever its source, the fixed frequency is fed to the demodulator system along with the sideband.

You can see one oversimple system in **fig. 1A.** (Don't bother copying it though; it's inefficient.) For distortion-free detection in any mixing-type ssb demodulator, the carrier signal must be much stronger than the sideband signal. One way is to attenuate the sideband signal; that's why the variable attenuator is included.

The arrangement in **fig. 1B** is a little more effective. The improvement comes from isolation provided by an amplifier between the carrier source and the mixer. The amplifier also gives that needed boost to the carrier signal.

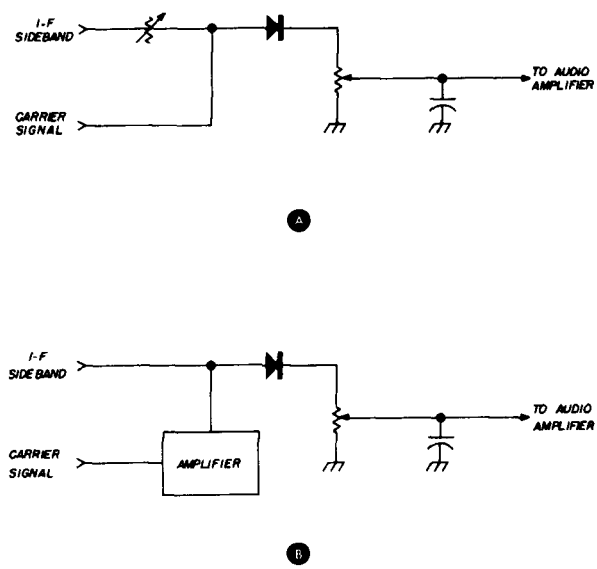


fig. 1. The simplest principles of ssb demodulation. There is no isolation between signals in A; isolation plus carrier amplification are provided in B.

toward a better way

Extra care must be taken with single-sideband detectors. Distortion is always a possibility, unless each signal is handled so that the only nonlinearity is in the detection circuit itself. Applying the signals to a single detector diode is not the most desirable way to get this particular job done.

Better efficiency can be had from the improved version in **fig. 2,** using two diodes. The carrier signal is applied to them in a

parallel mode (its coupling capacitor is connected between their cathodes). The i-f sideband signal, on the other hand, is in series with both diodes. This parallel-series hookup lets the two signals mix in the special way that produces an audio signal.

The special way mixing takes place in **fig. 2** as the result of how the signals are ing the output—should sound familiar if you've read earlier articles in this series. Beginning on page 24 of the May issue, I described balanced modulators in ssb equipment. They also use this two-mode way of handling input and output signals.)

Tubes offer a better means of isolating and mixing (see **fig. 3**). Furthermore, the tubes can build up the carrier-signal

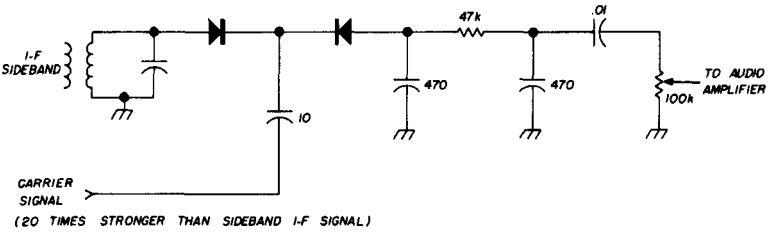


fig. 2. More elaborate diode circuit for ssb demodulation; circuit has fundamental resemblance to balanced demodulators.

brought together. The carrier signal is fed to the stage in a mode different from that of the sideband. The mixing generates a **product** of the two signals instead of sums and differences. (That's where the name **product detector** comes from.)

Furthermore, the output signal is taken from the stage in series—a mode opposite to the carrier input mode. This encourages cancellation of the input carrier, keeping it from the output. The **product** of this mode of mixing, therefore, is a relatively pure audio signal—the recovered voice signals that originally formed the sidebands. Any slight remaining carrier or sideband signal is eliminated by the 470-pF capacitors and 47k resistor.

(The parallel/series method of feeding the two signals into the stage—and of taking the output—should sound familiar if you've read earlier articles in this series. Beginning on page 24 of the May issue, I described balanced modulators in ssb equipment. They also use this two-mode way of handling input and output signals.)

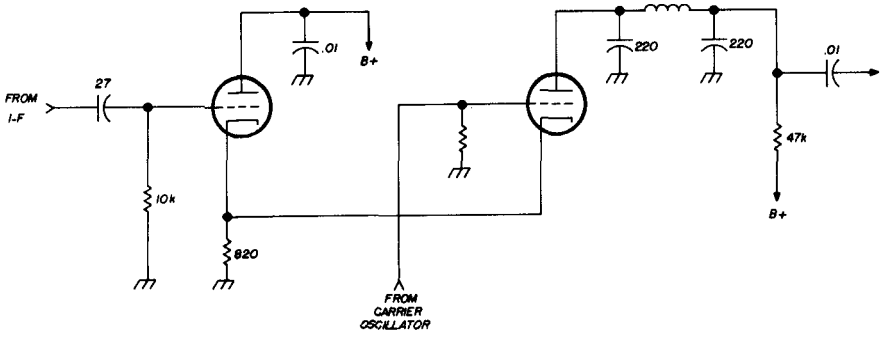
Tubes offer a better means of isolating and mixing (see **fig. 3**). Furthermore, the tubes can build up the carrier-signal strength. The sideband signal from the i-f amplifier is applied to a cathode follower; that isolates the signal source from the mixing circuit, without adding any gain. The sideband is then cathode-coupled to the mixing tube. Meanwhile, the carrier signal is also fed to the grid of the mixer, and is amplified.

These signals mix within the tube. The output is a **product** of both signals—a heterodyne product that includes the original modulation that has been carried by the sideband. All rf is filtered out by the pi-network, and clean audio is sent to the audio amplifiers.

balanced ssb detectors

You've already seen a simple single-sideband detector with characteristics ap-

fig. 3 Two-triode version exemplifies principles of sideband detection and is used commercially.



proaching those of a balanced modulator. The fact is, you can use a circuit very like a balanced modulator to demodulate sideband signals.

If you study the stage in **fig. 4A,** you will see that it differs only slightly from a balanced modulator. Both input transformers are rf types, whereas in a balanced modulator one of them would be an audio

the resulting audio signal while eliminating the carrier. The action in a balanced detector is thus very like the action in a balanced modulator; whatever signal is fed into the stage in a mode opposite from the output mode is canceled. This helps considerably in a ssb detector, since the carrier must be applied at a level so much higher than the level of the sideband.

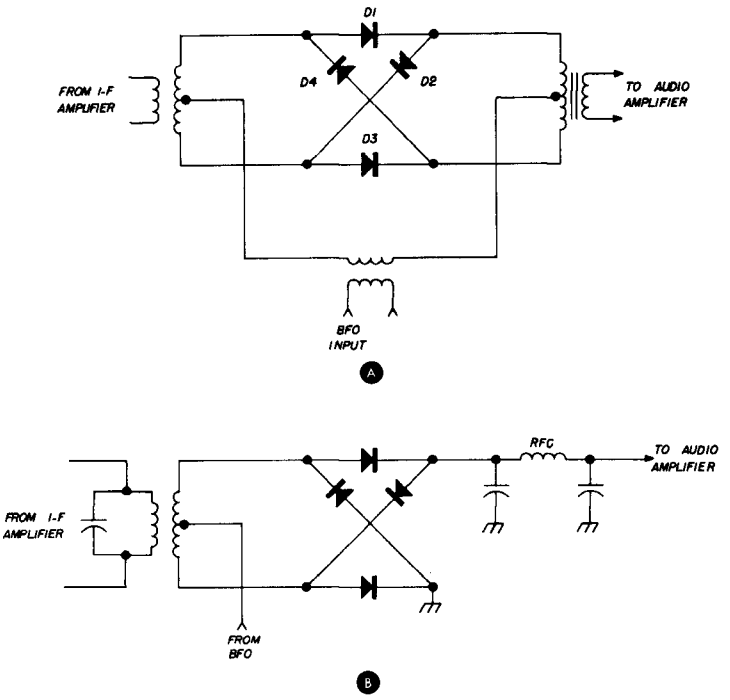


fig. 4. Ring type ssb demodulators. Primary version in A is balanced demodulator; eliminating the costly and bulky transformers doesn't alter stage operation (B).

type. The output transformer in **fig. 4A** is an audio transformer; in a balanced modulator it would be an rf type. What you see in **fig. 4A**, therefore, is a balanced demodulator.

The diodes are in what's called a **ring** arrangement; if you trace through them, you'll see they are essentially in series—'round and 'round. The name of the stage is **ring demodulator.**

Its operation is exactly what is needed to recover audio from sideband signals. It accepts the sideband i-f signal and the carrier signal (from a bfo or a carrier oscillator), reinserts the carrier so the signal can be demodulated, and then couples out

The ring demodulator can be simplified. Transformers are costly and bulky, and any circuit alteration that eliminates them has an advantage. An altered version is shown in **fig. 4B.**

Major characteristics remain. The solid-state diodes are hooked in a ring, the i-f sideband signal is applied in push-pull, and carrier signal is applied in parallel. With the bottom of the sideband-input transformer grounded (instead of the center tap), ground is made one side of a push-pull arrangement; the output is therefore effectively in push-pull, even though it is single-ended for any circuit following. The effect is thorough demodulation of the

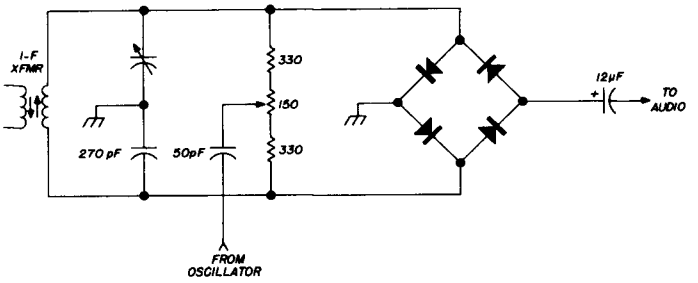


fig. 5. Commercial version of ring demodulator looks slightly different, but uses all principles of others.

sideband signal, with the carrier canceled in the output. The pi-network eliminates any slight rf that remains.

Does anyone use the balanced sideband detector? Yes. One version is part of the Sideband Engineers SB-34 transceiver. There's a schematic of the stage in **fig. 5.** I've redrawn the ring circuit to simplify the looks of the stage for you, but operation is the same as already described. The carrier signal, which in this case comes from the oscillator that generates the initial carrier for the transmit function, is fed to a resistive balancing network; the resistors also isolate it from the ring diodes.

The carrier is applied in the parallel mode, as you can see; the sideband input is push-pull, because of the "phantom" center-tap ground point offered by the ground connection between the two capacitors. In the ring circuit, input and output connections are the same as in **fig. 4B;** you'll see it if you trace them carefully, even though they may look different at first glance.

a one-transistor version

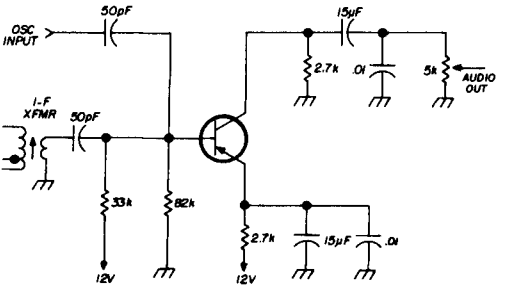
Diode sideband demodulators are all solid-state, since almost no manufacturer uses vacuum-tube diodes today. Semiconductor diodes are more efficient and less expensive. When you talk about solid state, though, you must include transistors. At least one manufacturer uses a transistor ssb detector.

You can see the circuit in **fig. 6.** This stage is from a Gonset Sidewinder transceiver. The pnp transistor is biased in a way normal for negative-ground power supplies—the emitter goes to the power-supply bus, and the collector goes to ground through its load. The two inputs are not isolated in this particular demodulator. The carrier signal is already amplified before it is applied to the transistor base (through the 50-pF capacitor). It and the sideband signal mix in the transistor.

What keeps this from being a simple amplifier for both signals is the bias level chosen for the transistor. The base-emitter junction is strongly backward-biased; the heavy carrier signal then is amplified class C, which is nonlinear.

Mixing in the base resistor as they do, these two signals generate considerable cross modulation. When the cross-modulation products are amplified by the Class-C

fig. 6. Transistor product detector, with no isolation between input signals. Bias of the transistor is what makes it a demodulator rather than amplifier.



transistor, the audio is easy to separate from the other products of this nonlinear mixer. The 0.01-μF capacitor across the volume control eliminates most of the rf signal that is left over. The original modulation, which has been masquerading as a sideband, is thus recovered.

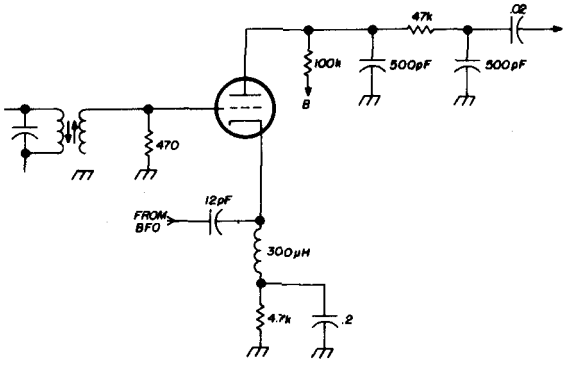
This transistor sideband detector hasn't become popular; no other set uses it that I know of. But transistors can be substituted in any triode-tube demodulator, provided you consider their dc supply requirements and their low impedance.

ssb detection with tubes

In any single-sideband detector system, isolation of the two input signals is desirable. One way to achieve this is in a simple triode product detector—**fig. 7.** The high-level signal from the bfo (or from the carrier oscillator) is fed to the cathode, using a 300-μH choke as high-impedance —and therefore efficient—input load. The i-f signal, which is the sideband to be demodulated, is applied to the grid, across a low-impedance load: the 470-ohm resistor. This disparity between the two input-load impedances goes part-way toward setting the 10-to-1 ratio you want between these two signal strengths.

This triode stage is another reminder of an important principle of product detectors. It isn't always the circuit arrangement that makes a stage detect sideband signals; it is the way the stage is operated. Without the high bias developed by the

fig. 7. Triode product detector with some isolation for signals. Same idea could be used with a transistor.



4.7k cathode-bias resistor, the triode would be nothing more than an amplifier. It would transfer both signals to its output, amplified but otherwise unaltered. It is the nonlinear operating characteristic that permits product detection—and therefore sideband demodulation.

The pi-network in the output of this triode single-sideband detector consisting of two 500-pF capacitors and a 47k resistor eliminates whatever rf products get through the detection process. Good rf filtering is more important in a detector stage like this than in a balanced type, simply because the balanced stage inherently keeps most rf from reaching the output.

The **fig. 7** circuit is popular because of its economy and simplicity. You'll find it in several Heathkit sets and in the Hallicrafters SR-2000 transceiver.

A tube version like the one I described in **fig. 3** is part of the Hammarlund HQ-180. The stage configuration is the same; the only differences are in parts values. A triode Colpitts bfo is used in the HQ-180 to furnish the carrier.

The other Hammarlund models revert to the single-tube product detector using a pentode: the HQ-110 and the HQ-145. There is no isolation between the two inputs; both signals are applied to the control grid. Some isolation is achieved by the weak coupling used for both signals. The strong carrier signal is applied through a 3-pF capacitor, small even in this service. The i-f sideband signal is coupled only by a twisted-wire "gimmick" capacitor offering less than 1 pF of coupling capacitance. The high gain of the pentode makes up for any expected weakness in the output—the demodulated voice signal.

using special tubes

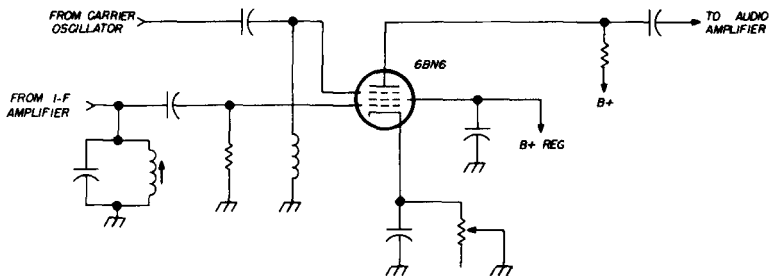
You read earlier that product detection is more how the tube is operated than what kind of circuit it's in. That being the case, imagination suggests that tubes with certain special operating characteristics could do an efficient job of demodulating single-sideband signals. That's right. One such tube is the gated-beam detector, a tube with pentode qualities and special construction that makes it particularly suitable for product detection. The beamed electron stream in a gated-beam tube is controlled by both the control grid and a special "gating" grid near the plate. Both grids have exceptionally linear control

over the electron stream, and very little effect on each other.

Combine these characteristics into the stage in **fig. 8** and you have a better-than-passable product detector. The sideband signal is applied to the control grid, or G1. The carrier-oscillator (or bfo) signal is applied to the special grid, G3.

The gated-beam stage is a little tricky to adjust. Unless the bias is just right for each particular tube, considerable output distortion is common. Designers also must carefully work out the strength ratios of signals applied to the two grids. Properly designed and adjusted, though, the gated-beam ssb detector does a good job.

fig. 8. Gated-beam detector, used for years in tv and fm receivers, can also make a good ssb detector.



An offshoot of the gated-beam idea is used in the Galaxy V Mark 2 transceiver. The circuit is shown in **fig. 9.** The tube is a 6GX6, a pentode specially designed for broadcast-receiver use in fm detectors. Its non-interacting quality between grid 1 and grid 3 serve sideband detection admirably. As you can see from the diagram, a crystal-controlled oscillator is formed by the cathode/grid 1/screen portion of the tube. The 6GX6 thus provides its own insertion carrier. Sideband signals from the i-f stages are applied to grid 3.

This tube, like the gated-beam detector tube, is touchy. Signal levels must be guarded to avoid crossmodulation that might upset output clarity. The diode between the i-f transformer and the tube input acts as something of a safeguard, to prevent overdriving grid 3. (The diode can't act as a detector because there is no carrier with the i-f sideband signal.)

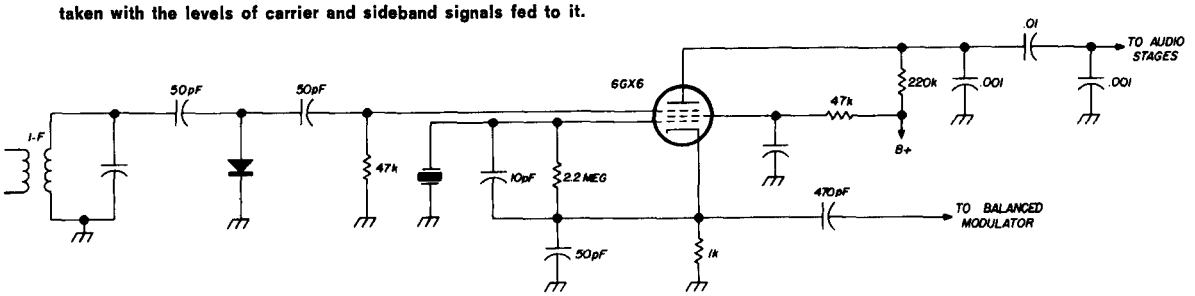
The connection going to the balanced modulator is shown because the oscillator portion of the 6GX6 circuit doubles—during transmission—as the carrier oscillator. That connection has no bearing on detector operation during recepiton.

beam-deflected ssb detection

If you did read the earlier article on balanced modulators, you may remember a rather unusual stage using a **beam-deflection** tube. The tube is an RCA 7360, and makes an efficient—though expensive—balanced modulator. This circuit can be altered slightly to become a balanced demodulator, as can other balanced modulators.

The sidebands are applied to the beam-deflecting plates in push-pull. (Supply circuits are not shown to keep the diagram

fig. 9. Another special tube, the 6GX6, makes an excellent ssb demodulator if care is taken with the levels of carrier and sideband signals fed to it.



simple.) The carrier signal is applied to the control grid, which puts that signal in parallel insofar as the rest of the stage is concerned. The demodulated audio output is taken off in push-pull by the simple device of grounding one output plate for audio through C4. (C5 is a different value because its purpose is to eliminate any stray rf signal that might be left over after demodulation.) Output is from the other plate, coupled to the audio amplifiers by C6.

in carrier and sideband signals instead of carrier and voice signals. A balanced modulator becomes a balanced demodulator by that simple switch; of course, you have to change one input transformer from an audio type to an rf type and change the rf output transformer to an audio type.

In demodulators of the product-mixer kind, you have to be sure the carrier signal is 10 or 20 times as strong as the sideband signal. That's because the carrier is always much more powerful than the side-

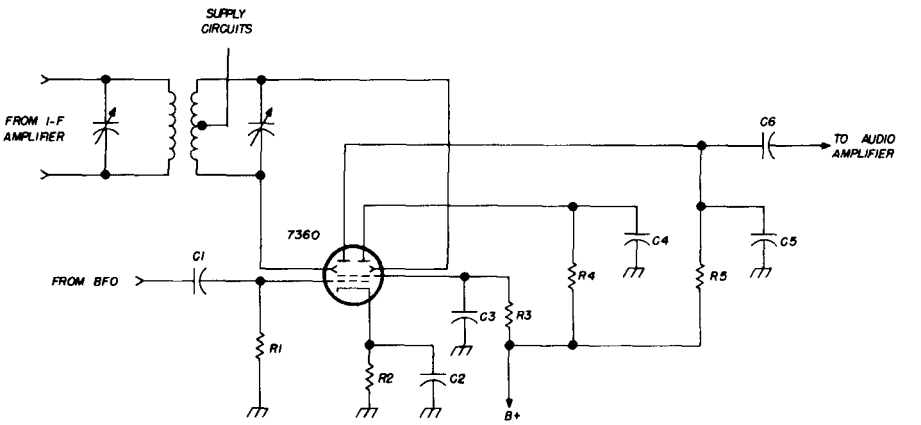


fig. 10. Tube developed especially for use in ssb communications; can be either a modulator or demodulator.

The beam-deflected method of ssb demodulation hasn't been used in any commercial ham equipment I know of. Its expense, though not great, is more than for diode balanced-demodulator systems; cost is always a deterrent to inclusion in factory-built equipment. The circuit is showing up occasionally, however, in home-brew designs. It is efficient, and you might want to try it in a receiver of your own. Operating characteristics for the 7360 can be found in The Radio Amateur's Handbook, in the Special Receiving Tubes table. From those, you can work out parts values.

sideband detectors in general

Summing up the characteristics of various single-sideband detectors, you can draw certain general conclusions. First of all, any ssb modulator can be altered to become a demodulator. You merely feed

bands in an ordinary a-m signal, and that relationship must be maintained for proper demodulation. If you're working up your own demodulator circuit, you should adjust the ratio between the two signals until you get the best output signal-to-noise quality; at the same time, keep both signals low enough in strength that one doesn't overload the tube(s) you're using.

With the information here, you should find that sideband detectors have few secrets anymore. You've seen both tube and transistor types, as well as solid-state diode types.

And—speaking of transistors—that's what we'll go into in the next article: transistors in single-sideband equipment. More transistors are being used, so there should be a lot of dope that will help you in your future ssb activities.

ham radio

solid-state antenna switching

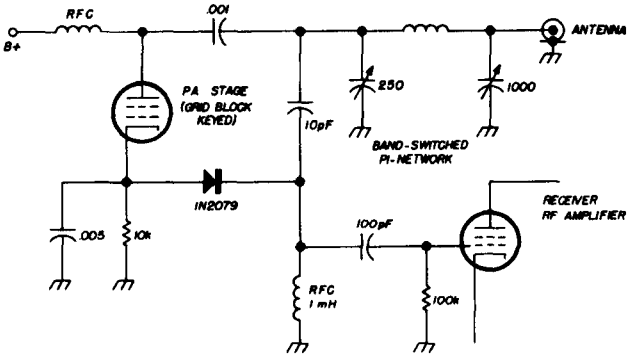
Solid-state antenna switching and how to adapt it to your own rig

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

Solid-state antenna switching has been widely advertised as a feature in a number of commercial transceivers. The advantages are numerous and the schemes used in transceivers can generally be adapted to your station.

One of the advantages of solid-state antenna switching is that the noise and contact problems are eliminated. Also, since the antenna transfer between transmitter and receiver is almost instantaneous, break-in CW is available with the

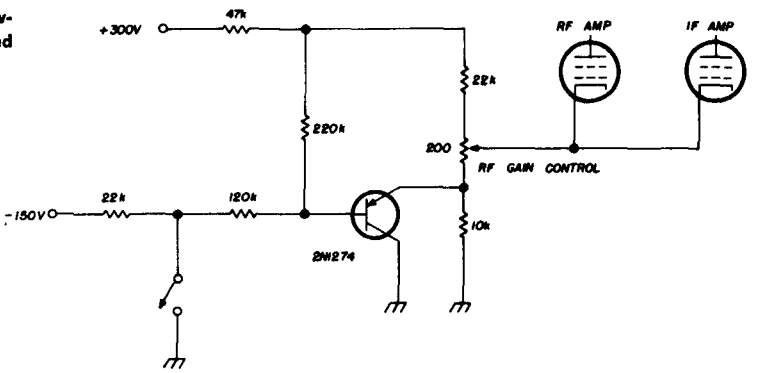
fig. 1. Basic circuit of the antenna-transfer circuit used in the Heath HW-16. The diode is a standard silicon rectifier rated at 500 V and 750 mA.



addition of a simple receiver-muting circuit. As a side benefit, the solid-state antenna switching circuit automatically protects the receiver front end from excessive input which may burn out the rf stage.

This article explains some of solid-state antenna switching circuits which have been used in commercial equipment. The circuits are relatively simple, and once their operation is understood, you should be able to develop a circuit that will operate with any transmitter and receiver combination.

fig. 3. Simple receiver muting circuit used in the Heath HW-16.

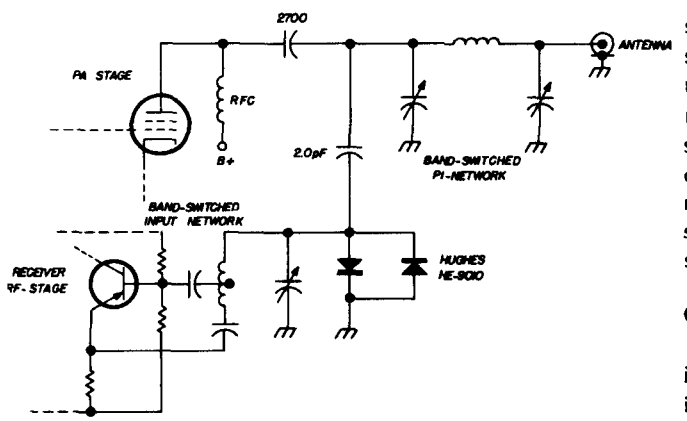


antenna switching

When you hear the term "antenna switching," you naturally think in terms of a relay where the center arm is connected to the antenna and completes the electrical path to the transmitter or receiver. However, this type of antenna switching is rarely used with solid-state equipment. Instead, the receiver is connected to the transmitter and antenna so that when the transmitter is not operating, a low-impedance path exists between the antenna and receiver. When the transmitter is turned on, the receiver is shunted to ground through a low-impedance path that does not affect transmitter loading.

The following circuits should make this clear. However, the idea of shunting rather than switching should be remembered to understand circuit operation clearly. In these circuits, the transmitter remains continuously connected to the antenna, and the receiver input is switched to either the antenna or ground.

fig. 2. Front-end protective circuit used in the SB-34.



If you don't want to use solid-state switching with diode switches and transistor muting circuits, a relay could be used across the receiver antenna terminals. Since the transmitter output is not switched, a small (and inexpensive) relay can be used, even for high-powered transmitters. The quicker switching action of a small relay permits break-in operation at slow to moderate CW speeds.

circuits

The basic antenna transfer circuit used in the Heath HW-16 transceiver is shown in **fig. 1.** It is typical of circuitry which can be adapted to other vacuum-tube

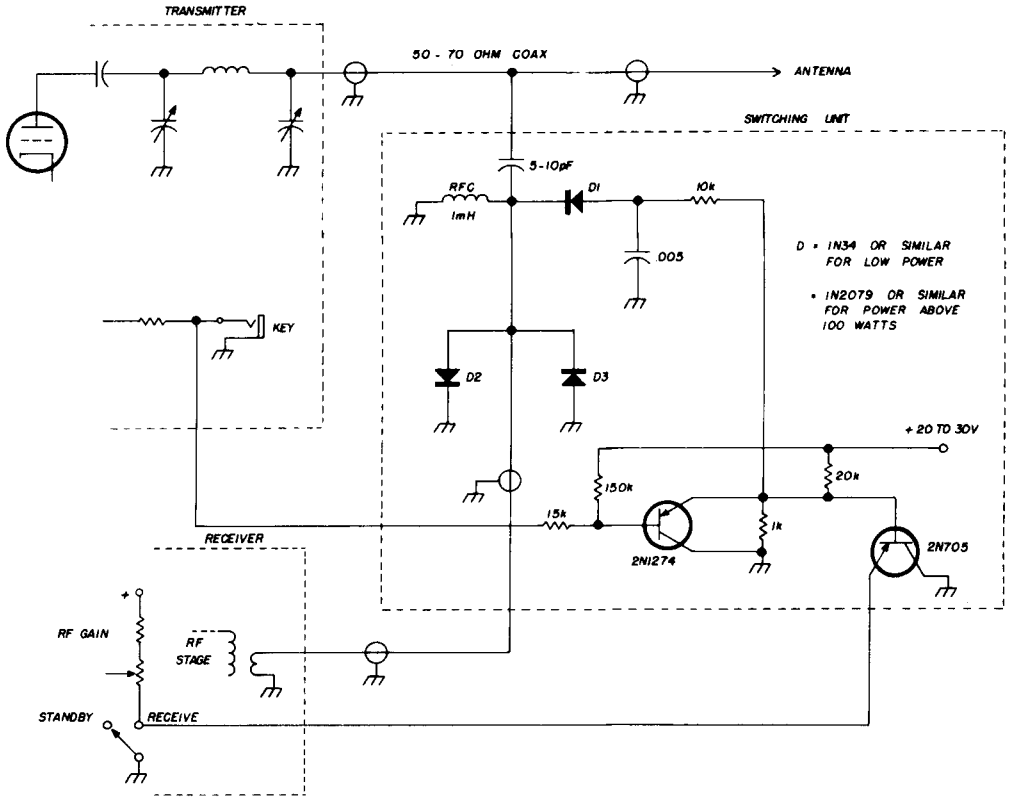
equipment. Circuit operation is relatively simple; grid-block keying is used, so when the key is open, the power amplifier is cut off. The voltage appearing at the PA cathode is almost zero—not enough to appreciably forward bias the diode. Therefore, a high rf resistance to ground appears along the receiver line (the rfc presents a constant high-impedance path). The antenna signal flows from the pi-network through the 10- and 100-pF coupling capacitors to the receiver rf amplifier.

When the transmitter is keyed, the PA cathode voltage rises to about 11 volts. This forward biases the diode into the low-resistance region so the receiver is essentially grounded for rf through the .005-μF cathode by-pass capacitor. The 10-pF capacitor is now in parallel with the 250-pF plate-tuning capacitor but produces no noticeable detuning effect.

Fig. 2 shows how antenna switching is accomplished in the SB-34. The receiver input is loosely coupled through a 2-pF capacitor to the pi network. During transmit periods the diodes operate as clippers and reduce the rf voltage across the receiver input circuit to a low value that will not damage the transistor rf stage. The diodes are high-speed computer types manufactured by Hughes.

Regardless of the method used to protect the receiver input during transmit periods, it must work in conjunction with a muting circuit that silences the receiver. One circuit that will do this is shown in **fig. 3**. When the key or push-to-talk switch is open, the transistor switch is biased into conduction and the lower end of the 200-ohm gain control is effectively connected to ground.

fig. 4. Switching unit for separate receiver/transmitter setups.



When the switch is closed, the 220k and 120k resistors form a voltage divider which puts about 30 volts on the transistor base; then the emitter-collector path is essentially an open circuit. A high positive voltage is placed on the cathodes of several receiver stages and they are cut off.

You can easily adapt the basic concepts of these circuits to suit your own equipment. If an adequate receiver-muting circuit is available, you only have to add the simple diode circuit shown in **fig. 1.** If you want to develop a completely new circuit, you can combine some of the ideas.

An example of a combined circuit is shown in **fig. 4.** This circuit is for use with a separate transmitter and receiver. Some advantage might be gained by coupling the receiver to the transmitter before the pi-network to take advantage of the selectivity added by the extra tuned circuit. However, the constructional difficulties involved when the transmitter and receiver circuits are in separate boxes usually outweigh the gain.

Operation of the circuit is a combination of **fig. 1** and **2.** Diode D_1 presents a high or low impedance to ground depending upon how it is biased by the 2N1274. Diodes D_2 and D_3 are additional protective circuitry for the receiver; they are generally not required for transmitters with less than 100-watts output.

In the receive mode, D_1 presents a high rf impedance to ground. The negative blocking voltage on the key biases the 2N1274 into conduction; this returns the base of the 2N705 essentially to ground so it can conduct and complete the circuit for the receiver rf gain control.

When the key is closed, bias is removed from the 2N1274, and a positive bias is applied to D_1 thus presenting a low-impedance path to ground. The 2N705 is biased so that its emitter-collector circuit presents a high resistance path and the receiver is muted.

construction

Whatever circuit you devise, care in construction is necessary for proper performance. The rf leads should be kept short and shielded although there is no particular restriction on any low-impedance coaxial-cable leads.

The circuit shown in **fig. 4** can be conveniently built and shielded in a small mini-box. If you use a small chassis, it's a good idea to put the rf components at one end of the enclosure and the control transistors and dc connections at the other. Muting can be checked by grounding the receiver input and noting that when the transmitter is keyed the receiver is silenced.

The protective circuitry for the receiver can be checked by measuring the rf voltage across the coaxial line to the receiver (with the receiver disconnected) under keydown conditions. There will be some voltage, but it should be far below the level that would ever damage the receiver circuitry.

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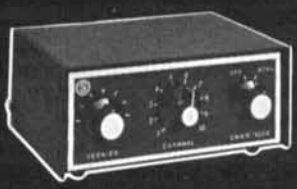

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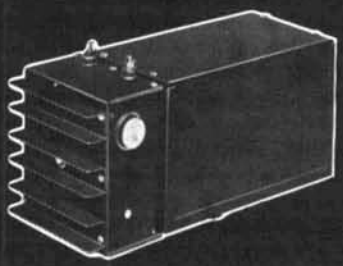

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pi and pi-L networks for linear-amplifier service

Almost all rf power amplifiers that are used with coaxial feedline use a pi or pi-L network to match the antenna; here's a simple approach to designing these networks

William I. Orr, W6SAI, Eimac Division of Varian, San Carlos, California 94070

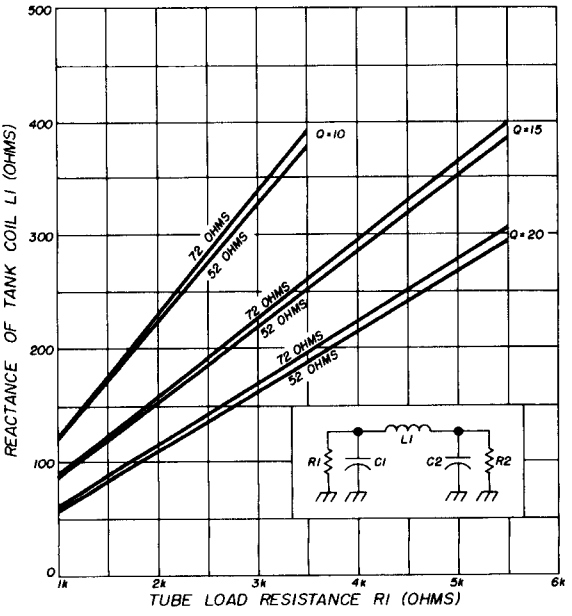
These graphs may be used to determine the component values used in pi and pi-L networks. The graphs cover the most generally used operating Q's, load resistances and antenna impedances. To use the charts, it's only necessary to know the plate voltage, peak plate current, desired operating Q and transmission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

using the pi-network charts

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

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

fig. 1. Tank-coil reactance as a function of tube load resistance for pi networks. R1 is the tube load resistance and R2 is antenna resistance.



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

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

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

2. Determine the approximate resonant load resistance from:

$$R_1 = \frac{\text{plate voltage}}{2 \times \text{plate current in amperes}} \quad (2)$$

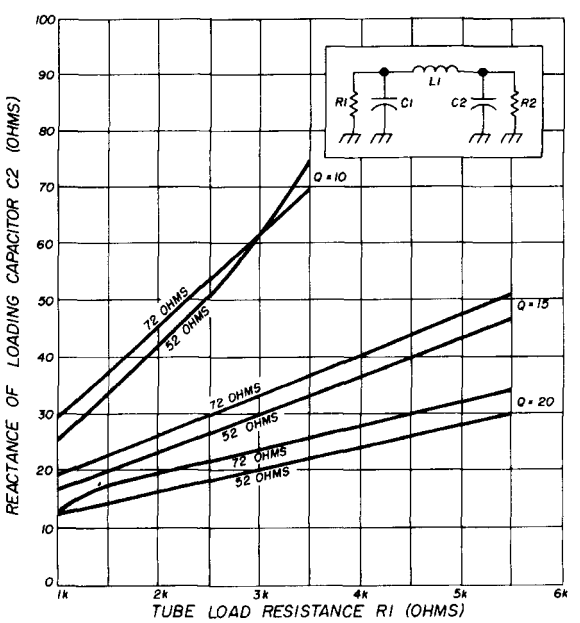
For the case of the 3-400Z's, the load resistance is: *2500/(2 x 0.8) = 1560 ohms.*

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

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

5. Find the reactance of the pi-network coil from **fig. 1.** For the case of two 3-400Z's operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the coil is approximately 120 ohms.

fig. 2. Reactance of the loading capacitor C2 as a function of tube load resistance for pi networks.



6. Find the reactance of the loading capacitor (C_2) from **fig. 2.** For the case of 3-400Z's operating with a load resistance of 1560 ohms and a Q of 15, the reactance of the loading capacitor is about 20 ohms.

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

For two 3-400Z tubes operating at a plate potential of 2500 volts with a peak plate current of 0.8 ampere (two kilowatts PEP) and a Q of 15, the values of the pi network plate circuit are: tuning capacitor (C_1) = 100 ohms; loading capacitor (C_2) = 20 ohms; pi network coil (L_1) = 120 ohms. As a quick check, note that the sum of the reactances of the two capacitors is equal to the reactance of the inductor.

8. Determine the capacitance and inductance values for the pi network. **Fig. 7** and **8** show reactance values of inductors and capacitors commonly used in rf circuits in the h-f amateur bands. For the reactances determined for the 3-400Z tubes, the circuit components may easily be determined for each amateur band. In the case of the 20-meter band, for example, the values are: tuning capacitor (C_1) = 100 ohms = 113 pF; loading capacitor (C_2) = 20 ohms = 565 pF; pi network coil (L_1) = 120 ohms = 1.36 μH.

using the pi-L network charts

Fig. 3, 4, 5 and **6** are used to determine pi-L network components.

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

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

Peak envelope plate current = 0.667 ampere

2. Determine the load resistance, as out-

lined previously in **eq. 2:** load resistance (R_1) = 2250 ohms.

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

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

5. Find the reactance of the tank coil (L_1) from **fig. 4.** For the case of the 3-1000Z operating with a load resistance of 2250 ohms, the reactance of the coil is approximately 215 ohms.

6. Find the reactance of the loading capacitor (C_2) from **fig. 5.** In this case, the reactance is about 47 ohms.

7. Find the reactance of the tuning capacitor (C_1) from **fig. 3.** In this case, the reactance is about 150 ohms.

8. Find the reactance of the loading coil (L_2) from **fig. 6.** In this case, the reactance is about 140 ohms.

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

9. Determine the values of the capacitance and inductance for the components of the pi-L network. **Fig. 7** and **8** show reactance values for inductors and capacitors in the range commonly used for rf circuitry in the high-frequency amateur bands. For the reactance values determined for the 3-1000Z tube, the circuit components may be easily determined for each amateur band. In the case of the 80-meter band, for example, the values are: tuning capacitor (C_1) = 150 ohms — 275 pF; loading capacitor (C_2) = 47 ohms = 900 pF; pi network coil (L_1) = 215 ohms = 9 μH; L network coil (L_2) = 140 ohms = 6.5 μH.

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

ham radio

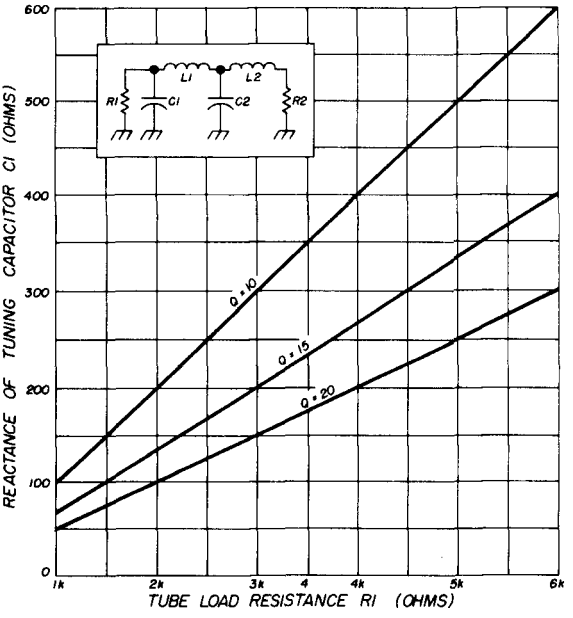


fig. 3. Reactance of the tuning capacitor C1 as a function of tube load resistance for pi and pi-L networks.

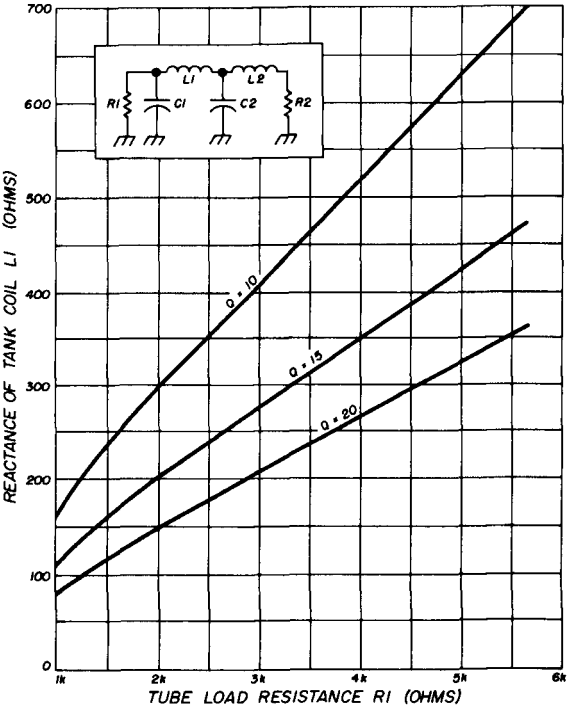


fig. 4. Tank-coil reactance (L1) as a function of tube load resistance for pi-L networks.

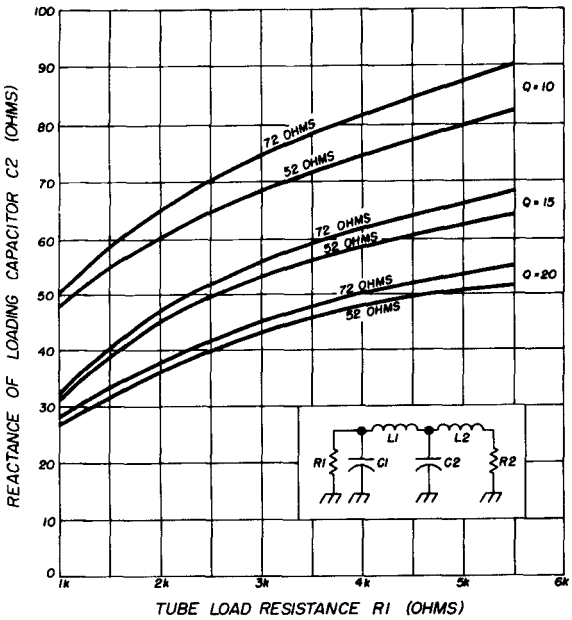


fig. 5. Reactance of the loading capacitor C2 as a function of tube load resistance for pi-L networks.

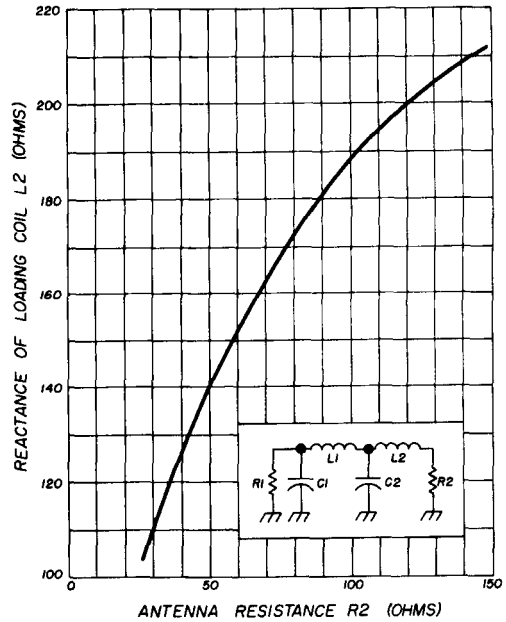


fig. 6. Reactance of the loading coil L2 as a function of antenna resistance (R2) for pi-L networks.

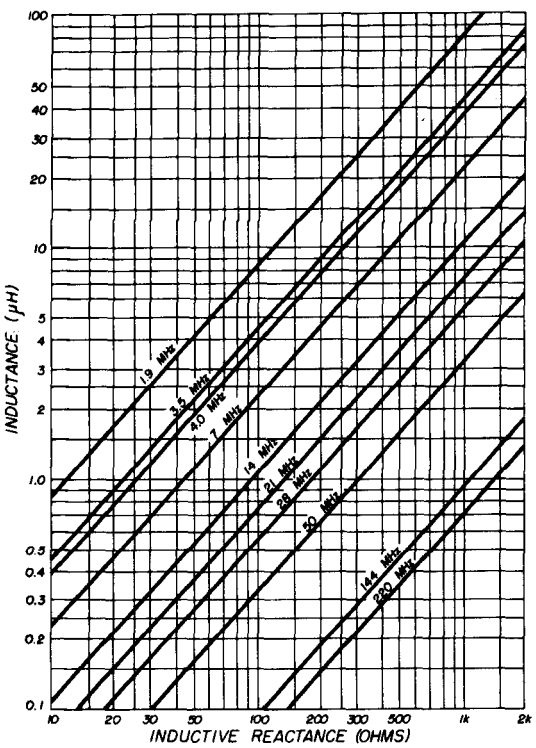


fig. 7. Reactance of inductors commonly used in the amateur bands from 1.9 to 220 MHz.

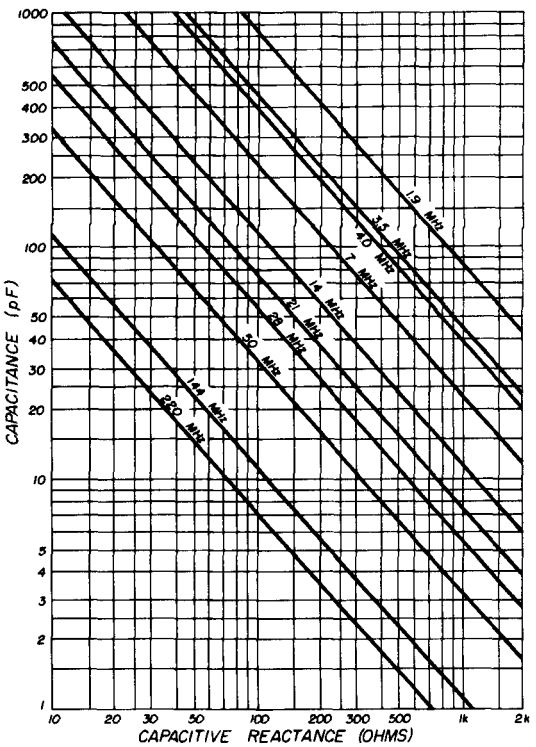


fig. 8. Reactance of capacitors commonly used in the amateur bands from 1.9 to 220 MHz.

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calibrators and counters

How to use inexpensive integrated circuits for accurate frequency-measuring equipment

Bill Conklin, K6KA, Box 1, La Canada, California 91011

I discussed amateur frequency measurements last month[1] and concluded that a temperature-controlled crystal in a calibrator is a pleasure to own, and calibrator harmonics every five or ten kHz are almost a necessity. I also suggested that a frequency counter is very useful for frequency measurements with accuracy better than one or two hundred Hz. These functions can be provided with convenience and simplicity by using the nonlinear integrated circuit (IC) known as a **J-K flip-flop** in single units, duals or decade dividers.

Building a calibrator and counter is so simple that it's completely dwarfed by one's inertia or stagnation in getting started. This is particularly true because we don't always know what we should about the IC's and must spend more time studying them than building. It is the purpose of this article to assist you in the planning job, and to demonstrate how easily the construction can be accomplished afterward.

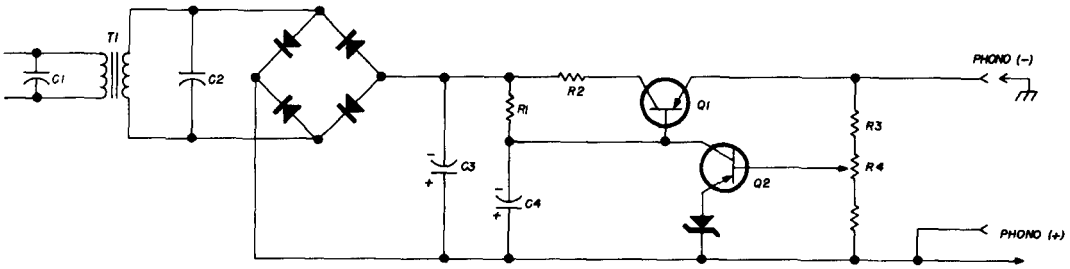
taming regulated supplies

Regardless of the voltages required, there is usually a power supply to design. Nonlinear IC's are not very critical as to voltage and don't need closely regulated power. However, there should be good regulation of the power to the crystal oscillator and a way of protecting

other units from overvoltage in case of loss of load. Unfortunately, most available transformers don't furnish the desired voltage; this leads to the need for a way of reducing the output without having it soar in the absence of part or all of the load. The simple answer to this is some form of regulation which can be as little as a zener diode and resistor. Articles discussing transistor-regulated supplies, or practical circuits, are listed in references 2 through 19.

It is not unusual for surplus zener diodes to vary considerably in their performance, to operate in reverse or not at all. More than one may have to be purchased for a particular job. On the other hand, it's not much more expensive to build a series-regulated supply which can be adjusted over a considerable voltage range. Taming one or two of these turned out to be an education and caused a delay in the much shorter job of wiring up the IC's. Some comment on **fig. 1** may prove helpful.

Many sources say that there should be a resistor in the transformer lead to protect the rectifier diodes. I found that the transformer resistance was sufficient—provided the current rating of the silicon rectifier diodes is around an ampere. These have always survived the short-circuit current through a very large surplus electrolytic filter capacitor. Good diodes are available as "glass amp" and other types at a few cents each from John Meshna, Poly Paks, Weinschenker and others. Be sure to test them; some operate in the reverse direction.



	5 V	12 V
T1	12 V	24 V
C1	.1	.1
C2	.01	.01
C3	large	large
C4	100	100
R1	350	220
R2	25 ohm, 25 W	200 ohm, 10 W

	5 V	12 V
R3	68	1000
R4	—	350
R5	120	600
D5	3.3 V	5.7 V
Q1	2N301	2N301
Q2	2N404	2N301

fig. 1. Regulated power supplies

table 1. J-K next-state truth tables showing output after next clock pulse. A applies to MC853, 9093, LU321A and 8822; B applies to MC790/890, μL923 and μL926.

J	K	Q_{n+1}
0	0	Q_n
1	0	1
0	1	0
1	1	$\overline{Q}$*

A

J	K	Q_{n+1}
1	1	Q_n
1	0	1
0	1	0
0	0	$\overline{Q}_n$*

B

* **Toggle condition to divide by two.**

Unless your filter capacitor has a grounded mounting, either pnp or npn transistors can be used by putting the proper polarity on them and grounding either power supply output. The selection of transistors is covered in theory and

table 2. Binary-coded decimal count-sequence truth table.

	output			
count	D	C	B	A
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1

arithmetic in the 12-page pamphlet, "Transistorized Voltage-Regulator Application Guide—1CE-254," available from RCA distributors. Practical circuits are found in the **GE Transistor Manual.**

Surplus 2N404-type transistors, Poly Paks number 14L492, are satisfactory for the amplifier job in the supplies that do not require a drop of more than ten volts or so. A good selection of low-voltage ten-watt zener diodes is available from Solid State Sales, so smaller ones may be avoided. The temperature-compensated GE type-RA1 reference amplifiers are also attractive for output voltages of 8 or more with precision control.

A bag of Poly Paks number 14L404 transistors provides the large ones for handling the series current and as amplifiers where a large voltage drop is involved. These are similar to the RCA 2N2869/2N301 three-ampere, 30-watt audio device in a TO-3 case. They may be used in the Darlington-configuration to increase the effective h_{FE}; with emitter-equalizing resistors they can also be used in parallel to handle more current.

Nothing here has more than taken the chill off of them. Don't forget to take all the burrs off the chassis holes and check for case-to-chassis shorts if they are mounted with mica kits.

The rectifier usually puts out about twice the ultimate voltage in order to provide room for control. The greater this voltage difference, the warmer the transistors get.

The resistor between the series transistor collector and its base must not be too low or the amplifier transistor will get hot. A high-wattage resistor directly at the series transistor's collector will provide short-circuit protection, which is a necessity. Its size depends on the maximum current to be taken from the power supply. See GE's manual.

The zener diode should have its resistance checked in both directions. As a test, put a large resistor in series with it and feed it some dc so the regulating performance of the zener can be measured

table 3. Some available J-K flip-flops and related devices.

	frequency limiting min	frequency limiting typ.	type	cost	V_{CC}	J-K to toggle
Motorola						
HEP558			single	$5.99	6	
MC1013P		85	single	4.55	—5.2	L
MC1027P		120	single	9.60	—5.2	L
MC1018P			translator	3.45	—5.2	
*MC838		30	decade	7.55	5	
MC853		12	dual	4.75	5	H
*MC790P	8		dual	2.00	3.6	L
MC890P	8		dual	2.30	3.6	L
Texas Instruments						
SN7490N	10	18	decade	11.10	5	
SN7473N			dual	5.90	5	
SN7476N			dual	6.20	5	
Fairchild						
9093		2	dual	7.45	5	H
9958		2	decade	11.20	5	
9960			decade	15.70	5	
μL923		2	single	1.50	3.6	L
μL926	8	20	single	4.50	3	L
Signetics						
*LU321A	10		dual	2.48	5	H
*N8280A		35	decade	8.90	5	
*N8822A	10	25	dual	5.90	5	H
N8826A	25		dual	5.90	5	H
N8829A	15		single	3.65	5	H
N8H22B	50	75	dual	15.30	5	
8T01B			decade	15.00	5	

* attractive

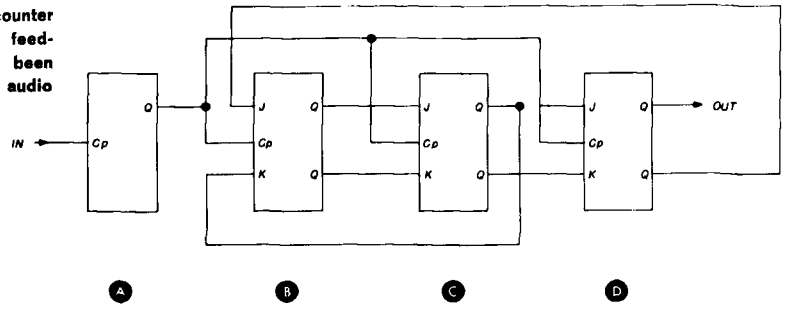
with both voltage polarities. The voltage drop across a reference zener may be around 40 to 80 percent of the desired power-supply output voltage.

The divider across the output can be selected to provide a light bleeder load together with a tap that will determine how much higher the output voltage will be compared with the zener voltage. A potentiometer connected in series with this divider permits adjusting the output; this is desirable during tests but can be avoided in the final assembly unless variation is required later, such as to provide a fine-frequency adjustment of the crystal oscillator. It is a little difficult to pick the two resistors in the divider because of resistance inaccuracies.

The power supply for the 6-V lamp count-indicator can be simple, as shown in **fig. 2.** It has a resistor, and 5- to 5.7-V ten-watt zener mounted on a bent sheet-aluminum angle as a heat sink. It is used only as a limitation on the maximum voltage that will reach the lamps. Their life depends on operating below their rating by at least a few per cent. They work fine with 4.5 to 5 volts on them. This arrangement permits using a 6.3- or 5-V transformer for 6-volt lamps. I used an expensive transformer (Lafayette 33H-3702, $1.39), although a new-condition Navy surplus one now furnishes all the power required for the equipment.

Some of the zeners may produce rf interference. A ceramic or mica capacitor directly across an offending zener diode will prevent this. Both polarities of each power supply are terminated in phono jacks for occasional use on experimental equipment. A shorted phono plug in the negative jack, captivated to a soldering lug under the jack, completes the connection to the crystal oscillator, the calibrator/time-base and the indicator lamps.

fig. 3. This decade-counter connection without any feedback NAND gates has been tested successfully at audio frequencies.

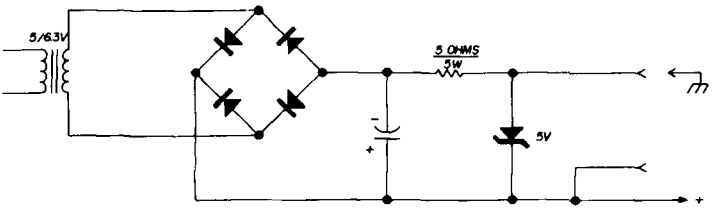


preliminary explanations

To help you select the IC's, some explanations are in order. Let's go through this. The arithmetical notation 1•2•3 or A•B•C refers to AND gates which occur at the input of some flip-flops. All inputs must be activated before a pulse can pass through the unit.

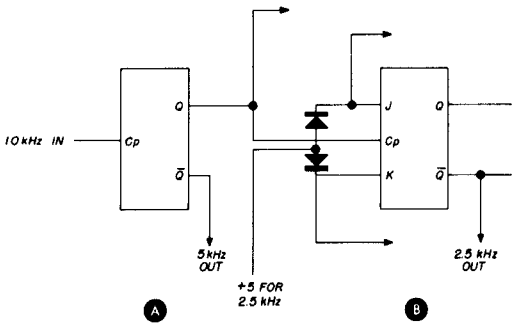
When you see 1+2+3, it refers to OR gates, any input to which will pass to the output—but this may not concern us here. The AND gates are built into J-K

fig. 2. Voltage-limited indicator-lamp power supply. Use phono plugs for output connectors.



flip-flops to make them flip in alternate directions from successive input pulses on a single terminal, which is referred to as **toggling.** This is the condition that divides by two.

fig. 4. Proposed method for inhibiting the decade count with +5 volts through diodes to the J and K inputs on flip-flop B in fig. 3.



It is well to recognize some of the designations used on the terminals. C_p and T mean clock pulse or input which is fed to both sides of the flip-flop to cause it to toggle (usually on its falling or negative edge) and divide by two when other inputs permit it. J and K inputs determine whether it will obey the next clock pulse. S and C, R and S, R_c and S_c, all refer to somewhat similar inputs. R, S and C are also used to indicate reset, set, and clear terminals, frequently with a subscript letter D to indicate direct action on the flip-flop regardless of the clock pulses or the other determining inputs.

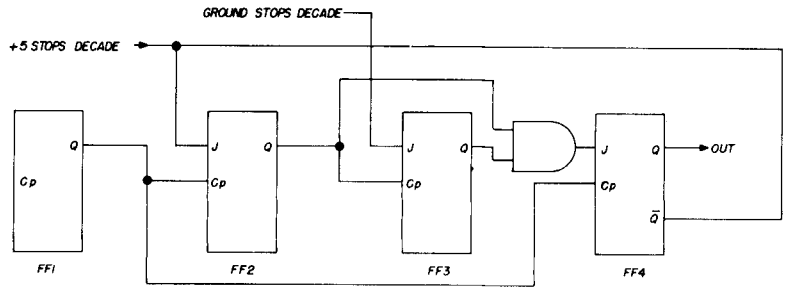
R_0 sometimes refers to reset to a binary zero, and R_9 to a binary count of nine, the latter being of no interest to us. Q or 1 is the normal output; $\overline{Q}$ (not Q) or 0 is the reverse of the normal Q or 1 output. V_{CC} is the power input to the collectors which is generally referred to as plus compared with ground; V_{EE} is also used in (M)ECL units other than the HEP558 where the power supply is minus on V_{EE} (the emitters) and V_{CC} is grounded.

Truth tables indicate what conditions will produce what results.[3] The J-K next-state truth table in **table 1** shows whether the auxiliary J and K inputs should be at high or ground voltage to cause the FF to toggle with successive clock pulses. There are several different types of count-sequence truth tables; the one in **table 2** shows the indicator lamps that light in a simple additive 8, 4, 2 and 1 binary-coded decimal.

Another[25], has more lamps on and deviates slightly from the simple addition of binary numbers to convert from the lighted lamps on an additive basis to the digital numbers of zero through 9. Still another differs greatly[19], tending to limit its use to the provision of decoding gates for full 0-to-10 decimal indication.

For frequency division, several kinds of FF's will work. The inclusion of AND gates to obtain decade operation is another consideration that makes the J-K flip-flop desirable.

fig. 5. A decade counter circuit using Fairchild 9093 flip-flops that requires an AND gate. Two ways are shown for stopping decade action to obtain a 2.5-kHz output from the 10:1 decade.



All counting flip-flops and the time-base gate must reset to zero just before a count starts. Thus, a reset or clear provision is required. However, if a flip-flop has only a **set** provision, this can be used to reset or clear simply by reversing the J and K inputs and the Q and $\overline{Q}$ outputs.

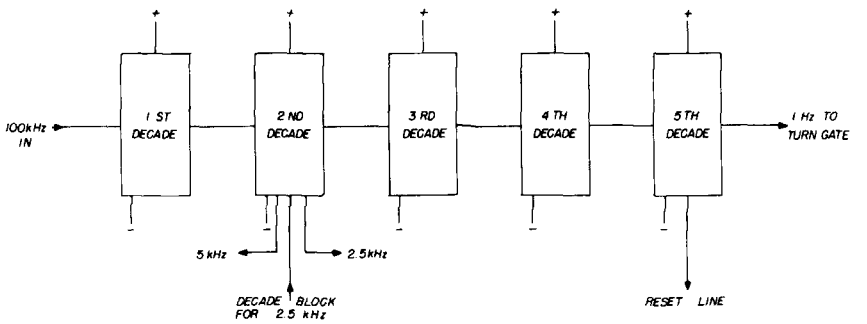
You should also consider the count-sequence truth table that will be produced by a decade divider using a particular type of IC and circuit, because of types of interpretations that may have to be applied to the indicator lamps. Two kinds can be mixed if you don't mind a little confusion.

other points to consider in mixed types.

Table 3 lists many of the available J-K flip-flops, decade counters, translators (if (M)ECL high-speed FF's must feed into the more common saturated-logic units) and decode/lamp drivers for direct conversion to full decimal read-outs. Inasmuch as the prices frequently decline 20 percent for quantities of 25 or more, the single and the dual J-K flip-flops for an entire calibrator and counter may cost less than indicated.

The MC790P operate only down to 15° C; many may prefer the MC890P which operates to freezing. The 9093's that I used operate

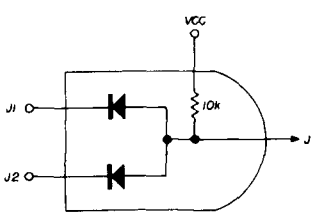
fig. 7. Block diagram of the calibrator time-base board. A shorter count delay results when the reset connection is moved one decade to the left.



available units

It is a convenience to have the IC's operate at the same voltage. The upper frequency limit desired for a counter may lead to selecting a more expensive first FF or decade, which could have a different requirement. If so, two or more series-regulator power supplies can be fed from the same rectifier. There will be some

fig. 6. Simple diode-resistor AND gate used with the circuit of fig. 5 to obtain decade operation.



from sine, sawtooth and square waves down to several cycles per second. The 8822 made by Signetics Corporation operates independently of fall-time of the clock pulse. However, most of the other units have specifications on fall-time which may require some form of trigger or squaring amplifier ahead of the first calibrator and counter FF's for proper operation. See references 2, 4, 13, and 20 through 24. Try them first without it, and let me know what happens.

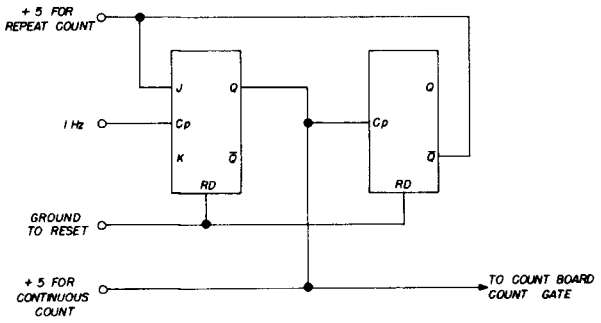
using the flip-flops

The most simple IC for calibrator/time-base and counter use is the decade. As indicated in **table 3,** these cost several dollars more than two dual J-K flip flops. They do not give access to the J-K terminals, so some provision must be made in the first counter decade to gate it. Other than putting a gate in front, an alter-

native is provided by the Texas Instruments SN7490N where the divide-by-two FF can be wired for use elsewhere; a separate high-speed FF with J-K inputs can be used as the initial divide-by-two part of the decade ahead of the remaining divide-by-five part. On the other hand, a quad-gate IC will do this job and do three other jobs elsewhere if desired.

You'll note that some units require the extra inputs to be grounded, and some require them to be brought to the V_{CC} or left open to toggle and divide the frequency. These two families require different frequency-division circuitry. Several articles have appeared giving the inter-

fig. 8. Time base and inhibiting connections.



connections for these types. See references 3, 4, 5, 13, 19 and 21.

In general, there must be a connection (internal or external) between the input and the output such as from J to $\overline{Q}$ and from K to Q to assure toggling. If no clock-pulse input is provided, the J and K terminals generally are connected; the input signal is fed to both of them.

The single or dual J-K flip-flop for decade operation requires some form of a feed-back connection so that four FF's will be fooled into counting to ten instead of sixteen. **Fig. 3** shows a circuit with 9093 dual J-K flip-flops I tested successfully in the audio range. It's probably satisfactory for all decade applications. If the 10-to-1 kHz calibrator/time-base decade uses this circuit, and if 25-kHz output is desired, it can be obtained from the second FF by connecting the J and K inputs of this unit to V_{CC} except during counting.

If only one of each of these inputs is available, two separate switch contacts might be avoided by feeding the plus voltage through separate diodes to the J and K terminals so that no back voltage can feed across. This is shown in **fig. 4.**

The above decade circuit is complicated by many J and K input connections. I will explain another divide-by-two and divide-by-five circuit to accomplish division by ten with 9093 FF's that I use. Refer to **fig. 5.**

This type operates by inhibiting an input to FF1 at the ninth and tenth counts, and by resetting FF4 on the tenth count or after every even-numbered count. FF2 and FF3 act normally through the 8th count. Then, the $\overline{Q}$ output of FF4 puts a 0 voltage on the J input of FF2, holding FF2 and FF3 at 0 during the tenth count when FF1 and FF4 also go to 0, ready to start at the beginning. Inasmuch as FF4 is fed from the output of FF1, but must not flip prior to the 8th count, an AND gate is placed at the J input of FF4.

This prevents FF4 from going to 1 until after both FF2 and FF3 are also at 1 at the sixth and seventh counts. **Fig. 6** is the simple AND gate I used and consists of inexpensive surplus diodes and a resistor of about 10k ohms. Some FF's already have extra inputs to do this. The gate must be capable of operating up to about one-half the input frequency of the decade.

A ground on the J input of FF3 (or +5 V on the J input of FF2) in **fig. 5** will stop the decade action so that 5- and 2.5-kHz outputs can be obtained from the Q or $\overline{Q}$ outputs of FF1 and FF2 in the 10-to-1 kHz calibrator/time-base decade for frequency-measuring purposes.

The equipment I use has a Monitor Products oven-controlled crystal oscillator with its output fed directly to the first calibrator/time-base FF. The output of that is fed to the input of the next, as shown in **fig. 7.** Wiring was done with small, flexible tone-arm twisted pair until the supply ran out, then with Japanese earphone twisted pair, which is larger and retains its kinks.

When the frequency has been divided to one cycle per second (1 Hz), it's necessary to add one more flip-flop to produce a time gate that opens for just one second, putting "toggle" voltage on the J and K inputs of the first count FF, then closes for one second. It is also desirable to provide still another FF (the second half of a dual unit) to permit locking up the time gate after the one second that it is on, so that the indicator lamps can be read.

After reset, the first negative slope from the 1-Hz decade in the calibrator/time-base board turns on the time-gate FF, putting plus voltage on the J and K terminals of the first count FF, which had been held low after the gate was reset. The second negative slope turns it off one second later, but this flips the one-shot FF so that its $\overline{Q}$ output goes to 0. This is connected to the time gate's J input and thus stops any further gating until reset. **Fig. 8** shows this circuit and the options provided by the count-selector switch.

I found that feeding the crystal oscillator output to a receiver caused a slight shift in frequency. As a result, the first output to the harmonic-selector switch is connected after at least one FF. The 10-kHz signal is available, and higher frequencies, but 5 kHz is more useful in my Collins receiver. A 2.5-kHz signal is also fed to the harmonic-selector switch, but produces 1 kHz during a counting condition of any type—repeat one-second counts, one-shot count, or continuous count.

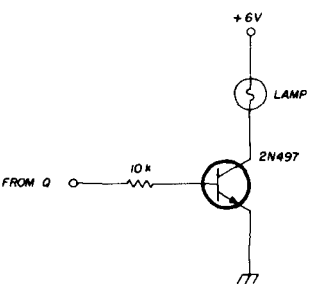
The harmonic-selector switch connects a ground to the J input of the third FF of the 10-to-1 kHz decade, stopping the divide-by-ten operation in the reset and indicator-lamp-off positions of the count-selector switch, thus making the decade divide-by-two twice, producing the 2.5-kHz output from the 10-kHz input. This is desirable with selective receivers to ensure that a suitable beat note is available at every frequency, including gaps left by lsb and usb selection in the receiver. The 1-kHz output during a counting period is mainly a novelty, but its 30,000th harmonic is strong and could be used for 1-kHz receiver calibration.

The crystal-oscillator output lead should not be cabled with ac wires unless you use RG-174/U. It might be just as well for the other leads to the harmonic-selector switch and the receiver's antenna input to be coaxial cable or kept out of the cabling, although this has not caused any trouble so far.

electronic counter

The frequency to be counted, such as an audio oscillator, was originally brought directly to the first J-K input of the first FF to provide this function when being driven directly by the time-base gate. It was found that a subsequent FF can also be gated to prevent strong input signals from activating subsequent decades when the time gate is off.

fig. 9. Lamp driver and indicator-lamp circuit.



Additional decades are added until the output reaches the maximum read-out frequency at which the counter will operate at if anything beyond audio frequencies is of interest. The first decade can be an especially high-frequency unit, and it can be selected so that its first FF is capable of toggling at the highest frequency you're interested in.

It's necessary to provide a reset or clear function for the counting decades so that the count will always start from zero. It is also convenient to connect this reset line to the time-base gate and to the associated one-shot FF that locks it up when a single one-second count is de-

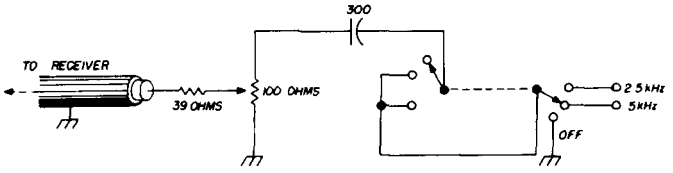
sired. Furthermore, the line can go to some of the decades in the calibrator/time-base board to set or reset them to delay opening the gate until all switching transients (if any) have decayed.

count indicator

The FF outputs from each decade can be decoded with diodes and resistors to drive numerical neon read-out tubes costing $8 or more per decade. IC decoder-drivers are listed in **table 3.**

After much consideration, I used the simple, convenient and inexpensive indicator suggested by Phil Brassine, K7UDL. This has four lamps per decade, the lamps from left to right indicating 8, 4, 2, and 1 in binary-coded decimal form as shown in **table 2.** These can be converted to a single digit by inspection, or by adding these digits. Then, more of these four-lamp sets are added, one for each decade to be indicated. Six sets will count to a million, or 1MHz. This is satisfactory for slightly higher frequencies, simply by watching the number of times that the count restarts during the one-second counting time. A seventh bank should suffice up to 30 MHz.

fig. 10. Harmonic selector switch with provision for "anti-leak".



The lamps are low-current Sylvania 6ESB six-volt slide-base lamps at 40¢ each. The unbased lamps at 31¢ would also work. Other voltages are available. Buss-bar wires are run across the board just below the holes for the lamps which are slightly larger than 1/4-inch. By **double** heat-sinking the slide base of the lamps, it's possible to scrape and solder the slide base to the cross buss-bar wire, holding the lamps neatly in the holes from the rear.

Holes are provided for the lamp-driving transistor and the base resistor leads connected as shown in **fig. 9.** Surplus 5 x 7-inch sheets of thin phenolic material are adequate to mount and hold seven decades of lamps (28 lamps). In the absence of any shielding, some flip-flops are triggered by rf entering the lamp-driving leads when transmitting at full power.

The lamp-driving transistors should not load the FF's heavily. With FF's that have a plus power requirement and output, npn transistors are selected as lamp drivers. New 2N1302 transistors vary a great deal; at least half had too low beta (collector current divided by base current) and required a low value base resistor. This resistance was around 3900 ohms. Other builders have selected the 2N1304.

After considerable checking, the 2N497 was chosen here, rejecting only about one-third of them after buying a bag of clean surplus devices from Solid State Sales at less than 20¢ each. About half of these work fine with a 10k base resistor, and almost all work with 4.7k or 3.9k. For convenience, 10k resistors are mounted in the indicator lamp board with the driving transistors; another resistor is soldered in behind the board where necessary to saturate the driving transistor and equalize lamp brilliancy. The fourth lamp of the first two decades does not light during a count but does when the count stops, because of the 20 per cent duty cycle.

interconnections

A double-pole double-throw toggle switch turns on crystal-oven heater power alone, or both this and all other power. The lamp power supply is also fed through a contact in the count-selector switch so that the lamps are extinguished except during counts and reading. The best sequence for a power switch is: off; oven heater power; add crystal-oscillator

power; add IC power for calibrator and counter; and add indicator-lamp power; in that order. Of this, indicator-lamp power and possibly IC power for calibrator and counter can be in the count-selector switch.

The harmonic-selector switch has an off-position for no signal to the receiving antenna; and then a 5-kHz contact. The final position connects either 2.5 kHz or, while counting, the 1-kHz signal. The output from this switch goes to a blocking capacitor and a potentiometer as shown in **fig. 10.** This is desirable to adjust calibrator output to obtain the strongest beat note with an incoming signal. A potentiometer in the regulator divider varies the voltage on the crystal oscillator from about 9.5 to 12.5 volts as a "fine" adjustment of crystal frequency.

There must be a "little red button" for grounding the reset line so that after a fractional second delay, a one-shot one-second count can be obtained without turning the count-selector switch if it's already in the right position. Since it's the only operation needed to make a new count, a phono plug and cable can be used (such as RG-174/U) with the little red button in a small minibox. This will start a new count from any convenient position, even out of reach of the counter itself.

construction

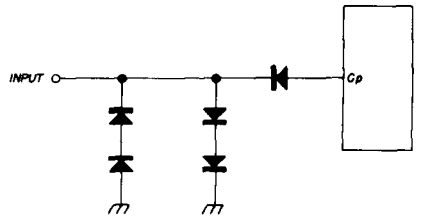
The chassis I used was overly large, since it was an available rack-mounted unit. It extends some inches behind the control panel which already has a 14-MHz front-end crystal filter,[26] audio switching for the loudspeaker and phone-patch, antenna switching and paralleling, and a small, easily-operated push-to-talk toggle for phone patching.

The chassis holds the surplus transformer and large electrolytic capacitors, the controls that are not brought out to the front panel, a Waber in-line grounded outlet box on the rear apron for 120 V to other equipment, and the count indicator, and leaves enough empty space for a good-sized cat.

Except for the heat-sunk transistors, the transformers and electrolytics, the three power supplies are mounted on Vectorboard with small holes on alternate 0.1-inch spacing inasmuch as there was to be some development work, and, later, the addition of the lamp supply. Quantities of Cambion solder terminals are mounted along two edges.

The oven-controlled crystal-oscillator plugs into an octal socket. It feeds the calibrator/time-base board which is

fig. 11. Protective diodes will only permit a negative input pulse to get into the first count flip flop. One pair of silicon diodes may not be required because of the series diode. If higher input voltages are desired, the number of series diodes may be increased.



made from phenolic material I had left over from "der Kleiner Keyer"[27]. It has a regular pattern of holes drilled to hold resistors and other parts along one edge, with holes here and there for transistors that may be needed in the future. The remainder of the board has sets of 14 holes for the plastic dual-in-line J-K flip-flop IC's. This board is 3½ x 6 inches.

It is stacked with the count board and mounts on bolts and spacers from the rear apron of the chassis. A part of the count board is reserved for the many terminals needed for the leads to the count-indicator which extend up from the chassis. The new Vector microboards with holes spaced 0.1 and 0.05 inches will accept dual-in-line IC's and transistors directly without any drilling, but they were not on hand at the time.

The 9093 units have a direct-set terminal which, when grounded, makes all FF's go to 1. To provide a reset function, all J's and K's were interchanged, and all

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signal/one

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Q and $\overline{Q}$ outputs were interchanged in the counter board and the gate FF's.

It is not easy to test the flip-flops prior to installation. One in each of several dual FF IC's turned out to be inoperative. Replacement was accomplished easily. Dc tests have contact-bounce problems although transistor keying or some of the proposals in reference 28 might help.

The best test method, in the absence of a low-frequency receiver to listen to the normal output frequencies of each FF when driven by the crystal oscillator, is to put an audio frequency in and observe the output at each FF and decade. This can be detected with headphones. The average current from the fourth FF in each decade is low because of the 20 percent duty cycle during a count.

The maximum frequency at which the FF will operate at can be determined by winding several turns of wire around the coil of a grid-dip oscillator and feeding it into the input—with a squaring circuit or trigger in between if required. The 9093 IC's that I used here reached the specified 2.2 MHz. I will probably add a faster decade ahead of the present counter so that the maximum frequency will be above 30 MHz to permit measuring all the oscillators in ssb gear. Then, by putting the counter on the vfo and making an addition (of any mixing crystals), fairly accurate frequency measurements will be available. However, for some purposes, the audio oscillator and oscilloscope will always be necessary.

In view of the direct application of inputs, there is some chance that overdrive might destroy the first IC. This doesn't happen with up to 6 volts from the audio oscillator, and counts are satisfactory down to about 0.06 volts. It is desirable to avoid a potentiometer input adjustment.

Just to be safe, two silicon diodes in series were placed across the input in one direction, and two more in the other direction, with a diode in series with the input to the IC to pass the negative swings of the input pulse. See **fig. 11.** Possibly the series diode makes one pair from the input line to ground unnecessary, but they haven't been removed yet. The diodes are inexpensive ones from Solid State Sales.

crystal adjustment

There is a little aging in the crystal oscillator until it operates for several weeks. The power-supply voltage adjustment is placed in midscale; the plunger-capacitor adjusting screw is then set for zero beat with a high harmonic of WWV; all subsequent adjustment is made with the potentiometer. Its position is noted for one cycle on each side of zero beat. At the midpoint of these two readings, only irregular fades in the 20-MHz WWV signal affect the receiver S-meter. With no evidence of a beat, it's helpful to be able to misadjust the power-supply potentiometer to determine that there is a beat note with WWV, and then to restore the pot to its previous setting.

There is another method of accurately adjusting the crystal oscillator to WWV without low-frequency transmissions.[29] It is described by Jay O'Brien, W6GDO. A 100-Hz output is fed from the calibrator/time-base board to a triggered external-sync oscilloscope input connection. When WWV transmits the 600-Hz tone for two out of every ten minutes, the scope sweep is set to show just one cycle from the receiver output. The crystal oscillator is adjusted so that the left-to-right movement of one cycle on the screen is less than one-tenth of a cycle during the two minutes. Of course, this must be done when the scope pattern is stable and does not shift left or right due to Doppler effect caused by propagation changes.

By using the first of these two systems to adjust the crystal oscillator, WWV has been measured on 20 MHz from the calibrator harmonic 2.5 kHz lower in frequency. This results in either a zero—or one—cycle error.

Better results probably would require going to a ten-second count period in order to avoid a possible phase error of one cycle in the count. The error does not occur when counting the crystal oscillator or any audio frequency fed into both boards—the result is always the same count as the nominal crystal oscillator output.

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In the ARRL frequency-measuring test, six measurements were submitted for the record. Five had a zero cycle difference from the official measurements made by the umpire. One had an error of two cycles. Propagation conditions can cause similar errors, especially when an elliptical pattern on the oscilloscope between the receiver output and the audio oscillator collapses repeatedly.

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ham radio

short circuit

Some errors crept into the excellent article on the kilowatt amplifier for 1296 MHz which appeared in the August issue. First of all, the width of part E in **fig. 2,** page 12, should be 1-5/16 inch, not 1-15/16 inch as shown. The cavity may still work this way, but the adjustment range of the input probe is insufficient.

Part J of **fig. 2** does not allow for the wall thickness of the cavity. It should be at least 1-1/8 inch wide initially and filed flat after soldering it to the top and bottom of the plate cavity. Its thickness depends upon the thickness of blade used to make the cuts in the cavity.

The perforation holes in part D of **fig. 2** should be 1/8 inch in diameter over an area of 1-1/2 inch with at least 50% open area. The toggle switch in the schematic labeled "driver off" should have two sections in series; a single toggle switch will not stand the voltage.

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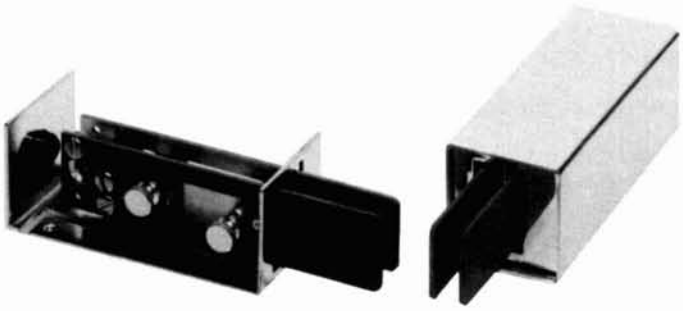

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more electronic units

More on the discussion of basic electronic units started last month

Jim Ashe, W1EZT, P. O. Box 343, Peterborough, New Hampshire 03458

What are some of the electronic units that trouble hams who have been working at radio for a while? I think there are four that could use some close attention; the cycle, the mho, the steradian, and the dBm. Each of these is a very useful idea, and any one of them can give you trouble way out of proportion to your real need for the term when you don't see what it's all about.

cycles

Cycle has a curiously indefinite, yet very precise meaning. Roughly, it means "through the sequence of events to where we came in, or to something like it." This tells us the term is applicable only to repetitive events, such as continuous movies, some electrical signals and some mechanical motions. Noise, while full of cycles, is not repetitive so we do not see a cycle of noise.

A definition in the 1968 Radio Amateur's Handbook is confusing, but will get by for some kinds of signals. However, it wouldn't do for the output of my thin-line generator, for example, which never goes negative. And modern communications and industrial electronic circuits can produce an infinite variety of repetitive electrical signals. We need a better definition: a cycle is the interval from any definite point in an electrical signal to the next point having the same amplitude, direction, and general shape. See examples in **fig. 1.**

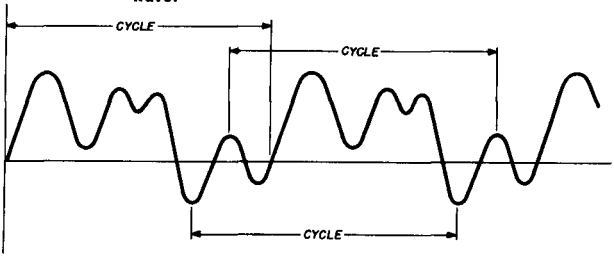
mhos and things

I hope you can stand an awful pun. I'm enjoying this. One of the most incomprehensible electronic units I've seen in amateur radio literature is the micromho.

Yet it is a very good term, once you know how to use it. If you are still working with vacuum tubes (and some of them **are** pretty good), micromhos are very handy thinking tools. But first we ought to turn them into something more sensible.

Reference to the books tells us a micromho is one-millionth of a mho, which is an amp per volt. So micromhos are microamps out per volt in, and the familiar old 6AK5 is rated at 5200 microamps anode signal current per signal volt into the grid. But why microamps? The first vacuum tubes must have been very weak. I prefer to think in milliamps and thousands-of-ohms anode load resistances.

fig. 1. This waveform emphasizes the "cycle" as a distance rather than a definite zero-to-zero interval along the wave.



When I look at that 5200 micromhos I can read it as 5.2 milliamps per volt times kilohms plate load resistance, for a rough output figure.

For instance, imagine you have provided a 6EJ7 pentode with a 15k-ohm plate load resistor. For a schematic you can breadboard, see **fig. 2.** Now, the 6EJ7 is rated at 15,000 micromhos, or 15 milliamps per volt. Assuming a 1-volt grid signal, 15-mA signal current times 15k anode resistance gives 225 volts signal out. You can do that in your head. In the real circuit there won't be enough anode voltage swing available for this large output, but we look for a voltage gain in the times-225 ballpark.

steradians

Sounds kind of hairy, doesn't it? If you're a vhf enthusiast you are sure to be concerned with steradians. What is a steradian? It is a way of describing how much sky your antenna illuminates.

Imagine a four-sided pyramid extending into the sky from your high-gain antenna as in **fig. 3.** How big is the end of that pyramid? Hard to say, not knowing how far away the end is.

But suppose we choose any convenient distance, call it R, and measure the lobe cross section area of the end in square R's. We'll get the same number whether R is measured in inches, feet or miles, and at any distance. Well, that pattern in space, measured in terms of the area in square R's at the end of it, is said to have a certain size or solid angle given in steradians.

A steradian is an area of one square R, whether it is a circle out there, an ellipse, or any other shape. Thinking is similar in style but more complicated in detail if the beam is multilobed in cross section. Next time you bump into 'steradians,' say 'square radiuses' and the shock will be diminished nearly to zero.

dbm

Here is a term that used to seem to me about like a magician's wand: wonderful, incomprehensible. But it's really a basic tool for you, if you're interested in long-distance radio communications. The dBm is not as hard as it sounds: as with other electronic terms, people started using dBm's in their thinking because it made things simpler. Once you see why, you have the whole thing. The dBm idea contains two other ideas tied together, which is the

table 1. Here are the power ratios for some common decibel values. The same figures, with appropriate signs, serve for power gain or loss.

power ratio	dB measure
100:1	20
50:1	17
20:1	13
10:1	10
5:1	7
2:1	3
1:1	0
1/2:1	−3
1/5:1	−7
1/10:1	−10
1/20:1	−13

cause of the comprehension problem.

The first idea is decibel measurement. What is a decibel? A decibel is a size of difference, rather than a size of a quantity or of a thing. A decibel change in signal strength is about the least your ear can detect by listening closely, regardless of the actual signal strength at the time. A 3-dB change will be quite perceptible, but if you are thinking about something else it could escape your attention.

A 10-dB change in signal strength will almost certainly be noticed whether you are listening or not. A 1-dB change is about 25% increase or decrease in power; a 3-dB change is a doubling or a halving of power, and a 10-dB change is by a factor of ten. See **table 1.**

You ought to try this on the bench. It's easy to do and will do more for your perspective than any amount of math or reading. Maybe it will change your thinking about spending more money to double your rig power too. Set up a signal generator, an audio amplifier and a loudspeaker as shown in **fig. 4.** Try changing audio levels in decibels and fractions of a decibel. It's surprising, even when you know what to expect and why you get this result. It is because of a built-in agc system in your ears, so that a "ten-times-as-much power" signal is judged to be, maybe, "twice-as-loud."

Now, all electronics signal devices respond to power input, not really to voltage input. It is an old custom, fortunately falling into disuse, to say a receiver responds to so-many microvolts input. The receiver actually accepts a certain definite signal **power** input, amplifies and detects it and amplifies it some more, and puts out a definite larger signal **power** output. How can we describe this performance?

We simply choose a unit of power and specify how much in terms of that unit must be fed to the input in order to get a satisfactory output. One commonly used reference power is the milliwatt applied to 50 ohms. Let's go ahead with a realizable bench test setup and see how it works. If we wanted to test a receiver we would set up things as in **fig. 5.** Suppose we haven't had too much experience with high-gain circuits?

A milliwatt signal is easy to make but hard to measure, so we start with a larger signal and pare it down to milliwatt size after verifying its power. Now, since the receiver surely has lots of gain, maybe even 1000 (30dB) or more, we apply our milliwatt to the receiver antenna terminals (variable attenuator switches all at zero dB) and get blasted as the receiver

fig. 3. To describe the space filled by an antenna lobe, use radius as the unit of distance and specify the amount of sky filled in terms of square radiuses.

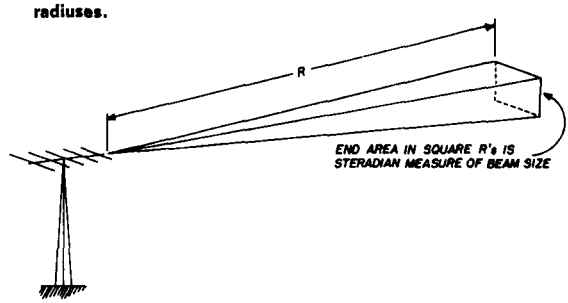
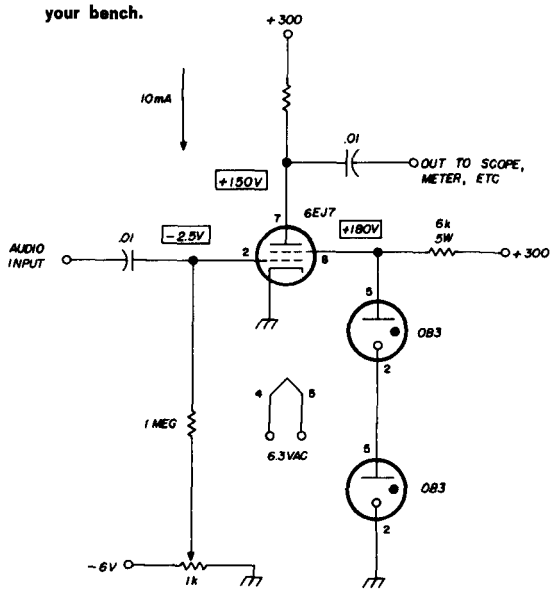


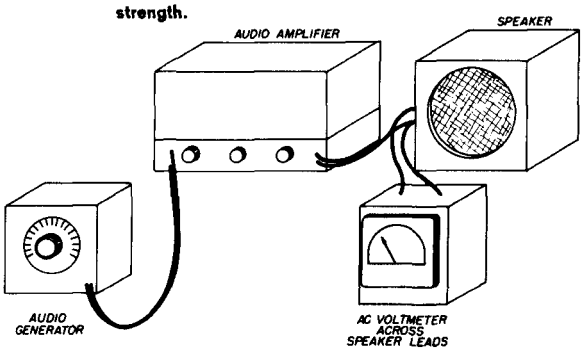
fig. 2. You can use micromhos to quickly estimate stage gain of a pentode amplifier; try this circuit on your bench.



overloads badly. But let's assume for a moment that the receiver did put out the watt specified in **fig. 5.** How would we describe this?

We'd say the input was zero dB different from 1 milliwatt, or zero dBm (into 50 ohms). The receiver would be showing a power gain from input to output of 1 milliwatt rf to 1 watt of audio, or 30-dB gain. This emphasizes that receiver sensitivity and receiver gain are different aspects of receiver performance. And it shows us that we cannot specify receiver gain **or** sensitivity until we have chosen an output power to use in the input-output gain figure.

fig. 4. Simple setup for demonstrating apparent loudness variation with dB change in signal strength.



Typically, receiver sensitivity is the smallest signal that will produce a satisfactory output, but what if you have the gain control turned down? The receiver has a sensitivity that is smaller than its best sensitivity, but the same thinking applies.

Now we continue our experiment. At one microwatt input (this is minus 30 dBm, now) we discover we are still overloading the receiver. One millionth of a watt! No wonder some circuits are very sensitive to noise and feedback. Input-output gain would be 60 dB if the receiver were working properly, but a weaker signal is still required.

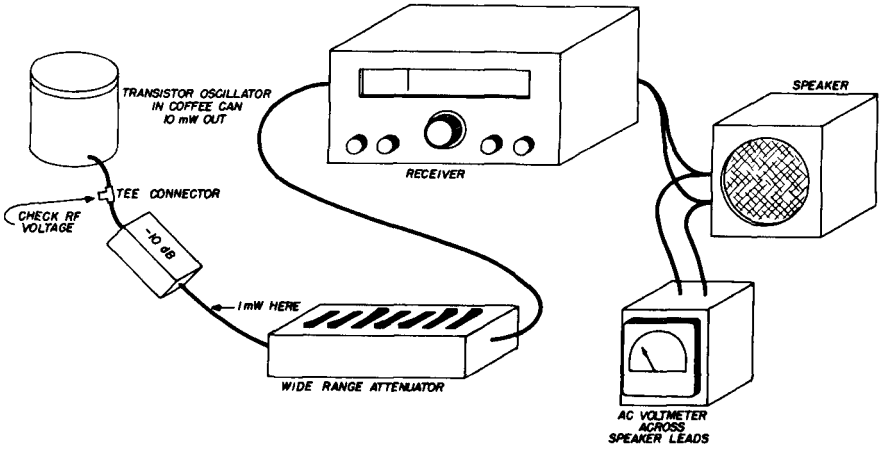
Finally we switch in enough attenuation for a normal 1-watt receiver output. Our generator is still feeding its milliwatt into the variable attenuator which now reads 110-dB attenuation. These are realistic figures for a rather average receiver. At last we know our receiver has a sensitivity, at certain control settings, of minus 110 dBm, and its over-all gain is 140 dB.

A lab visitor says, "Please translate that into microvolts." Converting from dBm to watts we have 10^{-14} watts input, and since $E^2 = WR$, we get about 0.7 microvolts across 50 ohms receiver input resistance.

I expect a number of you will see how well this procedure settles questions about receiver performance, and will want to start doing tests like this. Let me pass on a few thoughts.

Start with a signal large enough to mea-

fig. 5. Test setup for evaluating receiver performance



sure. A milliwatt signal is about 223 millivolts across 50 ohms: pretty small. A ten milliwatt signal would be just over three times greater in voltage. Calibrate your diode probe with audio, remembering to use capacitors whose time constants are not too small.

Base your attenuator design on Daughters' and Alexander's excellent article on attenuators.[2] This article is one of the best I have seen, because it is written from the viewpoint of a ham who is working with such things.

Set up for the test to minimize rf leakage from the oscillator-attenuator assembly **and** to minimize receiver sensitivity to rf from anywhere but the input terminals. You probably can't get perfection in either department, but a good job in both will produce the same practical results.

summary

That's enough on electronics units. The rest is up to you. When an electronic unit seems hard for you to use, spend some time reading about it. If you look for the way in which the term is built up from simpler terms or from common-sense ideas, it will gradually become more real to you.

Then try to put it to work in some simple breadboard circuit that tests or illustrates your thinking about the unit. Finding work for a new idea is the best way to make it stick.

references

1. J. Ashe, W2DXH, "100-kHz Thin-Line Pulse Generator," 73, February, 1968, p. 24.
2. G. Daughters, WB6AIG and W. Alexander, WA6RDZ, "Low-Power Attenuators for the Amateur Bands," 73, January, 1967, p. 40.

ham radio

unmarked filter chokes

Have you ever picked an old power choke out of your junk box and wondered how you could find out its inductance? You probably checked the dc resistance, scratched your head, and put it back in the junkbox as another of life's insoluble problems.

Well, there is an easy and painless way to figure out the inductance of any power choke. The only equipment you need is a filament transformer, a vtvm with reasonable ac accuracy and a composition resistor of known resistance (± 5% is good enough). There is only one limitation—you obtain the zero-dc-current inductance. If you're looking for the inductance of an audio filter, this is fine. For power filtering purposes with dc flowing in the choke, the inductance may be perhaps 30 or 40 percent less, depending on the characteristics of the core and the amount of current.

Let's look at a thumbnail sketch of the process first and then run through a sample calculation. The process is based on Ohm's law for ac circuits where impedance (Z) is substituted for resistance (R). The formulas are written the same way as the dc version of Ohm's law (e.g., $E = IZ$, $I = E/Z$, $Z = E/I$).

One further point: if the voltage is measured as rms, the current will also be rms. The same applies to peak-to-peak measurements. As long as the system of measurement is consistent, only the values are important. The second formula you need is a variation on $Z = \sqrt{X_L^2 + R^2}$. If you know Z—more on this in a moment—and R, the choke resistance, you can find the inductive reactance of the choke from $X_L = \sqrt{Z^2 - R^2}$. By now you may see where we're heading.

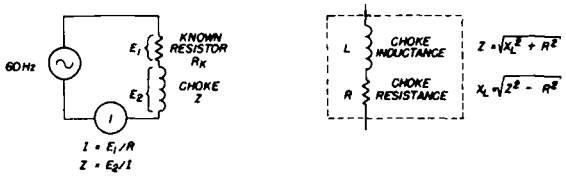


fig. 1. Here's the circuit used for determining choke inductance. The principles are explained in the text.

The first step in the process is to measure the resistance of the unknown choke. Your ohmmeter is accurate enough for this. Now connect the choke in series with a resistor of known value. Apply 5 to 10 Vac from a filament transformer to the ends of this RL network. Use the vtvm on its ac voltage ranges to measure the ac voltage across the known resistor.

With this information, you can calculate the ac current flowing in the circuit. Now measure the ac voltage across the choke. When you know the ac voltage across the choke and the ac current flowing through it, the impedance of the choke at 60 hertz may be calculated from $Z = EI$.

You already know the dc resistance of the choke, so by using Z and the formula $X_L = \sqrt{Z^2 - R^2}$, you solve for inductive reactance. As mentioned earlier, this will be the reactance of the no-dc-current inductance. While you're at it, you can also determine the Q of the choke from $Q = X_L/R$. This may be important if you are building an audio filter or other unit requiring a resonant audio circuit.

example

Assume a choke with 320 ohms dc resistance. You measure 4.7 Vac across a 1000-ohm resistor; this gives an ac current of .0047 amperes. You measure 7.5 Vac across the choke. Dividing 7.5 by .0047, you find the choke impedance at 60 hertz is 1600 ohms. From $X_L = \sqrt{Z^2 - R^2}$, where R is the choke resistance, you find X_L is 1570 ohms. To find the inductance, divide 1570 by 377 ($2\pi f$, where f is 60 Hz; the inductance is 4.17 henrys.

Now that you see how it's done, there shouldn't be a single unknown choke in your junkbox.

table 1. summary of process

1. *Measure the ac voltage across the series resistor.*
2. Compute I from E/R1 where R1 is the series resistor.
3. Measure the ac voltage across the inductor.
4. Compute Z from E/I.
5. Compute X_L from $\sqrt{Z^2 - R2^2}$ where R2 is the *internal resistance of the choke.*
6. Compute L from $Z/2\pi f$ where $2\pi f = 377$ for 60 hertz.

Bob Tellefsen, WØMKF

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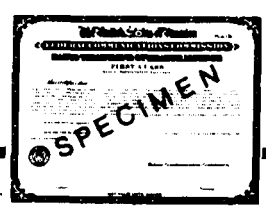

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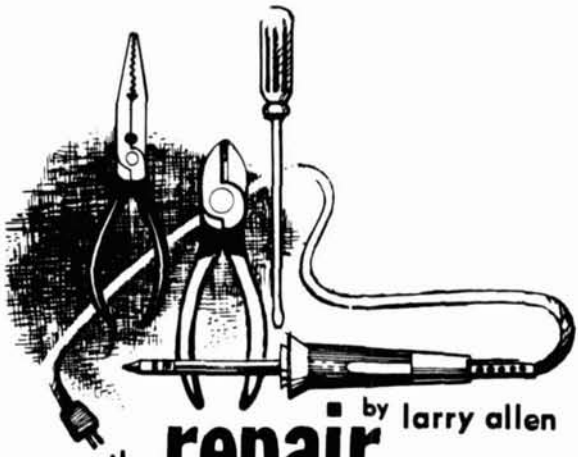


by larry allen

the repair bench

trouble shooting by resistance measurement

In a recent **repair bench** column* I said you should only work on transmitters with power off. I suggested checking resistances to find the trouble. Since then, readers without repair experience have asked how to go at this kind of trouble shooting. No one has argued the advice, but many don't see the approach.

The best excuse for using resistance measurement as a fault-hunting method is safety. You can check out most of a transmitter without any lethal voltages applied. Sure, for a few tests the transmitter or receiver has to be "live." But so many tests can be handled more safely with power off, there's just no excuse to ignore these advantages.

A few doubters may contend that the method isn't complete. Certain parts in high-voltage circuits break down under stress, yet test okay by resistance measurements, they say. That's true. But you soon learn to spot those troubles right away—usually by their smoke. Also, certain shorts and arcs in transmitters occur only after high voltage and drive are applied. The way to track them—safely—is by a process of elimination, part by part . . . but that's a story I'll save for a later column.

Resistance trouble shooting can be done with a simple volt-ohm-milliameter, or vom. That's a major advantage; other techniques demand fancier instruments. A vom is simple to operate, portable, and usually inexpensive. The main criterion is that the ohmmeter readings be accurate.

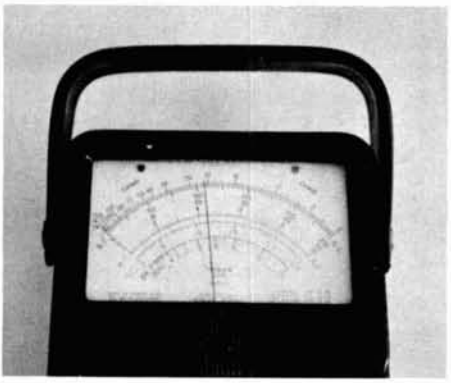
A vom has an advantage over a vacuum-tube voltmeter, even when the vtvm is more accurate. Since a vtvm must be plugged into the power line, it

Adjusting the mechanical-zeroing screw.

* *repair bench*, August 1968, page 52.

can introduce problems in certain equipment. With the vtvm common lead connected to some non-ground point in a transmitter or receiver, you may get some faulty readings. A vom is usually battery-powered; there's no power-line connection. You can get better readings without worrying about "unseen" ground paths caused by a test connection.

The needle should rest somewhere near midscale for best accuracy during tests.



The needle should move up scale when the test leads are clipped together.

reading the ohms scales

Dependable ohmmeter measurements are useless if you don't read them accurately. You can't even go on to the next step—interpreting their meaning—unless you get the numbers right. Three aids will help you read your ohmmeter correctly.

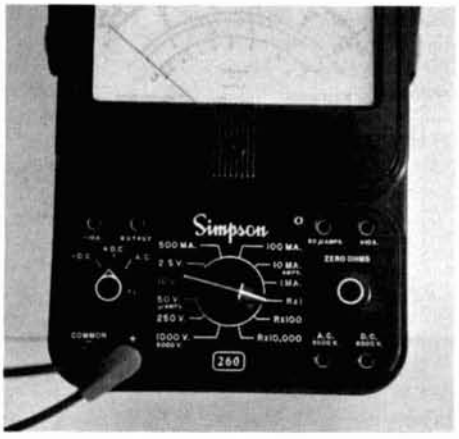
First of all, **calibrate** the instrument. Place the vom upright or lying down, in whatever position you'll use it. (Position will affect vom readings.) With the test leads plugged in but not shorted together, make sure the meter pointer rests exactly over the left edge of the scale. You may have to adjust the mechanical-zero screw that is just below the meter face. Tap the meter glass to jar loose any pointer friction.

Next, clip the two test leads together, making zero ohms between the tips. The pointer should move up to full scale. Zero ohms is at full scale since the ohms-scale numbers go right-to-left. Again, tap the meter glass to be sure the pointer rests normally. That takes care of calibration.

A second aid to accurate resistance measurements concerns the range switch and the multipliers you use with the scale readings. After you connect the test leads across the resistance you're measuring, rotate the range switch to a position that makes the pointer rest somewhere in the middle section of the scale. That's where readings are the most accurate because the numbers there are easy to read. You can see that they are, if you examine the ohms scale in the photos; it's the top row of numbers.

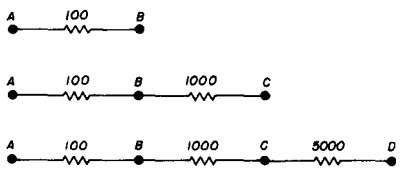
For a third aid to accurate readings,

Pay close attention to the multipliers on the range switch.



make sure you multiply the scale reading properly. Look at the multipliers on the range switch. Pay close attention to which one you set the switch at. Then multiply whatever number the pointer stops at by the multiplier from the switch. Do it carefully. It's awfully easy to drop

fig. 1. When resistances are in series their values add directly even if they're not arranged in a straight line as the ones shown here.



a zero or add one and end up with a wrong reading. Use scratch paper to be sure. Whenever the pointer stops toward the left end of the scale, you'll have zeros in the scale reading as well as in the multiplier; be sure you get those right, too.

simple series paths

What do you do with those accurate resistance readings once you have them? Analyze them, of course! But that brings up the next problem: How?

What you do first is learn to recognize the different kinds of resistive paths. There are plenty of them. Every ham receiver and transmitter has dozens of resistors. Besides them, many, many other components show readings on an ohmmeter. For example, the heaters of tubes; the windings of transformers; the leakage of electrolytic filter capacitors; the forward and backward effects of diodes; and, of course, transistors, which produce some readings that can really fool you. Resistive paths exist all around the chassis.

The trick of resistance trouble shooting is finding paths that aren't where they should be or paths that have too much or too little resistance. To do it, you'll have to know how to spot the paths on diagrams and figure out where and what they should be in the chassis.

The simplest, both to spot and to analyze, are series paths. **Fig. 1** shows a bunch of different kinds. The A-B path shown first can't very well be mistaken; it's 100 ohms, between points A and B.

In the second series path shown in **fig. 1,** the resistance between A and B again is 100 ohms. In series with the A-B path, however, is a path from B to C. Again, the over-all circuit is easy to figure out; the A-C path is the sum of the A-B path plus the B-C path. Resistance in the A-C path totals 1100 ohms.

fig. 2. Here are several series resistance paths whose values add directly like those in fig. 1 do.

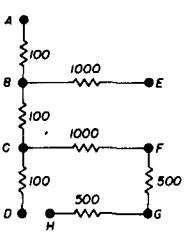
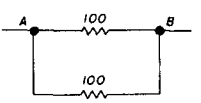
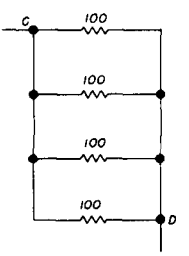
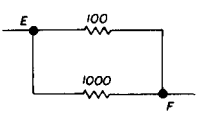


fig. 3. Parallel paths are more difficult to calculate; use the formula in the text to find path resistance.







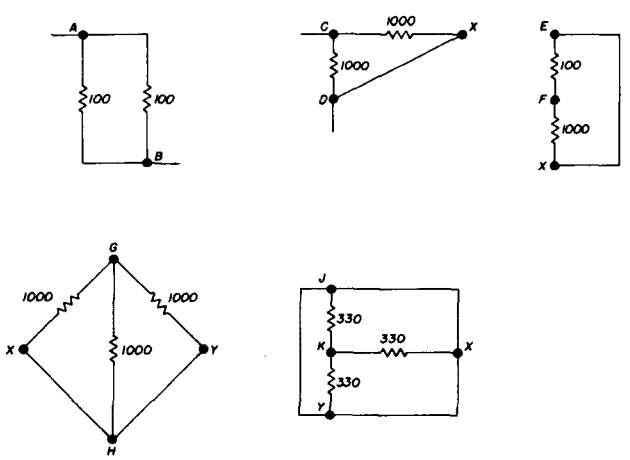
The third series path is A-D. This one, too, is pretty easy, even if it does have three series resistances. Add all three of them, and you have a total A-D resistance of 6100 ohms.

It's important that you recognize another relationship from **fig. 1.** Connecting the B-C and C-D paths had no effect on the A-B path; in all three cases, an ohmmeter connected to A and B measures 100 ohms. You can add a dozen more paths in all directions, but A-B remains 100 ohms—as long as all the paths are added **in series.**

Another thing. You can measure path B-C independently, and path C-D independently. The other series paths don't interfere with measuring either one. Also, you can measure path B-D, if you want to, getting a direct 6000-ohm reading between points B and D.

Take a look at **fig. 2.** Paths don't have to be drawn in a straight line to be in series. Nor do they have to be connected directly to each other. An ohmmeter connected to A and D measures the A-D path directly; the resistance is 300 ohms. None of the other resistances is of any consequence to that particular path, because they are not in series with it.

fig. 4. Different ways branches may be arranged and connected in parallel; don't let them fool you.



What about path A-H? Only the resistance **between** A and H have a bearing on the resistance of that path. An ohmmeter connected to A and H measures only the resistances between them. The resistances of path B-E and path C-D are ignored because they're not in series with path A-H. Path A-H measures 2200 ohms. Trace its path—from A to B, to C, to F, to G, to H.

Trace the path from E to G. It goes from E to B, to C, to F, to G. An ohmmeter connected between E and G measures only that path, and indicates 2600 ohms.

The D-H path is 2100 ohms, the B-F path is 1100 ohms, and the F-A path is 1200 ohms. Because these resistances are in series, they're easy to measure and easy to analyze.

Suppose you're trouble shooting a circuit like **fig. 2** and measure 2700 ohms between B and G. The schematic shows series resistances totaling only 1600 ohms in the B-G path. At least one resistance has changed value, and you have to figure out which one.

One way is to measure each resistance individually. There may be several in an actual circuit, and that can take a lot of time. A better way is to work your way toward one end or the other. Leave one ohmmeter lead on B and move the other one from G to F. If the reading drops to 1100 ohms, that means the B-F path is okay, so the F-G path must be at fault. Suppose, though, the reading only drops to 2200 ohms. That means F-G is okay; the 500-ohm change tells you so. The trouble is therefore in the B-F path.

Move the test lead from F to C. If the reading drops to 100 ohms, B-C is okay. You can then move both leads to measure path C-F, just to be sure. If it measures 2100 ohms, you've found the trouble; it should measure only 1000 ohms.

Those are the principles. As long as paths are in series, tracking down trouble is a cinch. Plenty of resistive (and continuity) paths in ham equipment are series types. With them, you'll have an easy time. However, the other kind can be a real pain.

resistive paths in parallel

The simplest kind of parallel path is shown in **fig. 3.** Parallel paths are hard to describe. As far as an ohmmeter is

concerned, the path between points A and B is merely path A-B. As you can see from the diagram, however, it isn't that simple. There are two paths. One is from A to B . . . and, yet, so is the other! You'll find several paths like this in electronic equipment, and very often they aren't at all obvious on the schematic diagram.

Resistances in parallel add inversely (another word for upside-down). Put two equal-value resistors in parallel and the resistance of their path is only half the value of either one alone. Take path A-B in **fig. 3,** for instance. The two 100-ohm ressistors in parallel make a path measuring only 50 ohms; that's what an ohmmeter reads. The C-D path, with four 100-ohm resistances in parallel, measures only 25 ohms with an ohmmeter.

From this you can figure that adding a parallel path always lowers the ohmmeter reading between two points. Parallel path resistance is, in fact, less than the lowest value of resistance in any one branch of the parallel path. The E-F path in **fig. 3** shows a situation where one branch has a much lower resistance than the other. Calculated by the formula for parallel resistances,* path E-F has a resistance of 91 ohms. That's what an ohmmeter measures between points E and F.

Consider what it means, then, if you measure path E-F and find a reading of 100 ohms instead of 91. You can assume the 1000-ohm resistance has been removed, is defective (open), or has changed value to some extremely high resistance. On the other hand, if the reading is 1000 ohms, the 100-ohm resistance is at fault. Get the idea?

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \cdots \frac{1}{R_n}$$

odd parallel connections

The trouble with parallel paths in typical ham gear is that they are seldom as obvious as those in **fig. 3.** For example, they could be arranged in all sorts of odd shapes, like some of those in **fig. 4.** Strange as they may appear, these diagrams are all of parallel resistance paths.

A-B is easy to recognize; it's like the one from **fig. 3,** only turned up on its side. Path C-D looks at first glance like a 1000-ohm path. Follow path C-X, though, and you'll find that X and D are the same point insofar as the circuit is concerned. The C-D path is therefore a parallel one involving both resistances. The true resistance is 500 ohms between C and D; that's what an ohmmeter reads.

Consider the E-F-X path. The two resistors appear in series; but they're not, because the jumper connection makes E and X electrically the same point. So, any current from E flows through both resistances to reach F. The true path is E-F (or F-E if you prefer). An ohmmeter at E and F measures the combined resistance of the two—combined in parallel, not in series. The path from E to F measures 91 ohms.

Path G-H is fairly easy to see, now that you're looking. The path is through three parallel resistances. Points X and Y are mere connections, and are the same (elec-

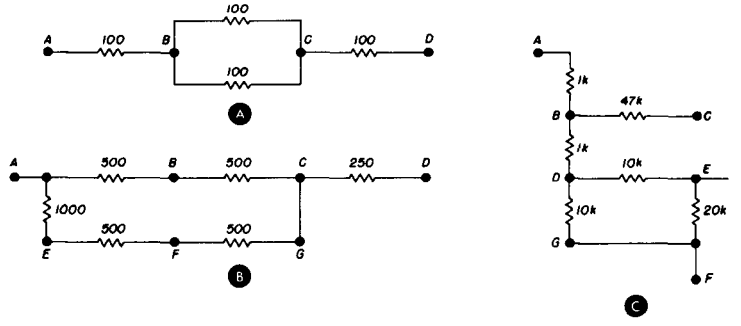
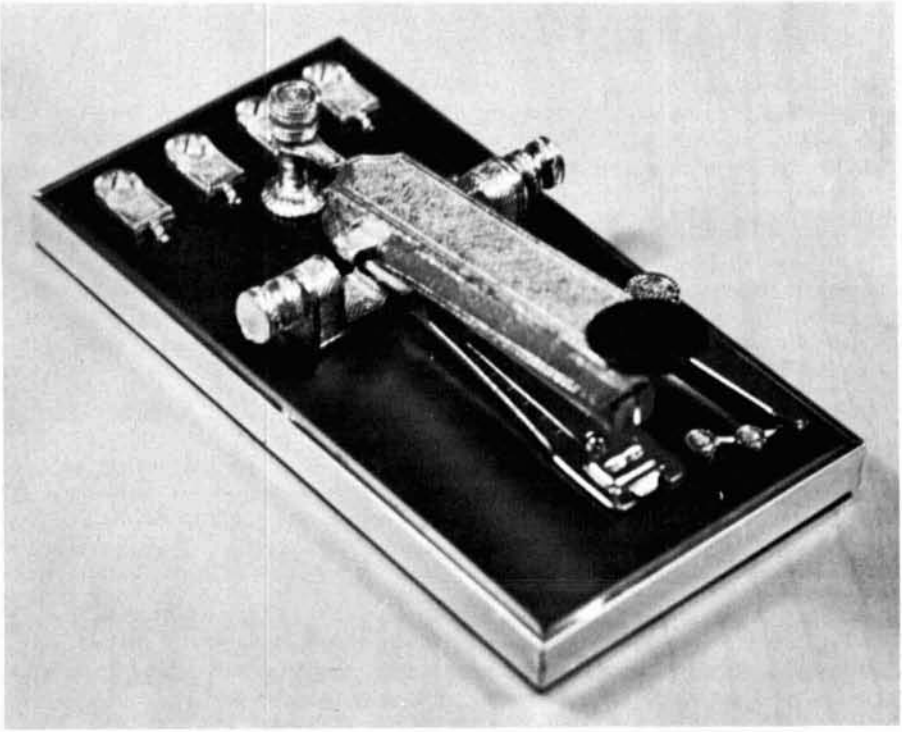


fig. 5. Series-parallel resistance paths are more like what you'll find in real equipment. Still . . . the method of figuring out proper path resistance is the same: calculate parallel paths and add series ones.



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trically) as point H. Combining the three resistances by the parallel-resistance formula, you can figure out the net resistance of path G-H: 333 ohms. An ohmmeter shows it to be around 330 ohms.

Whether it looks that way or not, path J-K is a parallel path through three branches. J, X, and Y are tied together by several jumpers, and can all be considered point J. The net J-K resistance is 110 ohms.

The point of studying odd resistance hookups like those in **fig. 4** is that you won't overlook a parallel connection just because it isn't obvious. In fact, many you'll encounter in actual equipment are even less obvious. Some are through components not drawn with resistor or wire symbols. In every case, keep in mind that a **parallel** path, whether visible on the schematic or not, always **lowers** the ohmmeter reading to some value **less than** the lowest value in any branch of the parallel path. If you forget that fact, resistance trouble shooting can be confusing; remember it, and you can use the technique to pin down some very elusive faults.

series-parallel combinations

A combination of paths is what you'll find most in equipment circuits. Some parts form series resistance paths, and some form parallel ones. You'll find them in all kinds of arrangements. A few possibilities are sketched in **fig. 5,** just to give you an idea.

In diagram **A,** paths A-B and C-D are in series with two-branch parallel path B-C. To know what resistance to expect between A and D, you first calculate the resistance of path B-C; that's 50 ohms. With that settled, it's easy to figure the rest: all paths are in series. A-B is 100 ohms, B-C is 50 ohms, and C-D is 100 ohms; the total is 250 ohms.

Fig. 5B is a little different. The paths are complicated. The path from A to C is made of two parallel branches, both of which have several series resistances. To find out what each branch should measure, solve one at a time. Path A-G totals 2000 ohms. Path A-C totals 1000 ohms. Points G and C are jumpered together, so are the same point. Combining the resistances of both parallel branches, you'll find the resistance of path A-C should be 667 ohms; an ohmmeter connected to A and C measures 660 or 670 ohms. The whole path, of course, extends from A to D. Combining path A-C with path C-D is simple addition, since they're in series. Path A-D should measure slightly over 900 ohms.

Fig. 5C shows what you might find in a real circuit. You can figure out what your ohmmeter should measure from any convenient point to any other point. Suppose F (and G) is ground; that's usually a good place to clip one test lead. Path F-E should measure about 10,000 ohms (10k), because the G-D-E path is in parallel with it. Path F-D should measure 7500 (30k in parallel with 10k). The path from F to B includes F-D in series with B-D and should measure 8500 ohms. The F-C path includes F-B as well as B-C and thus totals 55.5k (47k plus 8.5k). From F to A totals 9500 ohms or 9.5k; path B-C doesn't enter into the F-A path at all.

next month

You can see from all these diagrams that trouble shooting by resistance measurement is based entirely on figuring out from the schematic diagram just what a resistance path should be and then measuring it with the ohmmeter. If the resistance isn't what you expect, you isolate the one that's wrong. There, you'll find the trouble.

Principles alone aren't enough to make you feel at home with the resistance-measurement technique. In the next column, I'll show you some real trouble-chasing with this method. I'll show you how to track down parallel-series resistive paths, and how to use resistance tests to pinpoint specific parts troubles. By the time you're through with this subject next month, you'll know how to find faults in any kind of ham equipment with no power applied at all—the safest way you can troubleshoot.

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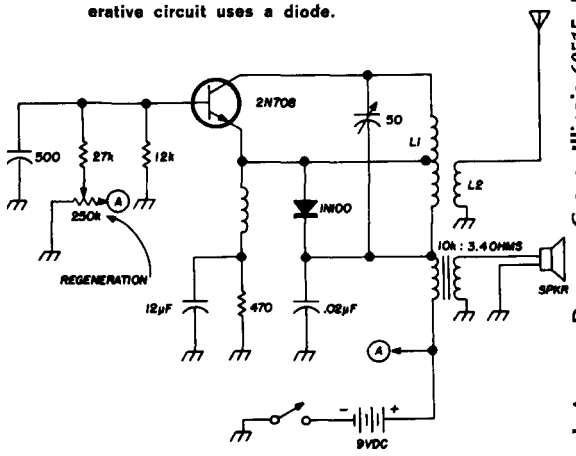
no-radiation no-hangover 28-mhz superregen receiver

Robert M. Brown, K2ZSQ, 5611 Middaugh Avenue, Downers Grove, Illinois 60515

Any amateur familiar with superregenerative receiver design is pretty well aware of their inherent disadvantages. Experienced hams tend to bypass this design altogether because of previous problems. Others simply ignore superregens because they tend to emit spurious radiations on other frequencies.

True, the circuit does have disadvantages—rather non-selective tuning, hangover and radiation—but it still represents a simple approach if you're looking for extremely good sensitivity at minimum expense. A look at the Heath Company's contributions on 28, 50 and 144 MHz in years past can attest to the sensitivity aspect. Now to cure some of the less desirable features.

fig. 1. New type superregenerative circuit uses a diode.



diode circuit eliminates radiation

After years of tinkering with superregenerative receivers, Bell Laboratories recently announced a major breakthrough: the addition of a diode shown in **fig. 1.** This diode eliminates the problems with hangover and radiation. While the selectivity factor remains, it has been shown that this simple addition promises better performance than any previous designs.

Residual oscillations persisting in the tank circuit of a superregen after active oscillation ends produce the well-known "hangover" effect—resulting in spurious responses and blocking action. Blocking inhibits detection because the receiver is swamped by its own residual signal.

By simply shunting a germanium diode across the tank coil, undesired energy is immediately dissipated after the oscillation burst. This eliminates hangover effects during the remaining sensitivity period of the decay slope. Note that the diode is installed so it doesn't affect the necessary positive feedback.

damping action

The problem of overcoming superregen radiation is also achieved by this diode-insertion technique. The diode damping action lowers the amplitude and significantly shortens the duration of the radiated pulse.

What happens to sensitivity? Since barrier potential—about 0.2 volt with germanium diodes—prevents conduction, tank loading with respect to incoming signals is also eliminated. Therefore, there is no change in sensitivity.

The circuit shown in **fig. 1** affords a practical opportunity to experiment with this technique. Any superregenerative receiver can be used, or you can start from scratch. The receiver shown here covers the 28-MHz band and is designed to match 50-ohm lines. The regeneration control should be adjusted just past the point a "hiss" is detected in the speaker output.

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propagation

predictions for november

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

As I write this column in August, some uncertainty exists about the trend of the present sunspot cycle. If, as predicted by the Zurich Observatory, the present cycle reached its peak this summer, then ionospheric conditions this November will very likely be similar to those experienced during November 1967. On the other hand, there is a 10% probability that the peak may yet occur during 1969, and that the muf's may be significantly higher this fall than last.

However, the smoothed sunspot number may not be the best predictor of quiet ionospheric conditions. The sunspot number most commonly used is the Wolf number, R, given by

$$R = k\ (10G + S)$$

where G is the number of sunspot groups, S is the number of observable individual spots, and k is a correction factor to account for different observers' characteristics. Daily sunspot numbers, obtained from the Swiss Federal Observatory at Zurich, are averaged over each month to obtain a monthly average sunspot number. The daily sunspot numbers vary widely because of the non-uniform distribution of sunspots across the sun's disk and its rotation (at an average 27-day period).

The monthly sunspot numbers are then smoothed by forming a running average over 12 months to form the sunspot numbers used in ionospheric prediction studies. Since the smoothed sunspot number is dependent on future observations, it can only be determined six months after the fact.

Predicted sunspot numbers are formed by extrapolating observed smoothed sunspot numbers and comparisons with past sunspot cycles. The prediction process is not particularly accurate. The National Bureau of Standards underestimated the peak of the last cycle by 50, and the 90% confidence level for last month's predicted smoothed sunspot number was ±29.

What do sunspots have to do with the ionosphere and why should present ionospheric conditions depend on anything that is going to happen in the future? There is no direct causal relation between sunspots and the solar radiations responsible for the ionosphere—both are different manifestations of solar activity. However, during the last two sunspot cycles a case has been built for statistical relations between the two.

A correlation between solar activity and ionospheric behavior was found by the National Bureau of Standards as early as 1936. In the same year, the first automatic full-range (500 kHz to 16 MHz) ionospheric sounder was used by the Carnegie Institution of Washington.

By the outbreak of World War II, ionospheric observations from only seven locations in the world were available to Allied powers; by the end of the war, 43 stations had been added, and ionospheric data was being published monthly.[1] Thus world-wide synoptic studies of ionospheric behavior have only taken place since about 1943.

By 1946, enough data had been gathered from soundings made at Washington, D.C., Huancayo, Peru, and Watheroo, West Australia to indicate linear statistical relationships between the ionospheric-layer critical frequencies and the smoothed sunspot number.[2] In general, the correlation was not as good if monthly average sunspot numbers were used.

By 1961, data collected over cycle 19 indicated a departure from the linear relationships between critical frequencies and smoothed sunspot numbers for sunspot numbers in excess of 150.[3] Some differences (less than 1 MHz of critical frequency for a given smoothed sunspot number) are evident between cycles 18 and 19, and between the ascending and descending parts of the cycle.

Solar cycle 17 peaked at a sunspot number of 113 in 1937; solar cycle 18 peaked at a sunspot number of 150 in 1947 and solar cycle 19 peaked at a sunspot number of 201 in 1957. Solar cycle 20 may have peaked near 101 this year. The sunspot numbers during these past four cycles were greater than during any other cycles since 1870. The point is, past correlation of world-wide ionospheric behavior with solar activity has taken place only during the past two solar cycles, which were of unusually high activity. The same correlation may not necessarily hold for future cycles.

propagation during the month

No matter what the actual smoothed sunspot number is for November, wintertime conditions will soon be here. Twenty meters will probably be closed most evenings before midnight, and fifteen meters will probably be closed within a couple of

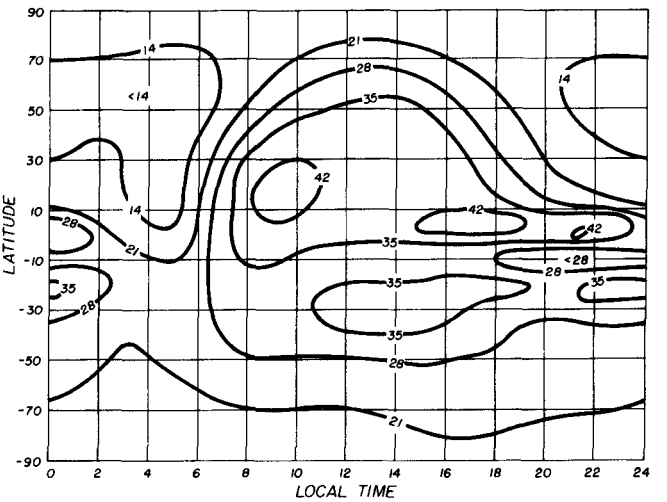


fig. 1. Maximum usable frequency curves for November 1968 based on 90° W longitude.

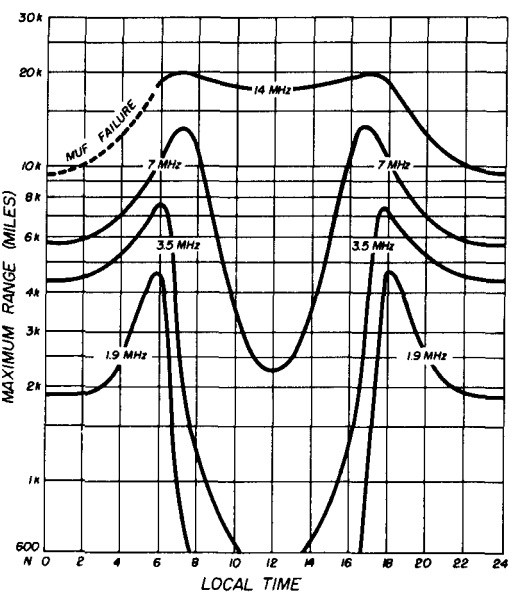


fig. 2. Maximum range to the north due to absorption.

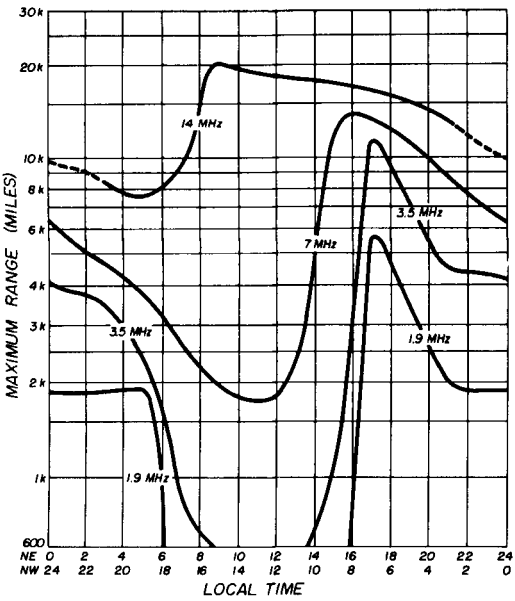


fig. 3. Maximum range to the northeast (top time scale) and to the northwest (lower time scale) due to absorption.

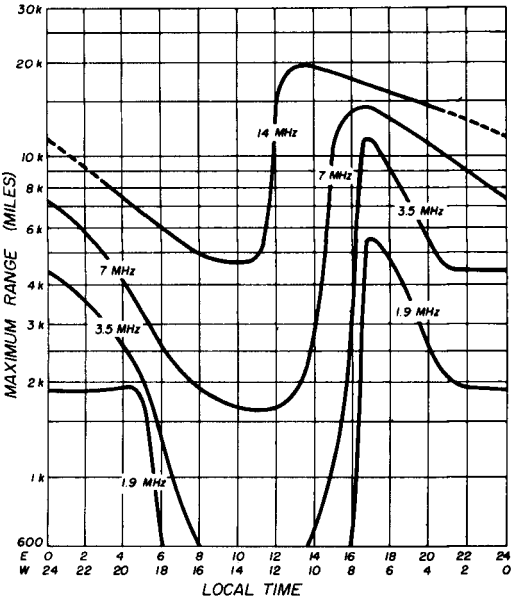


fig. 4. Maximum range to the east (top time scale) and to the west (lower time scale) due to absorption.

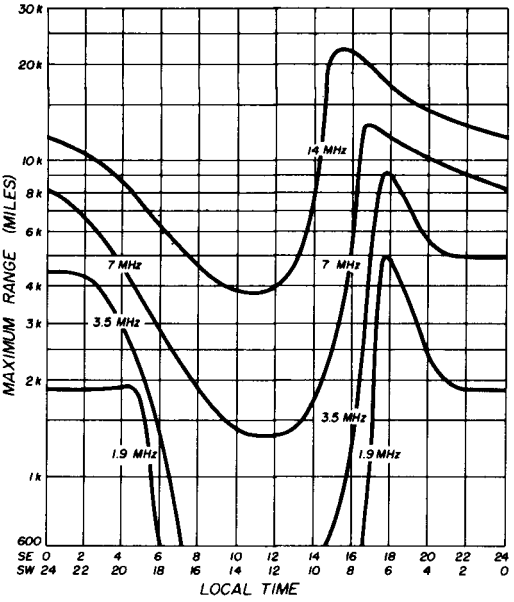


fig. 5. Maximum range to the southeast (top time scale) and to the southwest (lower time scale) due to absorption.

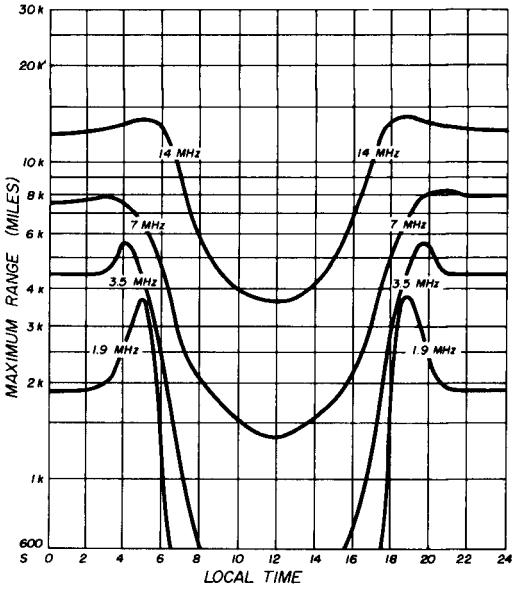


fig. 6. Maximum range to the south due to absorption.

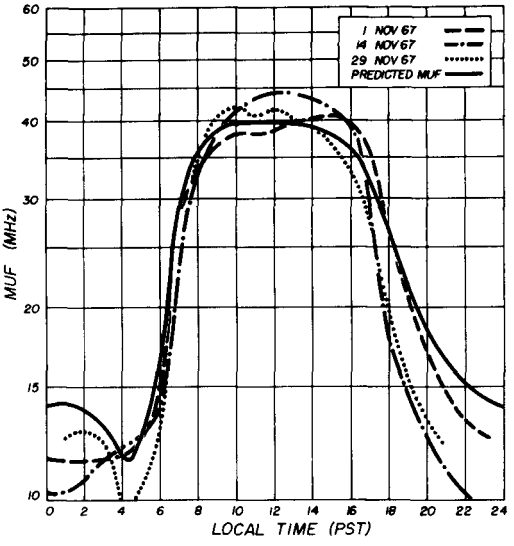


fig. 7. Graph of scaled muf's vs time of day for 1, 14 and 29 November 1967.

hours of sunset. The greatest uncertainties in this month's predictions are in the daytime muf's. Daytime muf's between 20 and 65 degrees North latitude are expected to be significantly higher during November than during October.

During the early afternoon (time of maximum muf's), the muf should increase by about ½% per degree of latitude between 30 and 50 degrees North latitude. Nightime muf's during November are expected to be significantly lower than during October. Twenty, fifteen and ten will open later and close earlier as the month progresses. Peak muf's will occur closer to local noon.

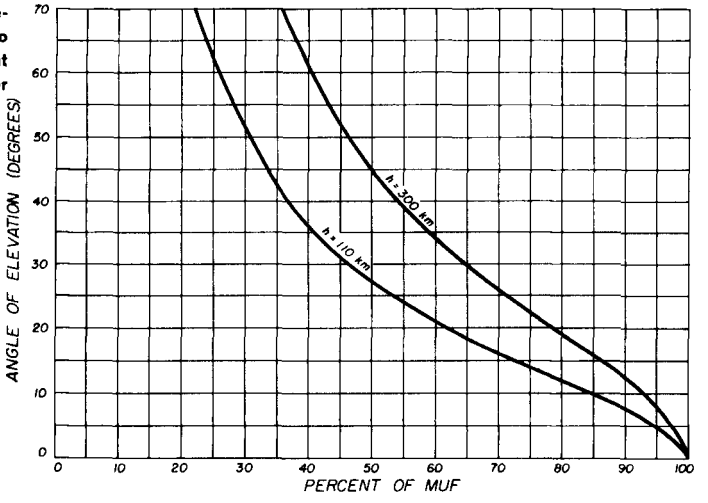
If the six-meter band has been open for single- or multi-hop F2-layer east-west paths during October, six-meters may open during November for two-hop paths between the West Coast and Japan and between the East Coast and Southern Europe peaking at 0000Z and 1600Z respectively, for about the same number of days. The predicted path muf for the two paths is only about 38 MHz. In any case, six-meter openings from the southern half of the U.S. *to the south should be possible at least* 20% of the days of the month.

Fig. 1 is a time chart of the predicted 4000-km F2-layer muf for November 1968. Maximum range vs time of day charts for 160 to 20 meters are shown in **fig. 2** to **6.** The box on page 76 tells you how to use these diagrams.

A graph of scaled muf's vs time of day for November 1, 14 and 29, 1967 along with muf (4000) F2 predictions for the location of the ionospheric sounder (35.5° N) is shown in **fig. 7.** The scaling procedure was discussed in the September column.

Fig. 8 is a graph of the maximum angle of elevation that is returned to the earth

fig. 8. Maximum angle of elevation that will be returned to the earth for a frequency that is a given percent of the layer muf.



how to use these propagation charts

1. To find the maximum usable frequency for F2-layer propagation in any direction, read the frequency at your control point on the muf time chart. Your control point is 1200 miles away from your station in the direction of propagation, about 18 degrees difference in latitude for a north-south path, or 1½ hours difference in time for an east-west path. To help you find the control point, a control-point circle has been drawn on the muf chart, centered on noon and 38 degrees N latitude. An overlay of this circle may be made on plastic or transparent paper and shifted so that the center of the circle is at your latitude and local time. Greenwich Mean Time may be used instead if you make an arrow along the bottom of the overlay at a time (your longitude in degrees W divided by 15) in hours to the right of center.

2. Over any particular path longer than 2500 miles, the path muf is the lower of your and the other station's control-point muf. Curved lines may be drawn on the overlay representing the great circle path. The great circle path may be found from a globe, a great circle chart or curves in NBS publications.[2, 3] However, this muf time chart (for a longitude of 90° W) will be somewhat in error outside the range of longitudes between 45° W and 135° W, and is least representative of actual conditions between 0° and 180° E longitudes.

3. To find the maximum propagation distance for 160, 80, 40 and 20 meters as limited by ionospheric absorption and atmospheric noise, refer to fig. 2, 3, 4, 5 or 6, depending on the direction you wish to work. Note that the time scales are reversed for westward propagation in fig. 3, 4, and 5. These curves are based on unity signal-to-noise ratio in a 6-kHz bandwidth with 100 watts output power (equivalent to 90% copy with 100 watts CW or 800 watts ssb) on 80 to 20 meters and ⅛ that power on 160 meters. The combined receiver and transmitter antenna gains over an isotropic radiator are −12 dB for 160 and 80 meters, 0 dB for 40 meters and +12 dB for 20 meters. On 10 and 15 meters, the communications range should not be limited by absorption to less than one transit around the earth, although marginal round-the-world propagation will occur at times of minimum range for opposite direction on 14 MHz.

The maximum distance curves were derived from consideration of atmospheric noise levels (from CCIR report 322) and calculated path losses at fixed distances in each direction from 38° N latitude. Only minor differences in maximum range would be noted due to change in absorption for stations located between 26° N and 50° N latitude. Somewhat greater daytime ranges could be expected at more northerly latitudes, and somewhat lesser daytime ranges could be expected at more southerly latitudes. The muf time charts were prepared from basic propagation predictions published monthly by the Institute for Telecommunications Sciences (ITS), Boulder, Colorado and available through the U.S. Government Printing Office.

for a frequency that is a given percent of the layer muf. There are two curves, for **true** heights of reflection of 110 km (E-region) and 300 km (F2-region). These curves indicate the importance of good low-angle coverage at frequencies near the muf. They also indicate that if your horizon is blocked, your muf will be lower than the layer muf.

Unfortunately a time chart of muf (2000) E is not available. It would indicate noontime E-layer muf's of about 16 MHz and night-time E-layer muf's of about 4 MHz. For frequencies below the E-layer muf there is a screening angle given by the 110-km curve. Below this angle the F2-layer is not illuminated, and paths longer than 2500 miles will have excessive attenuation. Thus, for long paths and frequencies below the E-layer muf, there may be both an upper and a lower limit to the angles of elevation that are propagated. If the screening angle is too high (depending on the path length) there may be no propagation at any elevation angle with low enough path loss.

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2. NBS Circular 462, "Ionospheric Radio Propagation," U.S. Government Printing Office, Washington, D.C., 1948, p. 60-61, 70.

3. K. Davies, "Ionospheric Radio Propagation," National Bureal of Standards Monograph 80, U.S. Government Printing Office, Washington, D.C., 1965, p. 144.

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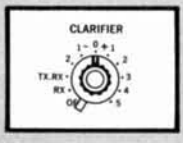

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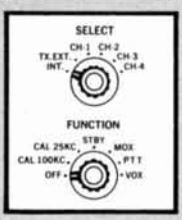

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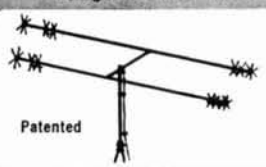

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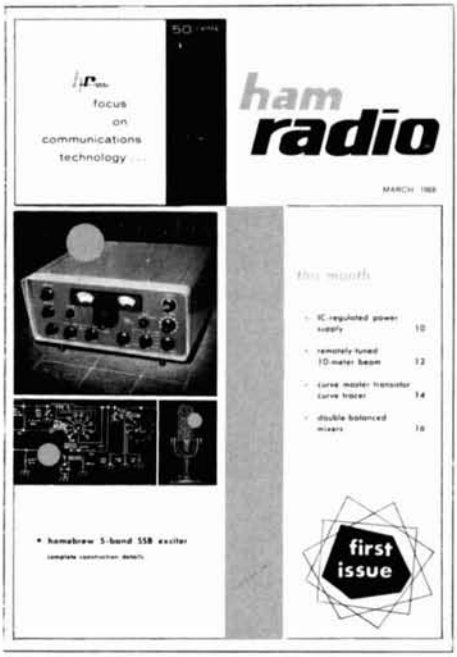

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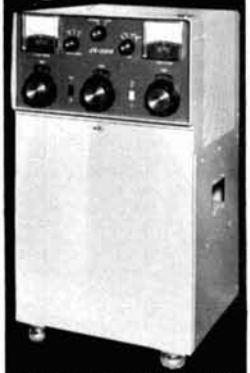

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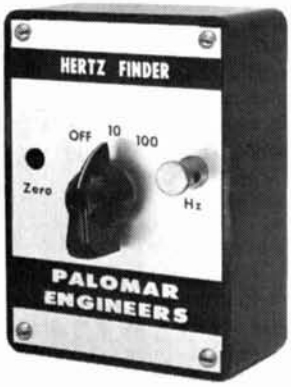

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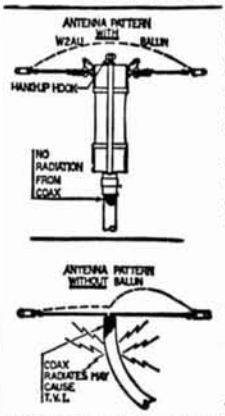

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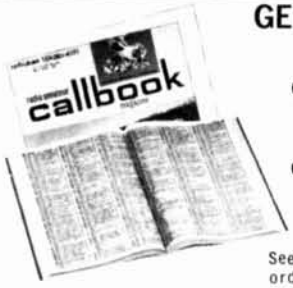

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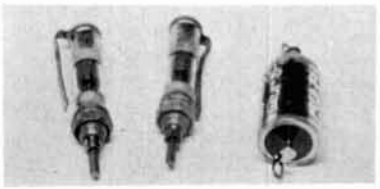

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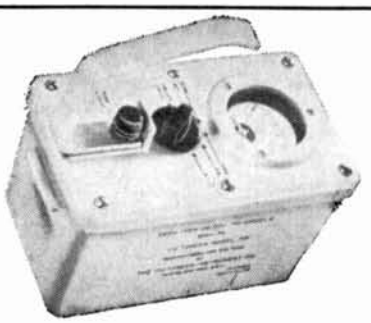

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SBT-20A **$350.00**
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TR-35A **$200.00**
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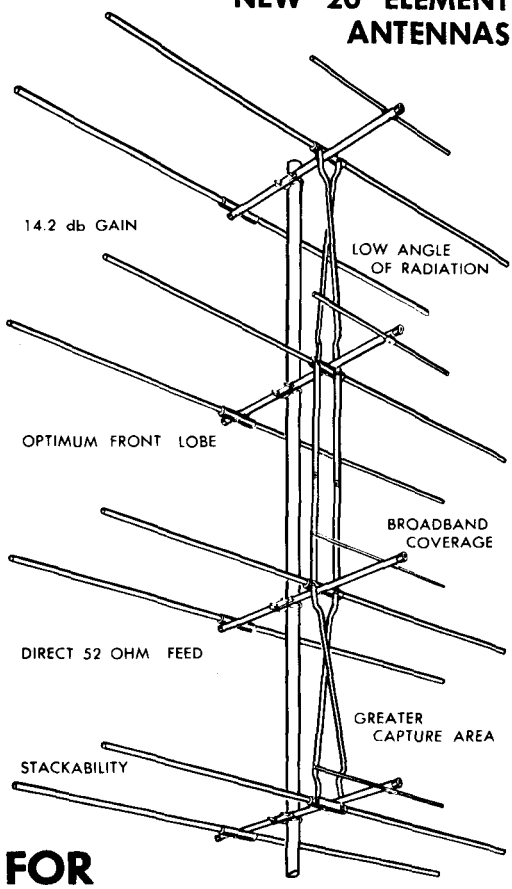

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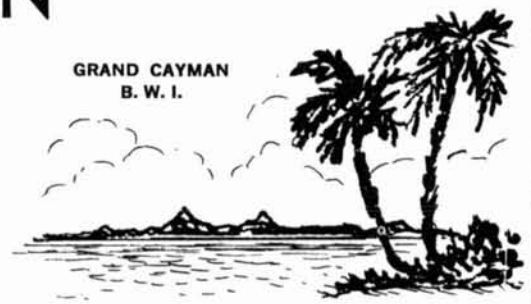

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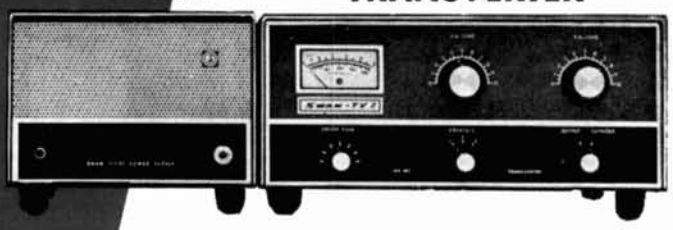

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Some Antenna Notes

Because an antenna is the most important part of your station

To start with, neither **QST** nor **Ham Radio** Magazine accepts antenna manufacturer's gain figures in their present advertising. This is because there is not any industry standard for measuring gain, a problem that is presently being worked upon by a committee of the EIA. Some manufacturers in their published literature compare the performance of their product with a somewhat theoretical device called an isotropic source, while others compare against a conventional half-wave dipole. Unless **You** know what the reference is you will likely not obtain accurate comparisons and, therefore, have false ideas of the product's performance.

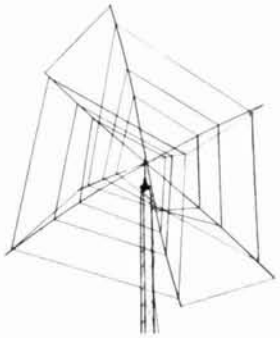
All other factors being equal, low radiation angle and maximum capture area are the most important antenna considerations. Long wire antennas, drooping verticals, and verticals all have their place — for certain kinds of work. But it should be realized that a vertical will radiate a donut shaped pattern in all horizontal directions while a horizontal antenna will radiate a half-donut sideways and upward. Upward doesn't count. A vertical is therefore very advantageous, but must usually have a good ground or counterpoise system.

When an antenna is excited at its resonant or fundamental frequency, it can radiate energy very much better than a random length of wire turned through a matchbox.

The useful frequency band of a particular type of an antenna is inversely proportional to its design frequency and is directly related to the design of the antenna and its installation. Typically, a long wire antenna of No. 12 wire cut at 3900 kHz will range plus or minus 12½ kHz for good results. At 40 meters this will double, at 20 meters the width will be a good hundred kHz while at 10 meters the response will typically be 300 or more kHz. The greater the cross section of the antenna the broader the response.

Multi-band antennas are **not** normally optimized for each band of operation. They will exhibit excellent results on one band, good on a second and fair response on a third. Typical tri-band beams or quads have an ideal electrical spacing on one band only and offer compromise spacing and feed matching on the others.

A theoretical antenna at resonance is purely resistive and, therefore, can offer the best match at this frequency. Since the maximum radiation occurs at the resonant frequency of the antenna, amateurs



need to know the actual response curves of their antennas. Several devices exist by which antennas may be evaluated. Rf impedance bridges, antenna scopes, antenna bridges or vswr meters may be used. But it must be realized that the typical vswr meter is inserted at the transmitter instead of where the most meaningful results can be obtained, right at the antenna feed. When inserted in the feed line at the transmitter the vswr meter reflects the terminating impedance of the coax, not the performance of the antenna although a resonant and properly matched antenna will exhibit a low vswr.

Transmission line loss invariably causes an artificially lower vswr than in reality exists at the feed point of the antenna. The loss in the feed line lowers the reflected power measured at the transmitter, but in no way improves the antenna itself. The reflected power as shown by the vswr meter is simply an indication of your complete antenna system's efficiency at the frequency you are working on.

It will help, therefore, to recognize that the practical standard of most manufacturers is a maximum 2 to 1 vswr. For maximum performance and reliability you should try to operate within this limit. One practical result of these notes is for the reader to prepare for each of his operating bands a graph showing vswr versus frequency and then to learn to operate within the limits of his antennas.

A possible second approach is to buy a Regina f321A three band quad for operation on 10, 15 or 20 meters. This exclusively distributed product is the only commercial device of its kind whose vswr is guaranteed not to exceed 1½ to 1 over each band. The gain of the Regina f321A is optimum and the new all aluminum version is mechanically superb. It is still priced at only $90.00 f.o.b. Harvard. Previous ads describe our product more fully or you may write for this information today.

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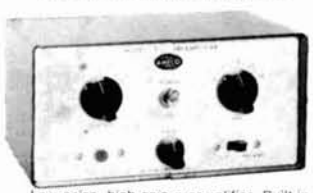

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