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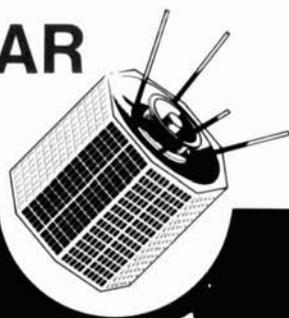
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APRIL, 1974

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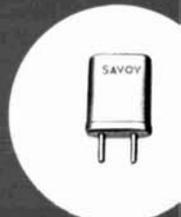


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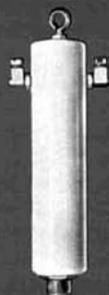


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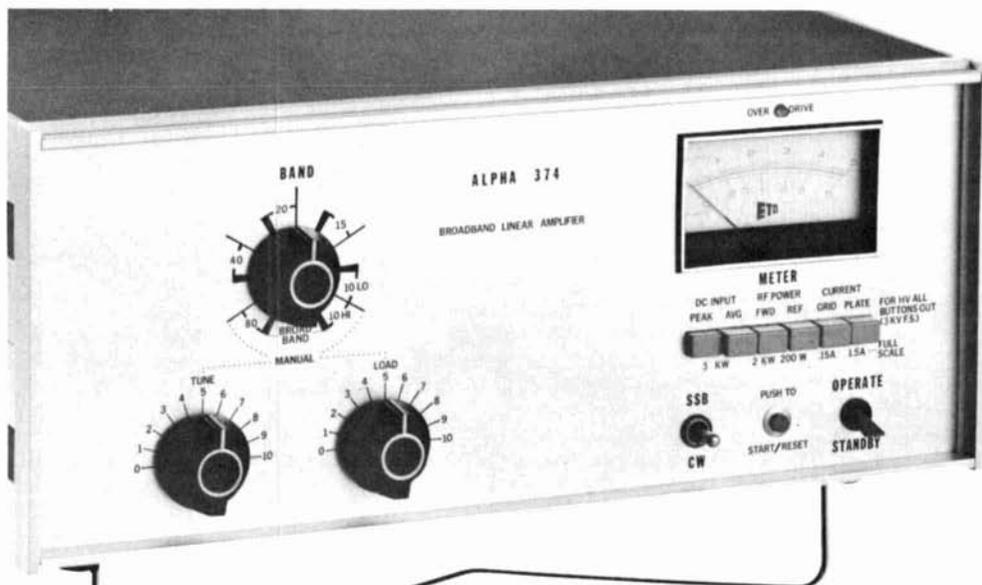
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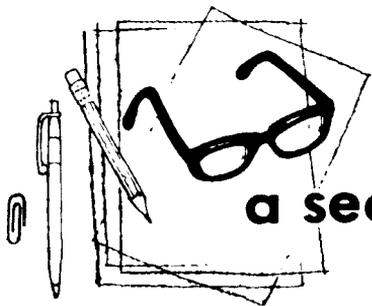
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a second look

by jim
fisk

It's **spring** and hamfest time again. No matter where you look you see an announcement for yet another convention, auction, hamfest or fm talk-in. The Dayton Hamvention in Dayton, Ohio, which is billed as the original hamvention, is one of the biggest of the year. Drawing upon a large amateur population in the Midwest, the Dayton show has provided the model for many successful amateur conventions around the country. Growing by leaps and bounds in recent years, more than 6400 hams were in attendance last year and 8000 are expected this year when the Hamvention opens its doors the last weekend in April.

This year the Dayton Hamvention Committee has gone all out to ensure a lively, interesting weekend for all. Bright and early Friday morning, the 26th of April, amateur radio manufacturers and distributors will start setting up their exhibits. At twelve noon the exhibition doors will be opened to the public. That evening the Old Old Timers and Quarter-Century Wireless Association will hold a dinner meeting in downtown Dayton.

By Saturday morning things will really start booming around Hara Arena. Vendors and traders from miles around will be setting up shop for the famous Dayton Flea Market, and the three-hour DX and vhf forums will be kicked off. A special ladies program, including luncheon, will begin at 11:30 AM, and an ARRL Forum is scheduled for the afternoon, followed by technical sessions on amateur television and troubleshooting. The traditional Saturday-night cocktail hour and banquet begins at 7:00—Senator Barry Goldwater, K7UGA, will be the guest speaker.

Sometime before the Hamvention is opened to the public, a 430-MHz transmitter will be hidden somewhere in the

Arena area. Transmitter hunts will start at 1300 on Saturday and Sunday. If you're going to Dayton and want to join the fun, write to Rudy Plak, W8ZOF, for an antenna design.

Sunday morning the flea market and exhibit area will open at 9:00 AM, and the *antenna and state-of-the-art forums* will get under way. In the afternoon there will be forums on fm and repeaters and space communication. In addition, there will be technical and group meetings for ARPS, MIDCARS, OSSB and MARS. Other special groups attending the Hamvention are the Ohio Sideband Net, Buckeye Belles, Country Cousins, Poverty, Cracker Barrel, Firebird, Post Office, Intercontinental Traffic and Young Ladies Radio League. Prizes will be awarded at the end of each technical session on Saturday and Sunday. If past performance and the 1974 schedule are any indication, it should be another great show.

For amateurs who arrive in trailers and campers, parking will be permitted in designated areas. For those who stay at hotels downtown, free bus service will be provided out to the Arena. An allotment of 500 rooms has been set aside for the Hamvention by the local hotels and motels, so all room requests should be directed to the Accommodations Committee so that rooms can be allotted within the available supply. For more information, and a Hamvention brochure, write to the Dayton Hamvention, Post Office Box 44, Dayton, Ohio 45401.

If you've never been to the Dayton Hamvention, but have considered it, this is the year to go. If you've been before, you already know what I'm talking about. See you there!

Jim Fisk, W1DTY
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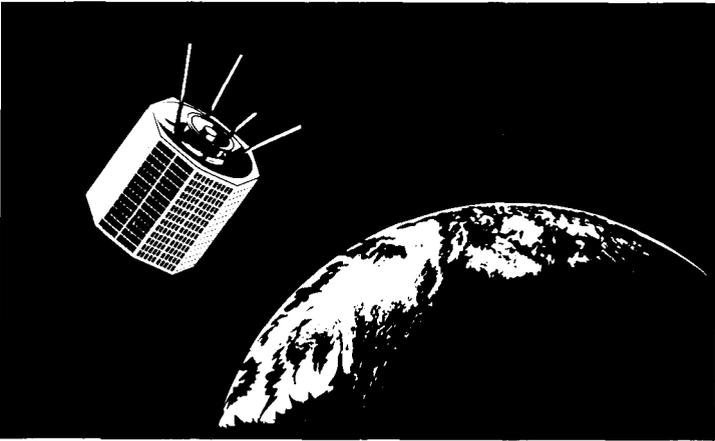
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communications techniques for OSCAR 7

A discussion of
the new OSCAR 7
amateur radio
communications satellite
and the equipment
for using it

Joe Kasser, G3ZCZ, 1701 East-West Highway, Apt. 205, Silver Spring, Maryland

The new OSCAR 7 communications satellite which will be launched into orbit in the near future is the most complex amateur-radio satellite built so far. It is the second in the AMSAT-OSCAR-B series of long-life amateur spacecraft, and is built in an octahedral configuration which provides surface area for enough solar cells for a positive power budget system. This means that, unlike OSCAR 6, it should not be necessary to periodically command the spacecraft into recharge modes.

OSCAR 7 will contain two repeaters and two auxiliary beacons, as well as Morse code and telemetry encoders. The two- to ten-meter repeater has an output power of two watts so signals received on the ground will be somewhat stronger than those received from OSCAR 6. The second repeater, which was built by a West German group, AMSAT-Deutschland, has an input at 432 MHz and an output at 146 MHz. The two beacons will be at 435.1 and 2304 MHz.

Ground control of the spacecraft is provided by command receivers in each repeater, redundant command decoders and a control-logic sub-system experiment. The whole spacecraft system was described in detail at the ARRL Technical Symposium on Space Communications.¹ The purpose of this article is to

present some ideas and techniques for using the new satellite for amateur communications.

working through the two-to ten-meter repeater

The ground equipment necessary for working through the OSCAR 7 two-to ten-meter repeater is identical to that required for use with OSCAR 6. Since the downlink signal will be transmitted with approximately twice the power of the OSCAR 6 transmitter, this will allow some (but not much) relaxation in the receiving equipment requirements. A ground-based transmitter with an output on the order of 80 to 100 watts effective radiated power (erp) will again be suitable. The same receiving antennas may be used.

The preferred antennas for both transmitting and receiving are simple non-directional ones such as a 5/8-wavelength vertical or a turnstile. Although the satel-

Also, in region 1 the amateur two-meter band spans only the frequencies from 144 to 146 MHz. Any transmissions above 146 MHz are non-amateur. In England, for example, the police use frequencies at 146 MHz for mobile communications and these signals were re-transmitted through OSCAR 6.

In the United States, frequencies above 146 MHz are used as input frequencies for two-meter repeater installations. Stations communicating through OSCAR 6 and working DX stations on their "own" frequency above 146 MHz were also being copied through their local repeaters. The simplest solution to these problems was to move the input passband to 145.850 to 145.950 MHz.

working through the 432-to 145.9-MHz repeater

Working through the 432- to 145.9-MHz repeater will be very much

table 1. OSCAR two- to ten-meter repeater passbands (± 3 dB points).

satellite	uplink	downlink	beacon
OSCAR 6	145.900-146.000 MHz	29.45-29.55 MHz	29.45 MHz
OSCAR 7	145.850-145.950 MHz	29.40-29.50 MHz	29.50 MHz

lite's two-meter antennas are circularly polarized, it is also preferable to have some sort of ten-meter polarization diversity so you will be able to receive both vertically and horizontally polarized signals.

The passband and beacon frequencies for the OSCAR 7 two- to ten-meter repeater are slightly different from those used in OSCAR 6 (see table 1). These new passband frequencies were chosen for several reasons. First of all, in region 1 (Europe) 145.950 to 146.000 MHz is used by beacon transmitters operating on a 24-hour-a-day basis. These beacons are used for propagation studies, setting up converters and as a general guide to the vhf propagation conditions prevailing at any time. These beacons, although transmitted through OSCAR 6, provided no communications service and unnecessarily drained the power supply.

like working through OSCAR 4, but in reverse. OSCAR 4 received signals on 144 MHz and re-radiated them on 431.9 MHz. The OSCAR 7 repeater receives signals on 432.1 MHz and retransmits them on 145.9 MHz. The OSCAR 7 repeater also features sideband inversion so that, for example, an upper-sideband (USB) input signal will be re-radiated as a lower-sideband signal. At present, many more stations are equipped to copy 144-MHz ssb than are equipped to transmit ssb on 432 MHz, and the convention is to use USB on 144 MHz. AMSAT suggests that USB be adopted as the standard for the uplink to make it possible to easily distinguish between satellite (LSB) and terrestrial (USB) signals on two meters.

The recommended transmitting power is 300 to 400 watts erp. This is best achieved by means of a high-power transmitter and a simple antenna. A ground-

plane, 5/8-wavelength whip or turnstile antenna do not require any pointing during the orbital pass, allowing the operator to concentrate on the important business of communicating.

Alternative methods of generating the required rf power are to use converted uhf fm transmitters, frequency converters or surplus tripler-amplifiers. Since the uhf repeater is a linear device, as is the two- to ten-meter unit, the recommended modes of operation are ssb and CW.

An alternative to ssb for voice transmission is series-grid amplitude modulation, also known as controlled-carrier modulation. With this system the modulation is applied to the screen of the final amplifier tube, the carrier output is set to a low level without any modulation, and the modulation controls the level of the carrier. Thus, the louder you talk, the more power you put out. Modulation can be set to a maximum of 95%. A suitable circuit is described in reference 2 and shown in fig. 1. The S-meter will fluctuate with the modulation. This can be annoying or impressive as the case may be.

The received signal cannot be distinguished from conventional plate-modulated signals by audio means. In fact, many ssb operators will not notice that the incoming signals are a-m. You might, however, get reports of excessive carrier on your signal. I know of one G3 who was using series-grid modulation on 20 meters about four to five years ago and he received many stateside QSL cards for two-way ssb contacts!

copying the RTTY telemetry

From the telemetry point of view, the biggest difference between OSCAR 7 and the previous amateur spacecraft is the fact that OSCAR 7 has the facilities for transmitting telemetry by means of RTTY. The RTTY will be FSK on the 435.1-MHz beacon and AFSK on the 145.98- and 29.50 MHz beacons. The Doppler shift will be about ± 10 kHz on the 435.1-MHz beacon, about ± 3 kHz at 145.98 MHz, and ± 600 Hz at 29.5 MHz.

The effects of Doppler shift on the telemetry signal will be copied by the ground station as a change in the carrier frequency and not as a change in the modulation frequencies.

An extremely simple arrangement will enable good copy of the AFSK signals on

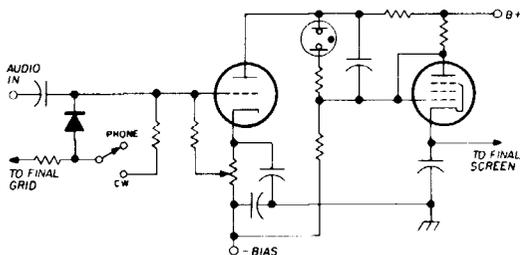


fig. 1. Basic circuit for a series-gate modulator (see text).

145.98 MHz. A typical set-up is a two-meter a-m receiver with a bandwidth of 5 to 10 kHz. A front-end preamplifier should be used between the antenna and the receiver. To copy a satellite pass it will only be necessary to tune to the low side of the signal for acquisition. The Doppler effect will cause the carrier frequency to have the appearance of a slow drift through the i-f passband of the receiver. The signal should still be within the passband at loss of signal (LOS). The detected RTTY tones do not vary during the pass so the terminal unit receives the correct tones throughout the pass. A terminal unit such as the ST-5 (see reference 3) would be very suitable for this application.

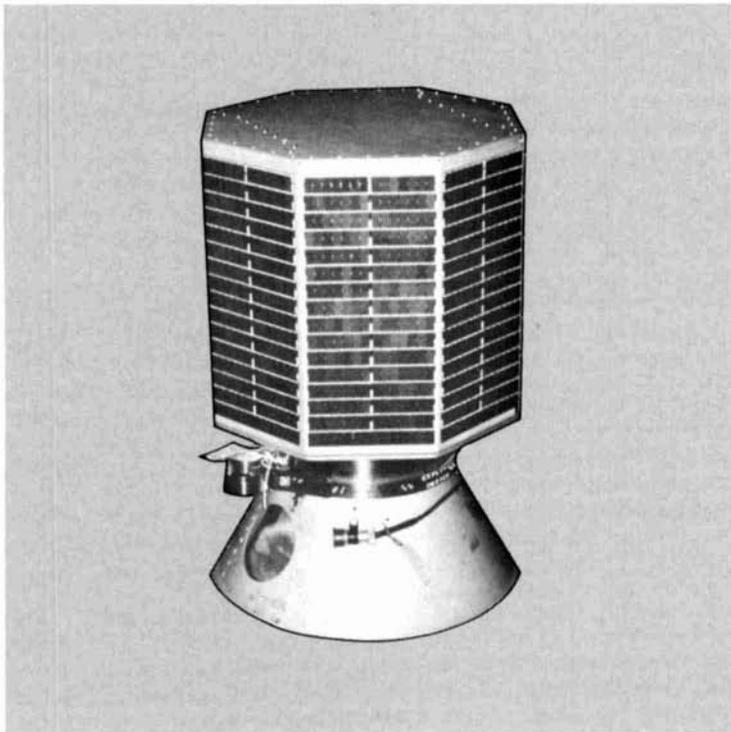
Copying the FSK transmissions on 435.1 MHz will be slightly more difficult. Tests were conducted from club station WA3EWJ in March, 1973, to determine the feasibility of copying satellite FSK RTTY signals with simple ground equipment, and to investigate how Doppler shift would affect the received signals. It was found that copying the teletype signals was quite easy. The terminal unit used was the ST-5.

The Doppler shift on the signal was found to appear as a gradual drift downward in frequency throughout the pass.

The received signal was monitored using the conventional cross-shaped oscilloscope pattern, and the vfo on the receiver was adjusted to keep the display correct. Frequency-selective fading was also observed on the signals. Although the ST-5 provided good copy (even after two tape

mark frequency and puts out a proportional positive or negative voltage (with respect to ground) as a function of how far (and which way) the receiver is tuned off the signal. If a center-zero meter is connected across this afc output, it will act as a tuning meter, showing how the

The AMSAT-OSCAR-B communications satellite which will be orbited as OSCAR 7 later this year. In this photograph the spacecraft is mounted on a sheet-metal cone which is installed in the launch vehicle. The OSCAR package is separated from this cone and ejected into a sun-synchronous orbit approximately 900 miles above the surface of the earth. The antennas, which are stainless steel tapes, are unfurled after the spacecraft goes into orbit. OSCAR 7 will carry 145- to 29-MHz and 432- to 146-MHz repeaters as well as beacons on 435.1 and 2304 MHz.



transfers), it was necessary to stay at the receiver to adjust the vfo every minute or so. It would be much nicer if the receiving system could tune itself, automatically tracking the Doppler shift, freeing the operator for other tasks.

Two designs for automatic frequency control of received RTTY signals have been published in recent years. Reference 4 describes a terminal unit using a phase-locked loop. The use of such a terminal unit allows the signal to drift through the passband while printing out good copy. However, when the signal moves out of the passband, copy is lost.

Reference 5 describes an add-on unit for the ST-5 or ST-6 terminal units which consists of a circuit which monitors the

receiver is tuned with respect to the incoming RTTY mark signal.

Reference 5 also describes how to modify the receiver vfo to accept and use the automatic frequency control signal. An alternative approach is to build a new external vfo for use with OSCAR 7. If a new vfo is to be built, a better approach is to build an afc-controlled oscillator for the front end of the 435.1-MHz converter.

copying the s-band beacon

Link calculations for a typical receiving station for the 2304-MHz beacon were given in a previous article.¹ It was shown that with a Doppler shift of about ± 55 kHz, a receiver bandwidth of 500 Hz and a four-foot dish with a pointing

accuracy of $\pm 7.5^\circ$, reception of this beacon presents a real challenge.

A simple front end converter for 2304 MHz is described in reference 6. It uses the classic trough-line front end based on earlier 1296-MHz units. Antenna construction plans are given in reference 7 in an article describing a pulse communication system. Amateurs who are currently tracking OSCAR 6 in both azimuth and elevation using a narrow-beamwidth antenna should already have the ability to track OSCAR 7 with a four-foot dish. Thus, reception of this beacon is not quite as difficult as it appears at first glance.

using medium-scan television

In most countries, wideband TV is an authorized mode of transmission in the 432-MHz amateur band. Since OSCAR 7 contains a repeater having an input in the 70 cm band it opens up the prospects of live, long-distance, real-time TV communications.

However, since the signals are re-radiated on 145.9 MHz, a waiver or special permit must be issued by the licensing authorities. This permission has already been requested of the FCC. Also, since the repeater has only a 50-kHz passband the transmissions will still be limited in bandwidth. This rules out standard fast-scan (525/625 line) pictures. Slow-scan TV with its eight-second frame rate is suitable (as demonstrated by a number of OSCAR 6 contacts) but could be improved upon, at least with respect to the frame rate.

Reference 8 describes a medium (or faster slow-scan) TV system. The specifications for this system are such that all frequencies used are four times the equivalent normal slow-scan rate as shown in table 2. The pictures thus have a frame rate of 2 seconds. These TV pictures are not currently used for on-the-air transmissions because the bandwidth is also four times the normal slow-scan TV bandwidth. The pictures are used for setting up cameras and monitors in the home station and are converted to slow-scan TV

table 2. Operating parameters of slow-scan and medium-scan television signals.

parameter	sstv	mstv
Line rate	15 Hz	60 Hz
Number of lines	120	120
Frame time	8.1 sec	2.025 sec
Sync frequency	1200 Hz	4800 Hz
Black frequency	1500 Hz	6000 Hz
White frequency	2300 Hz	9200 Hz
Horizontal sync pulse	5 ms	1.25 ms
Vertical sync pulse	30 ms	7.50 ms
Video bandwidth	900 Hz	3600 Hz

by the very simple method of dividing all frequencies by a factor of four (using digital ICs). The pictures are then indistinguishable from conventionally generated sstv.

Since the bandwidth of medium-scan TV (mstv) is less than 15 kHz, and the spacecraft repeater has a 50-kHz passband, there does not appear to be any reason why mstv should not be used as a communications mode through OSCAR 7, providing that the relevant permits are issued by the licensing authorities. With its two-second frame time, mstv is a vast improvement over sstv eight-second frame time.

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simple active filters for direct-conversion receivers

How to design
simple active
audio filters
for radio
communications

Many of the simpler communications receivers in use today are lacking in selectivity for CW work. To some extent this deficiency can be minimized by using audio filters. The more interesting designs, at least for me, are those where active RC circuits are used to replace the classic LC configurations. While high

quality active filters are available commercially,* they are also popular as projects for the amateur experimenter. Such filters are especially useful in conjunction with direct-conversion receivers, an approach in which all adjacent channel selectivity must be obtained at audio frequencies.

Several of the active filter designs available to the experimenter suffer from problems which make them less than optimum for general use. For example, many of the designs are aimed at achieving rather narrow bandwidths, often less than 100 Hz. While these units are quite useful for some specialized applications, I prefer a somewhat wider bandwidth for general CW work. A 0.5-kHz wide response is usually more than adequate if steep skirts are maintained.

Many of the popular active filter designs require tight control of component tolerances, leading to difficulty and excessive expense when being duplicated. The work described in this article is aimed at designs which use $\pm 10\%$ or even 20% tolerance components and feature wider bandwidths while maintaining sufficiently steep skirt response to be useful.

*Such as those manufactured by MFJ Enterprises, Post Office Box 494, State College, Mississippi 39762 (see *ham radio*, November, 1973, page 68).

Wes Hayward, W7ZOI, 7700 SW Danielle, Beaverton, Oregon 97005

lowpass filter

Shown in **fig. 1** is an abbreviated schematic of a 10-pole peaked lowpass filter I built. Although only two lowpass sections are shown, the filter contains five identical sections. Each of these has a 2-pole lowpass response which is peaked at the cutoff frequency. A similar response is that of the pi-network in a tube transmitter, again a peaked lowpass filter.

The Q of each active filter section is about 1.9 which yields a net 6-dB bandwidth of about 200 Hz. Skirt response, however, is not lacking. With a center frequency of 540 Hz, the attenuation is 75 dB at 1200 Hz. Net gain of the system is 28 dB at resonance. A single-pole highpass section is used at the input to bias the following stages and to provide some additional attenuation at the low frequencies.

The measured responses for one, three and five lowpass sections are plotted in **fig. 2**. This filter was built with 10% capacitors and resistors. However, the low Q of each pole pair would allow the use of 20% components with a minimal degradation in performance. Indeed, the slight stagger-tuning effect that could result might be quite desirable. The npn-pnp feedback transistor pairs are used as unity gain amplifiers and are not

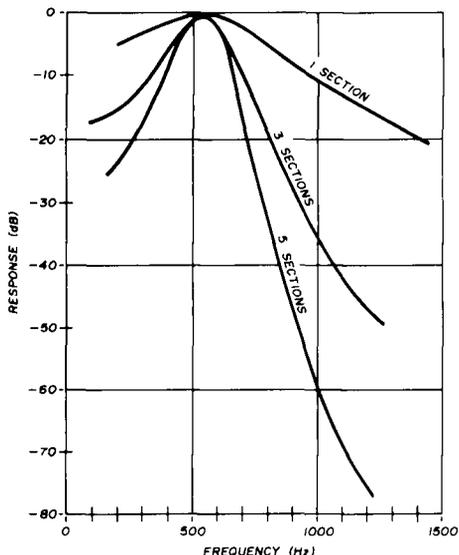
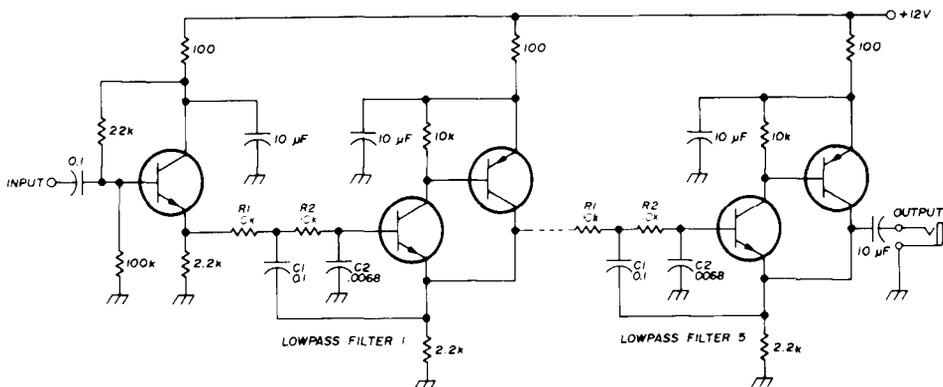


fig. 2. Measured response of one, three and five lowpass sections of the 10-pole active filter shown in **fig. 1**.

critical as to transistor type. The bias of the input section was chosen to compensate for the 0.6-volt offset introduced by each feedback pair, placing the filter output at half the power supply level.

From a practical point of view, the filter has been found to be an excellent performer. The moderately wide bandwidth makes the unit easy to use, even



- C1 0.1 μ F, 10% tolerance
- C2 0.0068 μ F, 10% tolerance
- R1, R2 10k, 1/4-watt, 10% tolerance

fig. 1. Abbreviated schematic of a 10-pole peaked lowpass filter. All five lowpass sections are identical. Npn transistors are 2N3565, 2N3904 or similar; pnp transistors are 2N3638, 2N3906, etc.

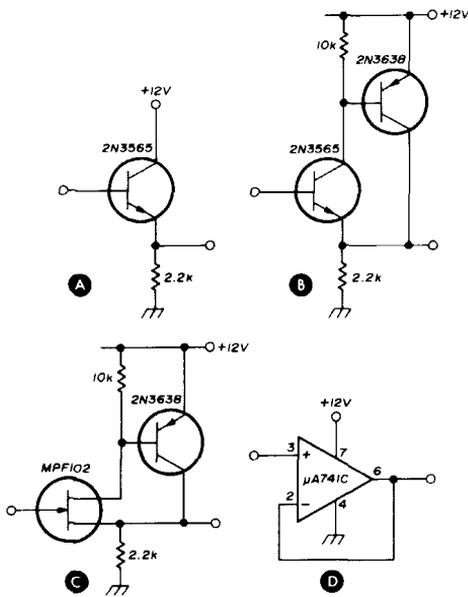


fig. 3. Four different types of unity gain, non-inverting amplifiers suitable for use in active filters.

with receivers with minimal bandwidth. However, the steep skirts insure adequate rejection of adjacent channels. When used with even the most simple direct-conversion receiver, performance is suitable for the majority of amateur communications.

other filter designs

The design outlined above should be suitable for those wanting a circuit to duplicate. However, the low cost of modern semiconductors and the ease of construction of audio circuits make active filters a very attractive area for further experimentation. The remainder of this article will present some possible variations for you to try in your own lab. If you have an analytical bent you will find

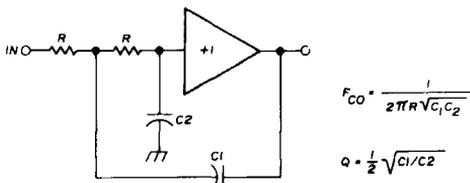


fig. 4. Basic peaked lowpass filter section. The cutoff frequency, f_{co} , and Q are defined by the values of R and C as shown here.

the analysis of the peaked lowpass design to be straight forward by using either classic methods¹ or the real-time approach outlined earlier.²

For the simple lowpass filter designs, the amplifier should have a unity, non-inverting gain and should exhibit good impedance-transforming properties. In many cases, a simple emitter follower using a high beta transistor will suffice. Integrated-circuit operational amplifiers such as the popular $\mu A741$ are excellent, although a lower noise type such as the LM-301 is sometimes preferred. Shown in fig. 3 are four possible configurations, all suitable for use with a single power supply. The fet input amplifier is useful for low noise applications.

The basic peaked lowpass filter design is summarized in fig. 4. Note that the Q of the circuit is completely defined by the ratio of the two capacitors. When designing a filter, a Q is chosen and convenient capacitors of standard value are then picked. Miniaturization and low cost would suggest choosing relatively

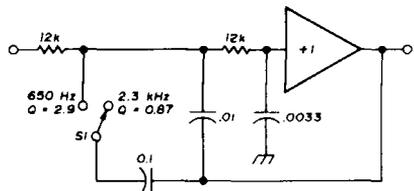


fig. 5. Dual bandwidth active filter. With switch S_1 open, cutoff is 2.3 kHz; cutoff is 650 Hz with switch closed.

low capacitor values. However, noise considerations often point toward the use of somewhat larger values. Once the capacitors are chosen, the frequency of the unit is determined by calculating the proper resistance values. Clearly, a tunable filter with a constant Q would result from the use of a dual potentiometer.

The filter shown will have unity gain at dc and a voltage gain equal to the Q at the "cutoff" frequency. The measured peak frequency will be slightly lower than the "cutoff" frequency defined by the equation of fig. 4. This error is largest at

low-Q values. For example, the measured center frequency of the 10-pole filter of **fig. 1** is 10% lower than that calculated. This effect is characteristic of any low-Q resonator.

The fact that both frequency and Q are dependent upon the capacitors can be used to advantage. Shown in **fig. 5** is a dual bandwidth filter. With the switch open, the cutoff is 2.3 kHz and Q is 0.87. When the switch is closed, the center frequency drops to about 650 Hz and Q increases to 2.9. A multiplicity of these sections should yield a filter useful for both CW and ssb work.

Obtaining electronic components is often a problem for amateurs, with precision capacitors being especially difficult to find. Occasionally, you will come across a large number of capacitors of identical value. These can be used in filter applications in conjunction with non-inverting amplifiers with a gain greater than unity. This lowpass filter configuration is presented in the schematic and equations of **fig. 6**. You can see that Q can become infinite for a closed loop gain, A, of only 3. This oscillating condition could be used to advantage in a transceiver application by using an fet or bipolar switch to alter stage gain, providing a simple sidetone function.

limiting amplifier

One drawback of many simple direct-conversion receivers is the lack of agc.

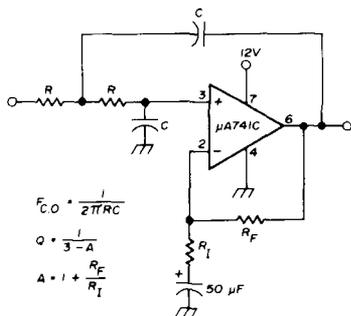


fig. 6. Alternate form of peaked lowpass filter using a non-inverting amplifier with gain greater than one. Cutoff frequency, f_{c0} , and gain, A, are defined by the resistance and capacitance components; Q is determined by gain.

This deficiency could be minimized for CW reception by careful application of limiting. For example, shown in **fig. 7** is a simple inverting, limiting amplifier with an adjustable limiting threshold. Below the threshold, the amplifier is linear with a voltage gain of 10. However, as soon as

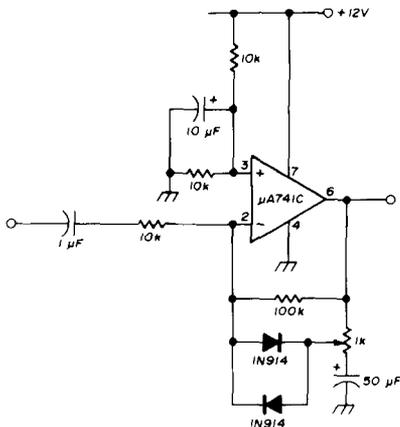


fig. 7. Simple inverting, limiting amplifier with adjustable limiting threshold.

the output is high enough for the silicon diodes to conduct, the gain drops below unity. This amplifier should be preceded by several sections of filtering, and followed by a single-section lowpass filter to eliminate the harmonic distortions generated in the limiting process. The resulting clipper-filter system would be a real ear saver when used with any receiver with poor or non-existent agc.

summary

There are many possible circuit configurations for active filtering and an equally wide variety of applications. Here we have considered only a very simple family of circuits. Perhaps this article will stimulate others to try their hand at this intriguing area.

references

1. Tobey, Graeme and Huelsman, "Operational Amplifiers, Design and Applications," McGraw-Hill, New York, 1971, chapter 8.
2. Wes Hayward, W7ZO1, "An RC Active Audio Filter for CW," *QST*, May, 1970, page 51.

ham radio

telefax transceiver conversion

Modifying the Model 6500A fax machine for F5 emission

Hams have long been an imaginative and ingenious bunch, taking older commercial equipment and converting it for their own use. Lately the Telefax Model 6500 facsimile transceivers have been on the market for 10 - 15 dollars each. Articles have been written on the conversion of these machines, but none have been very satisfactory as many of you may have discovered.

The machines were originally designed to work over landlines in conjunction with another setup located in a central office. This setup provided the sync detection and drum feed control for a pair of machines. The machine is basically an A4 emission device; that is, a varying

amplitude of the same frequency. One of the difficulties of A4 is the constant need for riding gain on the volume control to overcome band fading. Another problem exists in the legality of feeding this A4 emission into a 2-meter fm transmitter, which then makes F4 emission. This is not permitted under current regulations.

Included here are a set of electronics and standards to convert the final emission to F5, which is legal on 2-meter fm and which solves all the problems of fading on the low bands. The standards are simple; 2500 Hz for white, 2000 Hz for black, and the gray scale falls in between. The sync is a series of beeps at whatever frequency is coming out of the machine at that time. The first portion of the scan may be black, white, or gray — it makes no difference for the sync. Also included is an automatic drum feed. The receiving circuit detects the beginning of a picture and causes the drum to automatically begin the horizontal travel.

The schematic of **fig. 1** shows the unmodified Model 6500 fax machine. The following discussion describes the circuitry for adapting the machine for F5 emission and includes the sync and drum feed control.

interface circuits

Q5 and Q6 make up the multivibrator, which oscillates between 2500 Hz, the white frequency, and 2000 Hz, the black frequency (**fig. 2**). Q7 is the modulator, which detects how much output is com-

Fred J. Steurer, KØQMR, 11025 Patsy Drive, St. Louis, Missouri 63123

ing from the fax machine for a given picture and changes the oscillator frequency. Q8 is a simple emitter follower isolation. U2 is a flip-flop with a canceling input (fig. 3). On receive the phasing contact triggers this flip-flop, which drives Q3. Q3 in turn, drives PH-1, an LED lamp shining on a photo-resistor. This circuit provides isolation for the triac circuit, SCR 1. (Construction of PH-1 is shown in fig. 4.) Q4 is the triac control, which turns off the gate on the triac (SCR 1) when the phasing contact is triggered by the rotating drum. The triac interrupts power to the gray-colored motor (see fig. 1) to slow the rate of the revolving drum. When the received signal (beeps) coincides with the local phasing contact opening, the flip-flop is canceled and the gray-colored motor runs at normal speed. Both units (transmitting and receiving machines) are in sync. That is, the red line on each machine is in the same angular position.

U1 (fig. 5) is a standard limiter, which removes all a-m from the signal and causes a constant level regardless of the volume setting. The input is set for 500 ohms; and it would be a good idea to install a small transformer, 3.2 to 500 ohms, between the speaker terminals of the receiver and the 500-ohm input to match impedances.

The video detector is just a tuned trap in series with the limiter output and the fax input. Q1 and Q2 are the automatic drum feed and tuning circuit (fig. 6). Three seconds after video is detected, the relay closes and the drum feeds on the receiving end. U3 is a series-type regulator in the plus side of the power supply, fig. 7.

telexfax transceiver conversion

Remove the top cover and bottom plate and check the tubes. Check the stylus and replace if necessary with a piece of carbon-steel wire from a wire brush.

Carefully remove the exciter lamp, lens telescope, and projection tube. Remove the lenses from these assemblies and clean them. Replace the lenses in the

same order and in the same direction as removed. Remove the phototube, clean, and replace. Plug in the 117-volt line cord, and push the outgoing pushbutton. This turns on the lamp. Focus the light spot on the drum by moving the telescope back and forth. Put a piece of paper with typed letters on the drum, and focus the image on the pinhole of the projection tube by moving the projection tube back and forth. Take good care on this step if you are to transmit sharp pictures. Check that the red line on the drum is at the stylus position when the phasing contact is open and the free slack is taken up in the normally rotating direction. The factory setup may have slipped. Adjust with an Allen wrench. Burnish the phasing contacts and adjust them for 0.020 inch with a feeler gauge.

1. Clip the 51-ohm resistor (fig. 1) from the incoming switch on the front panel and the two other wires from this switch. This frees a set of contacts for future use. See fig. 8.

2. Clip the wire coming from relay LR, the normally closed contact, and going to relay HR, the moving contact.

3. Clip the wire on the rear, outer terminal of the out-going switch and run a wire from this contact to the moving contact of relay HR just made available. See fig. 8.

4. Clip the wires on all three lugs of the BR relay. Fold back and disregard.

5. Clip the two green wires from one side of the coil of the BR relay, fold back and disregard.

6. Connect a wire from the N.C. contact on LR, made available earlier, to the coil terminal just made available on the BR relay.

7. On the line transformer located under the gray-colored motor, cut all three wires from the terminals located on the transformer.

8. With a wire, ground the terminal closest to the chassis. A terminal lug is close by to solder to ground. See Fig. 8.

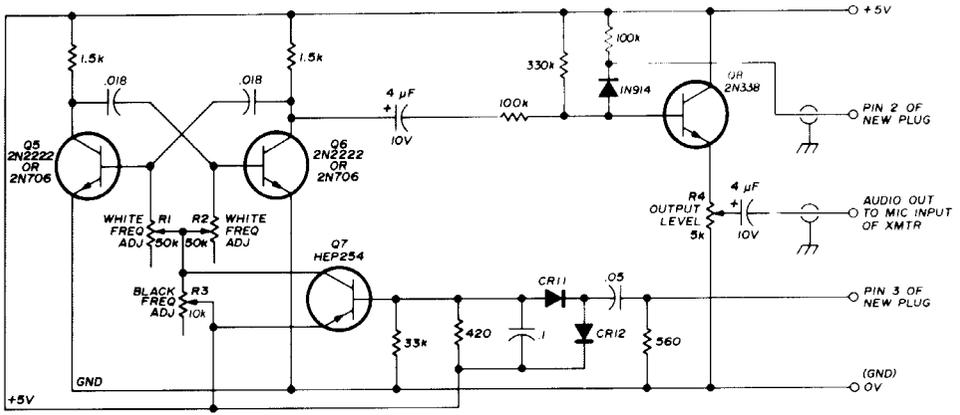


fig. 2. Modulated oscillator.

9. Clip the two wires (one from each lug) from the rear, inner spdt switch of the outgoing pushbutton and fold back the wires. See fig. 8.

10. To the center lug of these three, connect a wire to the last lug made available on the line transformer. No connection is made to the center tap on the line transformer. See fig. 8.

11. Extend the red wire (that was clipped from the line transformer) to the rear-most of the three contacts.

12. Extend the white wire (that was clipped from the line transformer) to the frontmost lug of the three. This transfers the line transformer from receive to transmit with the pushbutton.

13. Clip the two gray wires from the other coil terminal of the BR relay. Solder together and tape.

14. Connect a wire from this coil terminal just made available on relay BR to the N.O. contact, bottom stack of the PWR relay. There are several wires connected to this terminal of the PWR relay.

15. Make available a set of contacts on the TR relay as follows: On the moving contact of one of the sets is a pair of wires, one of which goes to the coil of the HR relay, the other to the 100-ohm, 10-watt resistor bolted to the chassis. After identifying this set of contacts, clip

the pair off the moving contact of the TR relay, solder them together, and tape.

16. Clip the other wires from this set of contacts, fold back and disregard.

17. Clip the wire on the drum phasing contact and remove.

18. Connect a piece of shielded wire from the phasing contact to the moving contact on relay TR just made available.

19. Connect another shielded wire from the N.O. contact of relay TR to the moving contact of relay BR made available earlier.

20. Clip the blue wire on the coil of the ACK relay going to the neon lamp, fold back and disregard.

21. Connect a wire from this terminal on the coil of the ACK relay to the N.O. contact of the PWR relay. (There are several wires on this terminal.)

22. Clip both wires from the acknowledge pushbutton on the front panel.

23. Remove the switch and replace with a spst, normally open pushbutton or similar momentary-contact switch. Wiring of this switch is described later.

24. Clip three wires on one side of the neon lamp holder. One side of the neon has a wire going to the acknowledge pushbutton. If this is the side you clipped first, identify this wire and discard it.

37. Pin 1 - Chassis ground.
- Pin 2 - Shielded lead to N.O. contact of relay BR.
- Pin 3 - Connect a wire to the *red* wire on terminal block for the line transformer.
- Pin 4 - Connect a wire to the *white* wire on terminal block for the line transformer.

Apply a source of 2500-Hz voltage to the 500-ohm input to the limiter. Connect a scope or a VU meter to pin 4 of the new plug (see fig. 5). Put a receiving blank on the drum. Push the INCOMING switch. Adjust L1 for a minimum reading on the scope or VU meter. Push the STOP switch. Put a new receiving blank on the drum. Feed 2000 Hz into the 500-ohm input, and push the INCOMING switch. When the neon light lights, push

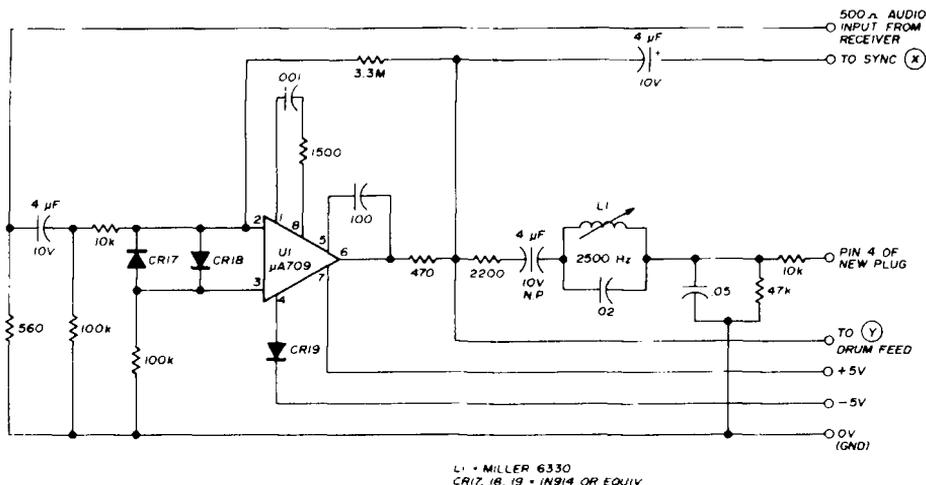


fig. 5. Limiter and video detector.

- Pin 5 - Connect to the N.O. contact, "A" set, of the ACK relay.
- Pin 6 - Connect to moving contacts, "A" set, of ACK relay.
- Pin 7 - Clip wire to service switch (chassis) and extend to Pin 7.
- Pin 8 - Connect a wire to the service switch terminal just made available.
- Pins 9 and 10 - Connect one pin to each side of power transformer primary.
- Pin 11 - Connect a wire to the N.O. contact of the set made available on the TR relay.

tune-up

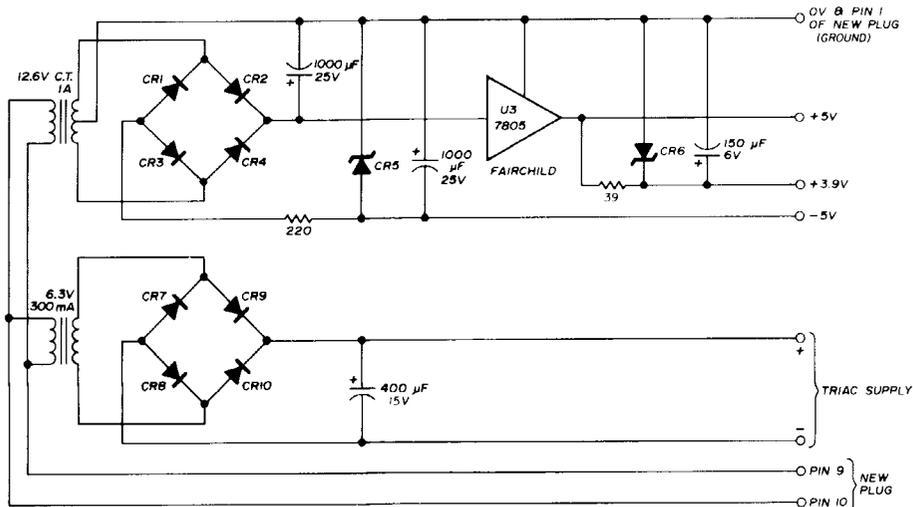
Push the OUTGOING button and check all the voltages on the board: +5V, -5V, and +3.9V. Push the STOP button.

the drum feed pushbutton (the new switch added in the machine conversion).

Adjust P1 on the fax chassis for a good black burn on the paper. Change the oscillator to 2500 Hz. Touch up P1 just so the burn disappears. As you go from 2500 down to 2000 Hz in steps, you will see the gray scale. No further adjustment is required on the video detector. Push the STOP button.

Connect a scope to the audio output of the modulated oscillator (see fig. 2). Connect a frequency counter or some means of determining 2500 and 2000 Hz reasonably accurately. An audio oscillator may be used as a BFO while listening with an earphone connected at the same point. Be as accurate as possible.

Put a sending blank on the drum. With a screwdriver, turn off the service switch



CR1, 2, 3, 4, 7, 8, 9, 10 = 100 PIV, 1A.
 CR5 = 5.1 VOLT, 1 WATT ZENER.
 CR6 = 3.9 VOLT, 1 WATT ZENER.

fig. 7. Power supply.

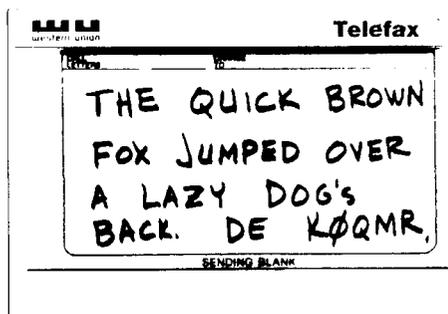
and the trailing lip of the paper located so that the stylus drags over the paper. Both machines are started, the OUTGOING button is pushed on the transmitting machine, and the INCOMING button is pushed on the receiving machine. Several seconds later the neon lamp will light. This indicates the tubes are warm and the machine is ready to use. Wait several seconds after the light is lit to be sure the receiving machine is ready and that sync has occurred. You can tell if the receiving machine is in sync when the hesitation in the gray-colored motor has stopped and the motor runs smoothly. The operator at the transmitting end pushes the drum feed button when he assumes the receiver is in sync (several seconds after the neon lamp has lit). After the button is pushed, the neon lamp will go out, signifying that a picture is being transmitted. The receiving machine will automatically detect the picture and cause the drum to feed. When the picture scan is complete, the machine will automatically stop, and the drum will return to its normal position.

If you push the wrong button, be sure to push the STOP button before pushing any other button. That is, if you push OUTGOING and meant to push INCOMING, push the STOP button before

pushing the desired button. This is to get the logic straight again.

summary

We find it easy to just hook the setups back-to-back for demonstration purposes and testing, or a landline may be used if a radio link is not desired. If wired back to back use two wires and install a T-R switch. Simultaneous connection of output to input causes the drum to feed prior to sync acquisition. If hum level affects sync acquisition, adjust the receiver volume control to a lower setting or install a Butterworth filter between the receiver and the input to the limiter.



An actual transmission. Note the contrast between black and white and note the positive picture which was inverted electronically. The words "sending blank" were transmitted.

The Telefax machine conversion basically consisted of relay rewiring. It is still well suited for its original emission of negative pictures without using any of the new parts on the circuit boards. The basic sending and receiving functions of the machine are left undisturbed. The control system was rewired to cause the drum to feed horizontally with a pushbutton and for sync sending.

The St. Louis Amateur Teleprinter Society (SLATS) is a highly technical

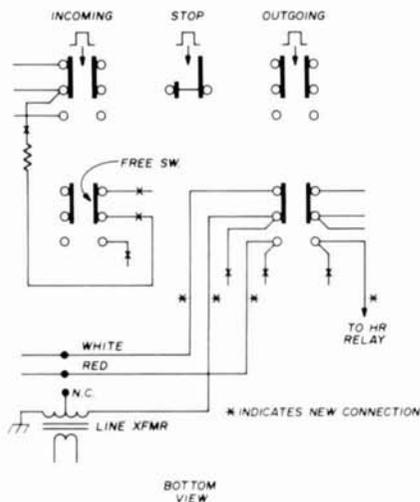


fig. 8. Incoming - outgoing switch modifications.

oriented group that prides itself in projects such as this. We find it quite handy to send schematics, photos, news clippings, etc., along with furthering our knowledge of linear and nonlinear circuits in specialized radio communications.

The author wishes to thank KØDOK, who sparked the development of the project; WAØIDS, who contributed the machine diagram and immense information on the machine; and the SLATS members, who inspired the article and acted as guinea pigs to prove the article could be understood.

A printed circuit board is available from the author for \$6.50 postpaid or a full kit of parts including the circuit board for \$62.00 postpaid in the U.S.A.

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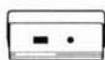
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Howard F. Battie, W7BBX, Falls Church, Virginia 22042

The Ten-Tec Argonaut leaves little to be desired by the QRPP enthusiast. However, in the 5-watt and under area, anything that will improve performance is well worthwhile. This article describes a tailor-made companion for the Argonaut transceiver, but individual modules may be used with transmitter-receiver combinations as desired. Project objectives were:

1. Speech compressor with variable microphone gain and compression.
2. A 2-kHz ssb splatter filter.
3. CW filter, vary narrow passband, no insertion loss.
4. Keyer with self-completing characters, 10-60 wpm.
5. Dc power supply, 12-14 volts.
6. Minimum current drain for portable operation.
7. Detachable key for ssb-only or mobile use.
8. Small size and weight.

the argomate

A review of the current literature showed that no better audio filters could be found than the MFJ Enterprises low-pass and CW units, which are extremely small, rugged, and require only a few milliamps of current. Recent editions of the ARRL Handbook¹ contain an excellent speech processor circuit, which provides up to about 3 dB processing gain. For its cost and simplicity, the IC keyer by W7ZOI² was a natural choice. The finished product is a versatile adjunct to any station, high or low power, fixed or mobile.

A block diagram showing the interface of the two units appears in fig. 1. The ssb and audio filters provide optimum performance while retaining the agc feature on the filtered af signal only.

Minor modifications required to the Argonaut to accept the filter circuits are described; other receivers should be modified similarly after carefully checking the receiver schematic. If audio filters are outboarded from the speaker or phone jack less satisfactory audio filtering and agc characteristics will result. No mods are necessary to the transmitter for the speech processor other than adjusting transmitter drive and mike gain/compres-

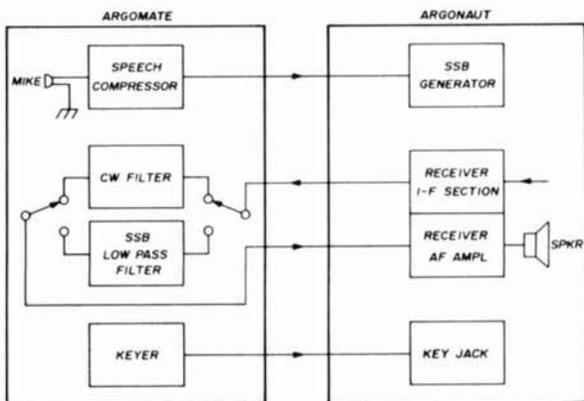
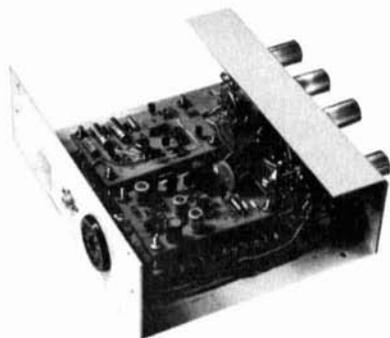


fig. 1. Interface between the Argomate and the Ten-Tec Argonaut transceiver.



Construction of the Argomate. Compressor board is on the left-hand side, keyer board is beside it. CW and lowpass filter boards are mounted under the compressor and keyer boards.

sion levels for optimum clarity and performance.

keyer

The W7ZOI keyer (fig. 2) can be easily constructed on a 2 x 3-inch PC board. It uses a $\mu A747$, which is a pair of $\mu A741$ Cs in a 10-pin TO-5 package. Resistor R2 is used to adjust the relative length of the first two dits to provide even spacing. The dot-dash ratio is determined by C3, C4; C4 is used only for the dot, and both C3 and C4 are used in parallel for the dash.

Q4 collector provides for keying a positive voltage to ground (20V or less). The keying transistor will handle up to 50 mA without a heatsink.

compressor

The speech compressor details, operation, and adjustment are well treated in reference 1. Coil L1 can be either a UTC DO-T8 or UTC ML-6. The circuit is shown in fig. 3.

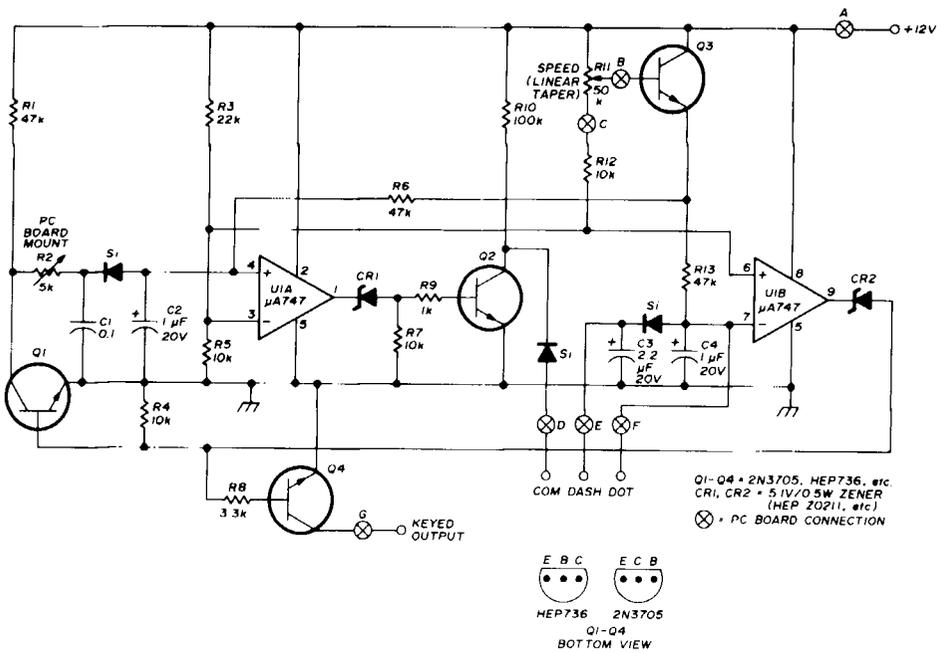


fig. 2. Argomate keyer schematic. Original circuit design, by W7ZO1, is described in the ARRL Handbook (reference 1).

cw filter

The CWF-2 filter is made by MFJ Enterprises and is available in either kit form or pre-wired and tested. Af selectivities of 180, 110 and 80 Hz are provided. This truly remarkable per-

former is a must for any CW operator. In the 80 Hz position, rolloff is 60 dB per octave, virtually eliminating any QRM not zero beat. Insertion gain (up to 2.4) is present in all positions. The high input impedance (680k) and low output impedance (less than 2 ohms) eliminate the

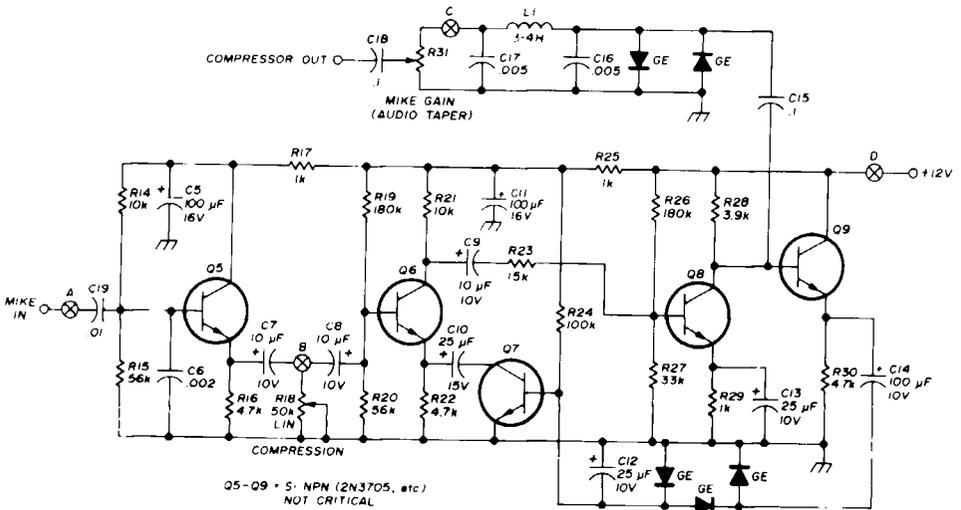


fig. 3. Speech processor.

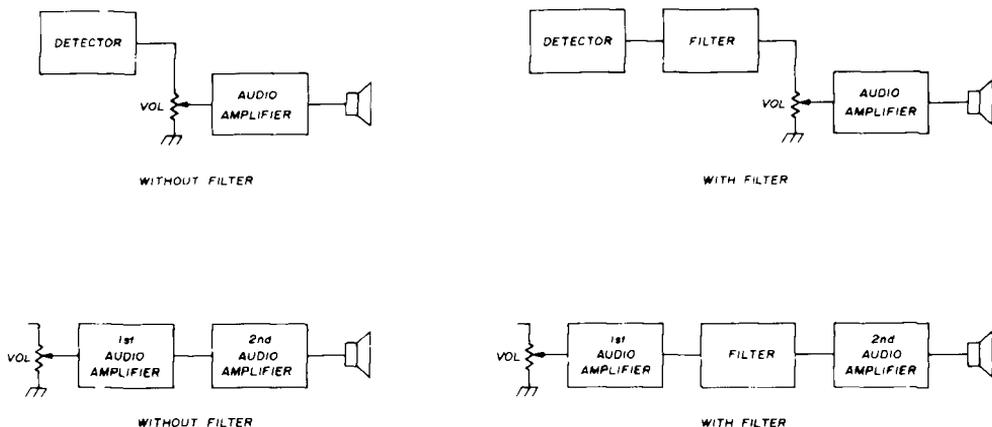


fig. 4. Method of connecting the CW and ssb filters. Better performance results if filters are inserted ahead of the audio amplifier.

necessity for receiver i-f/audio impedance modifications. The unit can be used with a 6-30 volt supply and draws 2-8 mA. The low-Q circuit design eliminates ringing, which is common in most conventional active filters. A 2 x 3-inch PC board is provided by MFJ.

low-pass filter

The LPF-1 low-pass audio filter is similarly constructed on a 2 x 3-inch PC

board. This filter is used only in the receiver audio line to reduce ssb splatter above about 2 kHz. The cutoff frequency can be adjusted as desired by changing the values of eight resistors on the PC board. Received audio intelligibility of a male voice is not impaired as long as the cutoff frequency is above about 1.5 kHz.

audio filter insertion

Although both the CWF-2 and LPF-1

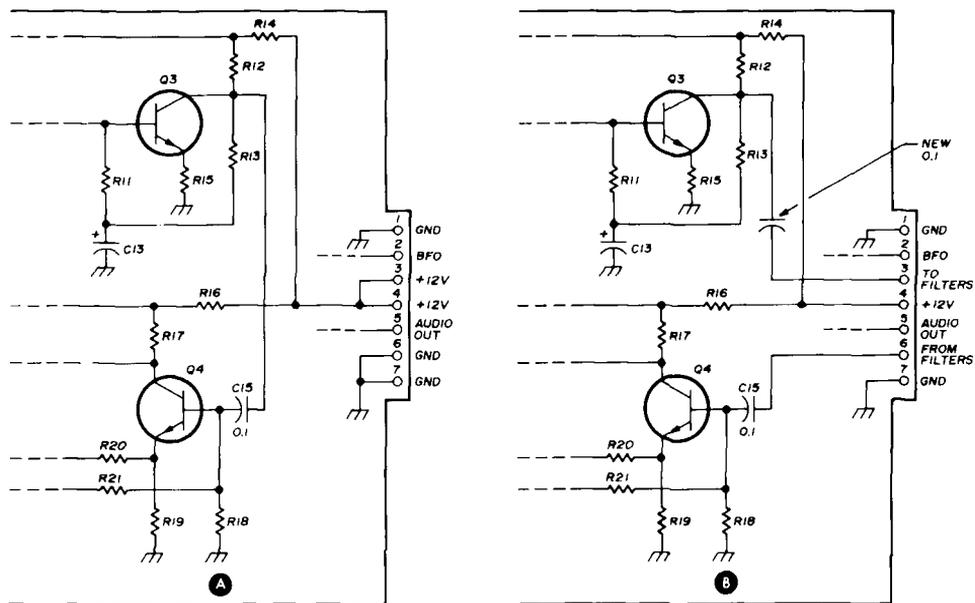
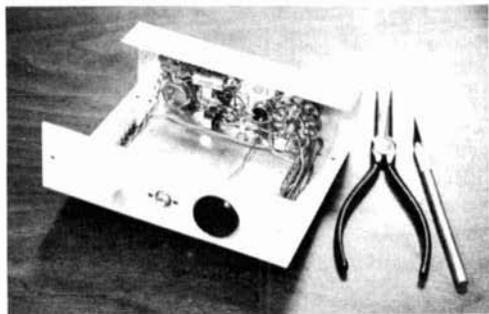


fig. 5. Argonaut i-f board modifications. A and B show before and after modifications. Component numbers are those of the Argonaut.



Front panel switching and control wiring.

filters can be connected directly to the receiver output jack, far better performance will be realized if the filters are inserted prior to the final audio amplifier in the receiver. Fig 4 shows recommended connections.

argonaut i-f board modification

Fig. 5 shows the modifications to the Argonaut i-f board for insertion of the filters. Mini-coax, such as RG-174/U, should be run to the rear panel connector for the filters. To perform the modifications, the copper foil joining pins 3-4 and 6-7 must be broken as well as the foil connection from C15 to the collector of Q3. C15 and the new 0.1 μ F capacitor may be soldered to the top side of the i-f board if necessary. Unsolder the wire connected to pin 6 of the front terminal pin jack for the i-f board and resolder it to pin 7 (ground). The Mini-coax connections are then made to the i-f board

terminals 3 and 6 and run to the rear panel accessory socket.

argomate chassis construction

A Ten-Tec JW-7 cabinet provides a handsome matching enclosure for the Argomate. Additional side shielding of aluminum sheet was added to prevent stray hum and rf pickup.

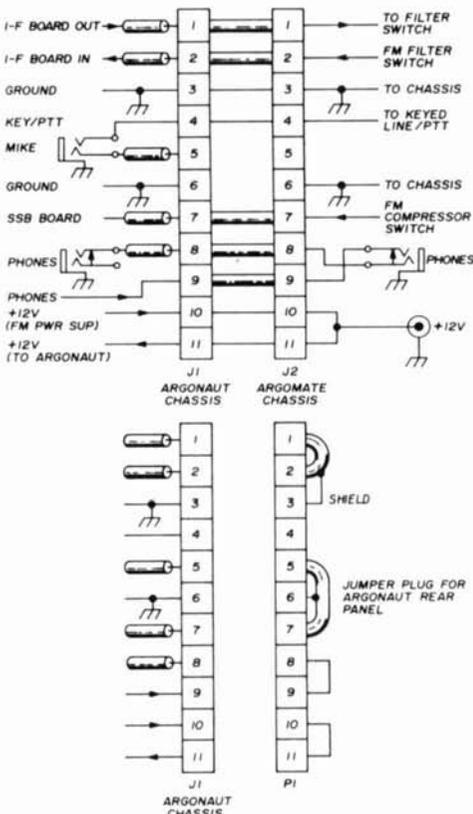
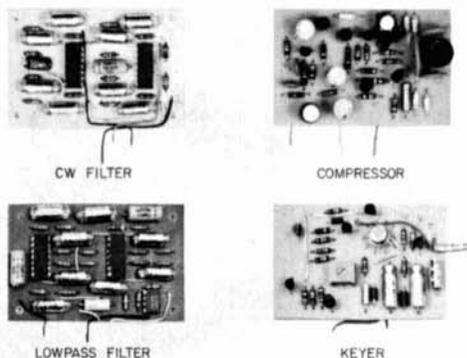


fig. 6. Interconnection wiring between Argomate and Argonaut chassis.

The four 2 x 3-inch PC boards were mounted in two stacks of two boards each in the horizontal position, with the CWF-2 and LPF-1 on the bottom. An 11-pin connector provides cable connections to the Argonaut, including power for the Argomate, as shown in fig. 6. A three-wire connector is provided for the cable to the Ten-Tec KR-1 paddle assembly. The mike PTT and keyer output



Four circuit boards used in the Argomate.

lines are connected together, as well as to the key jack on the Argonaut rear panel. Check the individual keying/PTT circuit before using this arrangement with other units.

The four PC boards are interconnected to multiple-pin terminal strips rather than with direct point-to-point wiring, which makes a neater installation and allows one or more PC boards to be removed if necessary. The terminal lugs serve as handy test points for every PC board input and output. Fig. 7 illustrates the chassis connections. A jumper in the +12-volt leads on TS3 (pins 11 to 12) permits measuring the current drawn by each PC board when a milliammeter is connected between the two pins. Fig. 8 shows switch and interconnection wiring.

The chassis cabinet and additional aluminum sides provide adequate shielding for all wiring and components inside the chassis. However, in areas where ac hum or rf pickup are severe, additional RG-174/U will reduce these effects inside the cabinet. Coax or other shielded connectors must be used between the Argonaut and the Argomate.

An additional 11-pin jumper plug must be used when operating the Argonaut without the Argomate. This plug rewires the i-f/audio connections to the original condition. The Argonaut +12 V supply line is jumpered in the plug and the Argomate rear apron plug. This feature prevents unauthorized operation if no

plug is inserted into the Argonaut but may be eliminated if desired.

To install the 11-pin jack, J1, on the rear panel of the Argonaut, it will be necessary to relocate the adhesive serial number and the existing rf output jack.

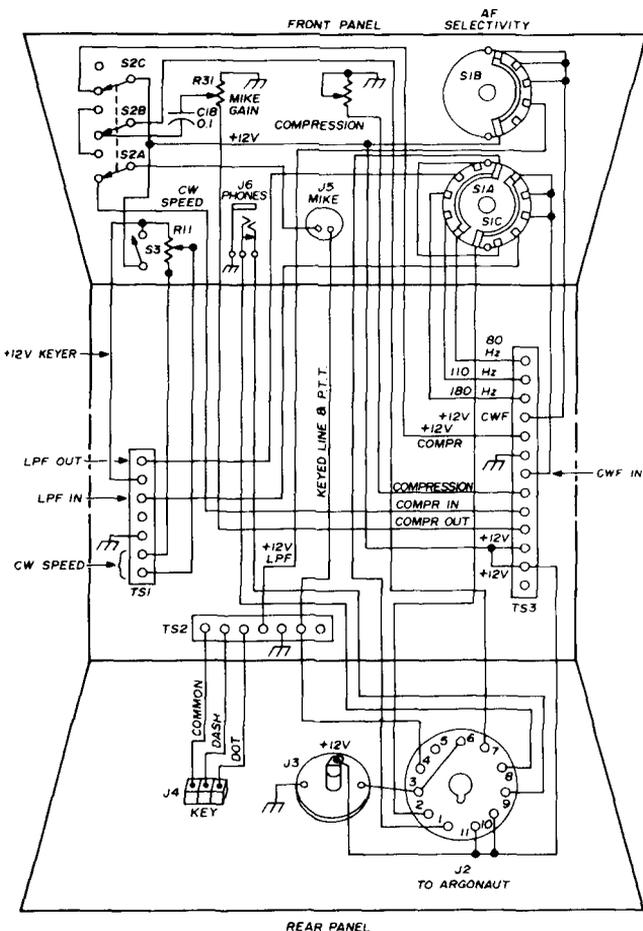


fig. 7. Pictorial representation of keyed chassis connections.

Space is limited, but with judicious placement, sufficient clearance for the relocated rf output jack and J1 is available. A resistor and a capacitor are soldered to the Argonaut microphone jack; relocate these with the mike lead to pin 5 of J1.

operation

For ssb, the mike connector is brought out to the Argomate front panel for

convenience and easier accessibility. The Argonaut mike connector is not used when operating the Argomate; however, the speech compressor may be bypassed on the Argomate front panel by S1. The Argonaut mike jack is reconnected to pin 5 of J1 only. The Argonaut mike PTT line remains connected to the keying line and is additionally connected to pin 4 of

Use of the Argomate will not require modifications to an external linear amplifier as long as the linear can accept the increased duty cycle when operating the ssb speech compressor. The Argomate mike gain and Argonaut ssb drive controls can be easily adjusted for the most effective speech compression and intelligibility.

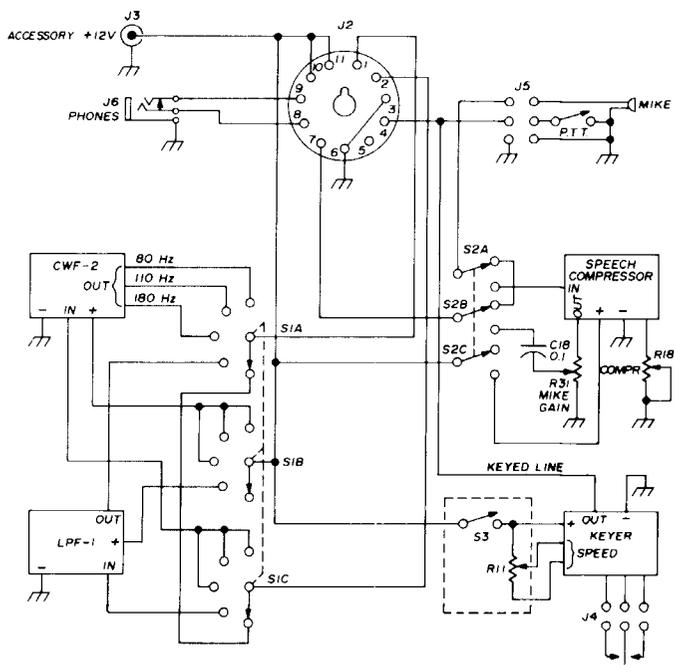


fig. 8. Details of component interconnections.

J1. For ssb operation without the Argomate, the Argonaut mike connector is activated by the jumper wire in plug P1.

On CW, full QSK operation is retained with or without the Argomate. The Argonaut key jack and ssb PTT controls remain activated when the Argomate is connected.

The Argonaut and Argomate phone jacks are connected in series; therefore, phones or an external speaker may be connected directly to either jack, which disables the Argonaut internal speaker. With no phones or speaker connected externally to either the Argonaut or Argomate, the Argonaut internal speaker remains connected.

use with separate transmitter and receiver

For use with a separate transmitter and receiver, the Argomate can easily be adapted with cables to each by addition of a separate multi-connector plug. Since the speech processor, keyer, and audio filters can be controlled independently, only those functions desired need be included in the Argomate. However, the design shown here has the advantage of requiring only a single +12-volt power supply for all functions, while providing the most essential improvements on both transmission and reception for the most discriminating ssb and CW operator.

mobile operation

The Argomate can tolerate supply voltages to about +18 volts; however, if the Argonaut/Argomate combination is used for mobile work, care must be taken to ensure that the alternator or regulator output of the car's electrical system is not higher than the recommended Argonaut supply voltage, +14 V. To provide proper voltage regulation when a voltage dropper is needed, the voltage drop must be independent of the difference between the current drawn on transmit and re-

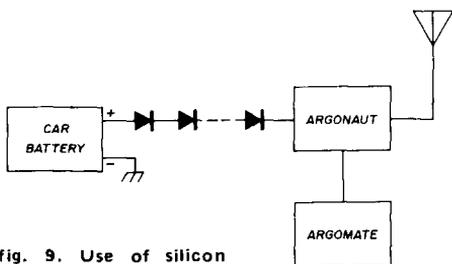


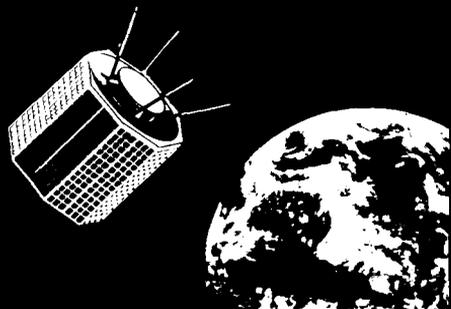
fig. 9. Use of silicon diodes between car battery and Argomate/Argonaut for mobile operation (see text).

ceive. This implies that the voltage dropper must have near-zero ohms internal impedance. This is easily accomplished by using a series of silicon diodes from the car battery (or cigar lighter) to the Argonaut +12 V supply input, as shown in fig. 9. To determine the number of diodes required, measure the battery voltage with the engine running, and figure on 0.6 volt drop across each diode (0.3 if germanium diodes are used). Each diode should be rated for at least three times the maximum current drawn by the complete installation in the transmit mode.

references

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2. Wes Hayward, W7ZO1, "An Integrated Circuit QRP Keyer," *QST*, November, 1971, page 38.
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low-cost monitor receivers for two-meter fm

How to convert
fm broadcast
receivers to
144-MHz
fm service

If you are an avid fm repeater user, you have probably had occasion to wish you could monitor the frequency at your leisure, without having to blow your bankroll on an expensive vhf monitor receiver. This article proposes an alternative: converting an fm broadcast band receiver to the vhf high-band. If you have a little-used a-m/fm portable or even an

old clock radio lying around the house, chances are that it can be converted to two meters. I have performed such surgery on several fm radios with surprising success. An additional benefit is being able to monitor police and public service broadcasts as well. The basic modification is quite simple and straightforward and should be easy to complete on most fm radios.

The superheterodyne circuit used in fm broadcast radios primarily uses two tuned elements to make the receiver tune a particular band. Those elements are the LC tank circuits in the rf amplifier/pre-selector and in the local oscillator. This means that all you must do to change the received frequency is to vary the range of the resonant frequency of the preselector and the local oscillator tanks. This can be done easily if, the radio uses transistors that will operate well at the higher frequency, and the radio has sufficient sensitivity at the higher frequency to enable you to hear the vhf transmissions. It is also very important to have readable schematics, as this helps considerably in making the proper modifications.

A good rule-of-thumb guide for selecting a receiver to modify is to find a recent receiver model (implying better high fre-

James E. Trulove, WB5EMI, 1409 SW 70th, Oklahoma City, Oklahoma

quency transistors) with excellent sensitivity. A good way to judge sensitivity is to retract the antenna all the way — or disconnect it altogether — then try to tune in an fm station normally. If you still receive most stations at full quieting, then the receiver should perform well with the weaker signals present in the public service and amateur bands. However, don't expect the superior sensitivity

converter (which serves the dual purpose of local oscillator and mixer). The rf amplifier is originally designed to tune 88 to 108 MHz, tracking with the local-oscillator tuning. This must be changed to cover the range of 140 to 160 MHz, providing a central frequency of about 150 MHz. You can achieve this by decreasing either the capacitance or the inductance of the rf tank. However,

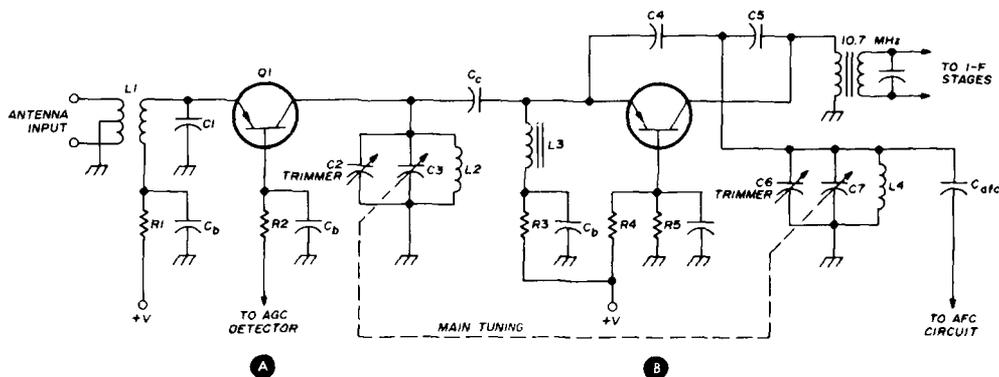


fig. 1. Typical rf amplifier and converter stages found in modern fm broadcast radios. In the rf amplifier (A), L1 and C1 form the low-Q input tuned circuit and L2, C2 and C3 make up the high-Q output circuit. In the converter stage (B), local oscillator tuning is controlled by C6, C7 and L4.

and selectivity of a \$150 crystal controlled, dual-conversion receiver.

the modifications

Most of the fm radios currently in use, use some variation of the rf amplifier and converter circuits shown in figs. 1A and 1B, respectively. These two circuits normally precede the a-m section of the receiver. Bandswitching disables the fm section and changes the role of the first two transistors in the a-m section from i-f amplifiers to an a-m band rf amplifier and converter, respectively. The modifications, then, will be directed at the fm portions of the radio only, leaving the a-m section substantially unchanged.

The primary modification is to the inductances in the rf amplifier output tank and in the oscillator tank of the

reducing capacitance will also have the undesirable effect of reducing the tuning range since the main tuning capacitor is not changed. Thus the inductance must be lowered. This will expand the tuning range somewhat, but will not seriously limit receiver performance.

A quick calculation from the resonant LC formula shows that to increase the resonant frequency by a factor of 1.5 you must decrease the inductance by a factor of roughly 0.5. Of course, this could be done by removing the coil presently installed and replacing it with one of half the inductance. To make matters simple, all you have to do is duplicate the coil and place it in parallel with the present one.

Take a look in the radio you have: the coils should be about three or four turns,

either air wound or on a plastic coil form about 3/16-inch (4.8 mm) in diameter. You may need to do a little visual circuit tracing to find the exact coils involved. The precise dimensions of the new coil are not too critical, as the inductance can be changed quite a bit by compressing or



fig. 2. Layout of the main tuning capacitor. Preselector and local-oscillator trimmers for the fm section are marked on top. An unused connection to the high side of the preselector trimmer is at (A); the corresponding connection for the LO trimmer is at (B). A ground strap for the modifying coils is available at (C).

expanding the turns of wire. Three or so turns of number-20 bare wire was found to work well on various radios.

The same technique is used to raise the local oscillator frequency. It doesn't go up by quite the same factor, so you might want to use a half turn or so less for this coil.

The next thing to decide is where to place the coil. Shown in fig. 2 is the tuning capacitor module commonly found in these a-m/fm radios. It has four sections, two of each rf and oscillator tuning for the a-m and fm sections. The four trimmer capacitors in the top of the case each adjust one section of the main

ganged capacitor. As seen in fig. 1, one side of each capacitor is connected to ground. As both fm trimmers have connections appearing on the top of the case, this is probably the easiest place to mount the two new inductors. Placing them across the trimmers to ground amounts to paralleling them with the original inductors. The pictorial in fig. 3 shows the placement of the new inductors. Note that they are placed so that the axes are 90° apart. This minimizes coupling between the two stages.

One word of caution: some receivers are designed so that the unused lead of the trimmer capacitor is not at dc ground. If you have such a receiver, install a 0.001 μ F ceramic disk dc blocking capacitor in series with the new coil to prevent damage to the radio. Better yet, choose a capacitor with leads longer than 2½ inches (64 mm) and form the coil from one of the two leads.

alignment

Rough alignment is really fairly simple, and it will probably be quite sufficient if you only wish to tune one small band of frequencies, such as 146 to 148 MHz. First you need a fairly strong signal at the frequency you are trying to tune. A dummy-loaded transceiver should suffice for setting up the two-meter band, as most dummy loads have a little rf leakage. If this method is not convenient, you may be able to use transmissions on radiotelephone at about 150 MHz, or even your local repeater.

With the signal source turned on, adjust the main tuning dial to a convenient location near the middle of its tuning range. Then, with a small screwdriver, carefully turn the local oscillator trimmer until the signal source is well tuned in. If you can't find the signal, try adjusting the inductor by compressing or expanding the windings. Then repeat the above procedure again. You may need to rewind the coil if you still have difficulty. If you have a variable signal generator or a grid-dip oscillator, use it to find where you are tuned. Remember that you

haven't adjusted the preselector yet, so you should be receiving two frequencies: the desired one and its image. The local oscillator will be between these two frequencies.

Assuming, then, this alignment has been accomplished, the preselector must be set up. This is easy. Leaving the signal source on and the radio tuned in to it, adjust the preselector trimmer for maximum signal. Since the tuning of this tank circuit is coupled through the converter transistor to the local oscillator tank, there will be some interaction, so go back and retouch the oscillator trimmer. Repeat this procedure a few times. If you have difficulty with the preselector, use the hints given in the previous oscillator alignment. A grease pencil may be used to make reference marks on the receiver tuning indicator for the new band.

This takes care of the initial alignment. You should now be able to tune in stations well on the band of frequencies around the one where the alignment was performed. But proceed with caution; the following "fine tuning" adjustments are much more difficult to make and may well require some specialized equipment. If you are satisfied with the sensitivity, selectivity and tuning range of your converted radio, you are finished with alignment. For the more adventurous souls, or the perfectionist, read on!

input tuned circuit

The antenna input is normally a balun transformer with a tuned secondary as shown in fig. 1. The Q of this tank circuit is low, broadbanding its response, and it primarily provides protection from frequencies considerably removed from the fm band. This circuit can best be retuned by removing the capacitor and substituting one of half its value.

tracking

If you wish to tune the entire public service band, proper preselector tracking needs to be taken into account. Tracking refers to the fact that for proper tuning the preselector must be tuned exactly

10.7 MHz away from the local oscillator frequency. When originally manufactured, your fm radio was adjusted to do just that.

Normally, the receiver is designed to have three points of perfect tracking, shown as zero tuning error in fig. 4.

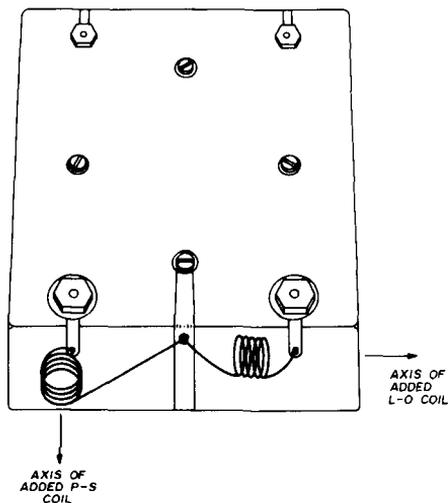


fig. 3. Mounting the new coils. The coils should be soldered between the unused trimmer tabs and the ground strap with their axis at right angles.

Tracking varies over the tuning range because of the interaction of the two steps.

Since several of the component values have been changed (as well as the operating frequency), you can presume that the preselector no longer tracks the received frequency. The solution is to adjust the value of the preselector inductance, then readjust the preselector trimmer capacitor until the receiver more or less tracks in the range of interest.

Pick two points on the dial to adjust the tracking, one at the high end and one at the low end. If you have access to a signal generator, any alignment points will do. If not, set the tracking at two discrete frequencies to which you wish to listen. Alternately adjust the preselector inductance (by compressing or expanding the coil turns) and the preselector trim-

mer capacitor at both high and low ends of the dial until the preselector stage tunes properly at both points.

i-f selectivity

As the commercial fm broadcaster uses 75-kHz deviation, broadcast receivers use an i-f bandwidth of about 150 to 200 kHz. Likewise, the detector is broadband. There are two methods of achieving this broadband response in the i-f stages: stagger tuning and resistive loading of the tuned circuit.

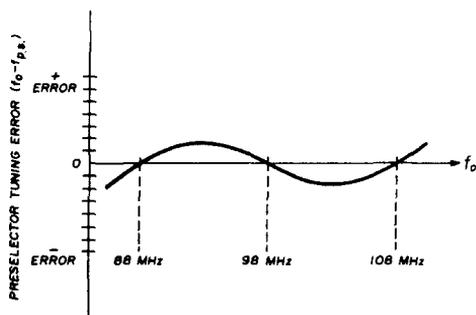


fig. 4. Points of perfect tracking, showing the three points of zero tuning error, where the tuned frequency (determined by the sum of the 10.7-MHz i-f plus the oscillator frequency) equals the preselector center frequency.

In stagger tuning, stages are purposely misaligned to yield an overall wide pass-band. This technique naturally decreases the total gain of the i-f system. Here the selectivity can be improved by tuning the i-f stages to the same frequency while receiving a fairly weak station or other fm-modulated signal source. Take care that the increased gain produced by your retuning does not cause oscillation, even if it means detuning one or more stages.

With resistive loading, a resistor is normally placed in parallel with the inductor, although sometimes the effective resistance may be in the coil itself in the form of small wire size and a large number of turns, or in the loading effect of the transistor. The only modification that is practical here is to clip out the discrete loading resistor. This should be

done only if the resistor is not part of the transistor bias circuit.

Frequently a combination of stagger tuning and resistive loading is used. It is likely that retuning the stages will be sufficient.

automatic frequency control

For some reason most fm radios are equipped with afc. This was great in the days of thermionic emission devices (tubes) where the source of instability was extremes of temperature as the filaments heated up. In transistor equipment, the afc circuit is more of an advertising point, and frequently introduces hysteresis, making tuning more difficult. While this is no real disadvantage for 200-kHz wide fm broadcasts, it is definitely harmful when trying to tune intermittent transmissions only 10- to 30-kHz wide. In the first radio that I modified, a poorly designed afc system actually pulled the local oscillator off frequency after each transmission began! The afc will also pull the receiver off a desired transmission onto a strong undesired one if it is within its range.

To disable this circuit, simply cut out the afc coupling capacitor, C_{afc} , shown in fig. 1B. This capacitor couples the capacitance of the afc transistor, which varies with the magnitude of feedback, from the fm detector circuit. Clipping it out will only affect the tuning of the local oscillator tank and will not affect any dc components in the afc or detector circuit. The local oscillator should normally have sufficient stability to preclude serious drifting.

conclusion

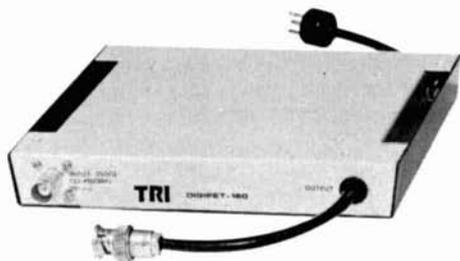
It is hoped that this article has provided a simple way of monitoring two-meter transmissions. Converting a radio for a friend is a great way to give him a taste of two-meter operating. Performance should be adequate for casual monitoring, though selectivity and sensitivity will not be as ideal as in an expensive transceiver.

ham radio

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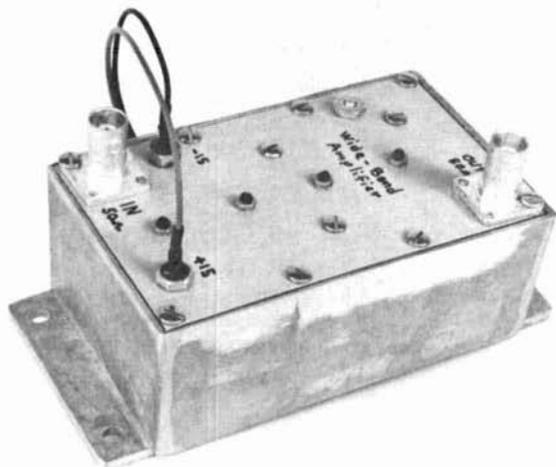
However, if you operate up around 140 MHz, you'll want the Digipet 160 converter. It costs an additional \$50 and, mated-up with the Digipet 60, measures the critical range from 130 MHz to 160 MHz. Its AC or DC operable with complete overload protection, plus being stable (aging rate: 1 part in 10^6 /week), small (7" deep x $2\frac{1}{2}$ " high), sensitive (50 mV/m's), flexible (five numerical-tube digits) and accurate (resolves to 1 kHz or 1 Hz, depending on gate time selected).

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Broadband rf amplifiers are becoming quite common nowadays with the availability of ferrites and transistors with high gain-bandwidth products. The current fad is to spend all your design time on the ferrite transformers so that the 50-ohm output impedance can be transformed to some higher impedance that may serve as a reasonable collector or

drain load for the transistor or fet you use. Such transformer design is beyond the ken of most hams, even if they could get the required ferrite cores. The general principles of broadband transformer design are described in two rather good articles.^{1,2} Beyond these basic articles are a host of little practical tricks and facts, many of which are not available in text at all. One transformer manufacturer has gone to such lengths as to put a sheet of lead inside the epoxy case of his latest rf transformer when it was on loan to a large systems company so that the potential customer couldn't x-ray it to see how it was constructed (before quantity ordering).

distributed amplifier

The broadband amplifier described here does not use any special ferrite transformers, and uses plastic economy-type transistors. The construction techniques are easily within the ham-experimenter's ability, and the finished amplifier has rather good performance for its cost.

An old, neglected technique of broadband amplifier construction was selected

Hank Olson, W6GXN, Post Office Box 339, Menlo Park, California 94025

because it requires no special parts. The distributed amplifier is the name by which this amplifier goes, and it dates back to 1937. The distributed amplifier was originally designed to be used with vacuum tubes, so that the input capacitance of each grid could be lumped with the shunt C of an artificial transmission line on the input side of the amplifier. In a like manner, the output capacitance of each anode was lumped with the shunt C

be used in the distributed amplifier. The base-emitter junction is used *in series to ground* with the shunt C of the artificial transmission line. In this way it is the *current through* the shunt C of the transmission line that drives the base of the transistor — which is just what the bipolar transistor requires. Reference 4, which is now over ten years old, describes a practical amplifier of this type as well as the basic design equations.

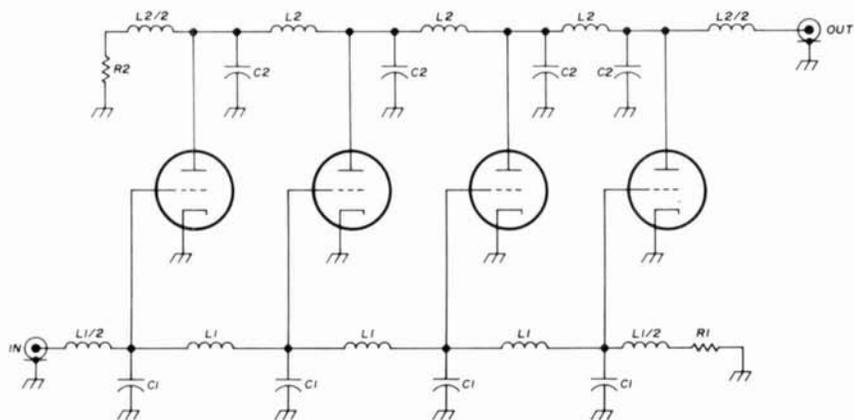


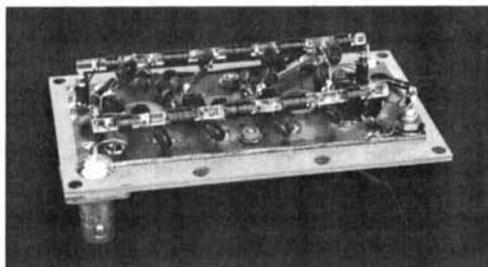
fig. 1. Basic distributed amplifier.

of an output transmission line. The basic distributed amplifier (with tubes) is shown in fig. 1. The gain contribution of each tube, in phase with the amplified wave as it passes down the artificial transmission line, adds to the contributions of the other tubes. Therefore, the whole distributed amplifier acts as if the transconductances of the tubes were all in parallel; i.e., the stage gains add.

bipolar transistor amplifier

You can easily see how the principles for a vacuum-tube distributed amplifier could be almost directly applied to one using fets. This has been done, in fact, in a technical article that will soon be published by Siliconix, one fet manufacturer.³ It is a bit less obvious, however, to see how the principle of the distributed amplifier can be applied using bipolar transistors. Fig. 2 shows the method by which bipolar transistors can

The broadband amplifier shown in fig. 3 was built using the method described in reference 4. Since ± 15 volts is a rather common type of power supply nowadays (because of the wide use of operational amplifiers), the amplifier was designed to operate on this source of power. The transistor type chosen for the amplifier was the Motorola MPS-U05, a silicon npn type in a small plastic power package.



Layout of the distributed amplifier showing the two delay lines. Input is to left, output to the right.

This transistor is truly a marvel of power handling and gain-bandwidth product, considering its cost is in the one-dollar vicinity. Fig. 4 shows how the gain bandwidth product and beta vary with current of the MPS-U05.

transistor, mica washer, aluminum washer and aluminum plate clamping. Silicone grease was used in all these thermal interfaces to increase heat conduction.

The inductors that make up the various L values in the artificial transmission

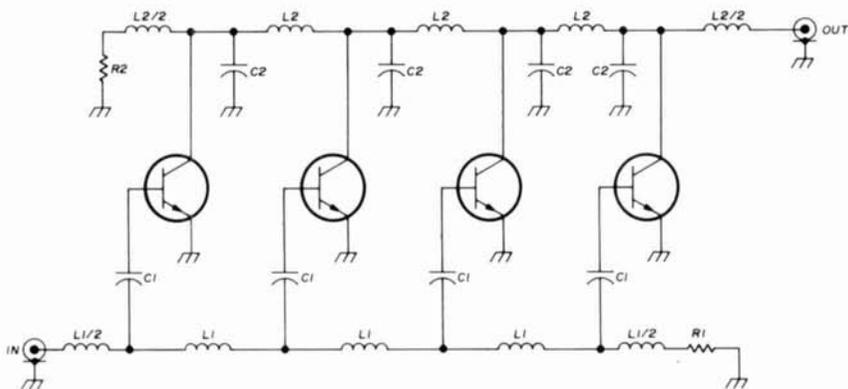


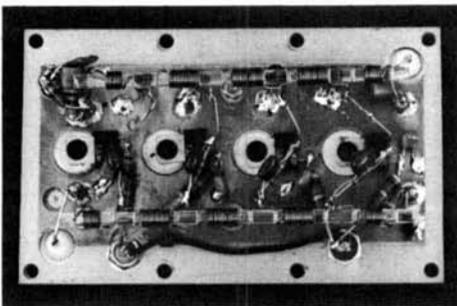
fig. 2. Distributed amplifier using bipolar transistors.

construction

When constructing the distributed amplifier, note that the four transistors must have their collector tabs installed in a heat sink. In this design the heat sinks are 1/2-inch (127 mm) diameter aluminum washers set into holes in the piece of copper laminate on which the circuit is built. The tabs of the transistors actually are mica insulated (electrically) from the aluminum washers, and the washers conduct the heat from the transistors to the aluminum outer plate of the box chassis. Nylon 4-40 screws were used for the

lines were wound on two 1/8-inch (3.2 mm) diameter lucite rods (one for the input 50-ohm line and one for the output 50-ohm line). The dimensions of these multicoil structures are not at all critical. In fact, the first version of this amplifier was built using ten standard-value rf chokes and a handful of standoff terminals. The single rod, hand-wound coil structures were used simply to conserve space and save costs.

The 100-pF capacitors in the two transmission lines should be silver-mica types. The DM5 version of silver mica is smaller and easier to use here, but larger types should work. The emitter-bypass and output-coupling capacitors should be low-inductance ceramic types. The Red-cap 0.1- μ F capacitors by Erie are just fine. It would be better to use 0.01- μ F ceramic disc capacitors (and have the low frequency end suffer a bit) than to use 0.1- μ F capacitors of foil construction (mylars, etc.) if the low-inductance 0.1- μ F types cannot be readily obtained. Do not parallel capacitors for emitter bypassing. Such practice seems like a good idea, but usually results in a vhf parallel-tuned circuit, which causes the amplifier to take off.



Plan view of the distributed amplifier showing the four transistor stages and home-made delay lines. Input is at right, output at left in this photograph.

There are three 1000-pF standoff or feedthrough-type capacitors in the amplifier. These are used in noncritical points in the circuit for getting into the box chassis or where a tie point is needed. The 1000-pF standoff type is located at the

should be tested. The amplifier should draw about 120 to 150 mA with ± 15 volts applied to it. The gain is about 18 dB, flat to within 3 dB from 1 to 36 MHz. The noise figure was measured only at 30 MHz with an A.I.L. automatic noise

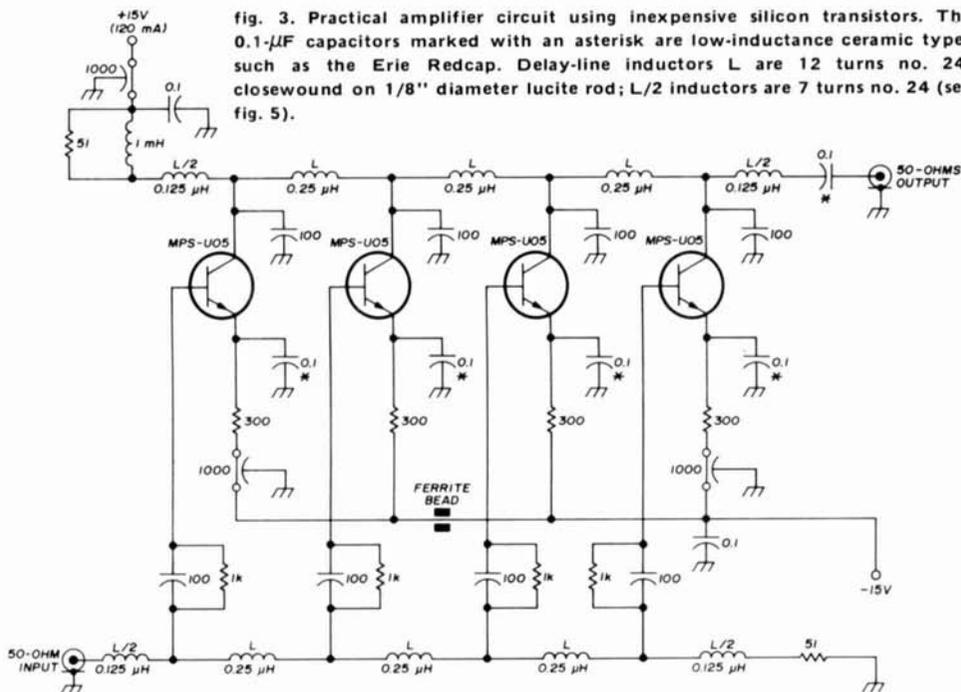


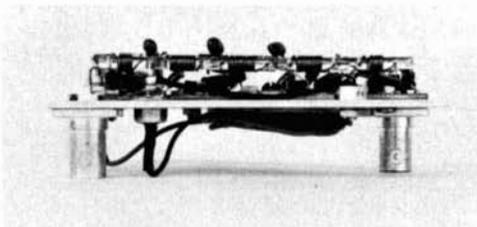
fig. 3. Practical amplifier circuit using inexpensive silicon transistors. The 0.1- μ F capacitors marked with an asterisk are low-inductance ceramic types such as the Erie Redcap. Delay-line inductors L are 12 turns no. 24, closewound on 1/8" diameter lucite rod; L/2 inductors are 7 turns no. 24 (see fig. 5).

junction of the two 300-ohm emitter resistors of the first two stages. A 1000-pF feedthrough type is located at the junction of the two 300-ohm emitter resistors of the last two stages and serves as the -15 volt connection to the outside of the box chassis. Between these two 1000-pF capacitors is a piece of hookup wire with a ferrite bead on it to damp out a possible vhf parasitic resonant circuit. These ferrite beads are in common amateur use now and are available from Amidon Associates.*

testing

With the input and output terminated in 50-ohm test equipment, the amplifier

figure meter; it was 8 dB. The amplifier was capable of putting out +20 dBm, or 100 milliwatts before a compression of 1 dB was encountered. In fact, by increasing the 300-ohm emitter resistors to 1 watt and using ± 30 volt supplies and higher-voltage MPS-U06 transistors, it was possible to get 1 watt output before 1 dB compression set in.



Side view of the distributed wideband amplifier showing position of delay lines.

*Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607.

applications

Assuming a careful job of construction and testing as described above, you now have a broadband hf amplifier. What can be done with it? The first thing that may come to mind is its use as a preamp for hf receivers from 160 through 10 meters.

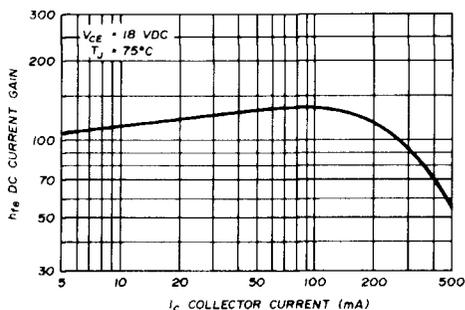
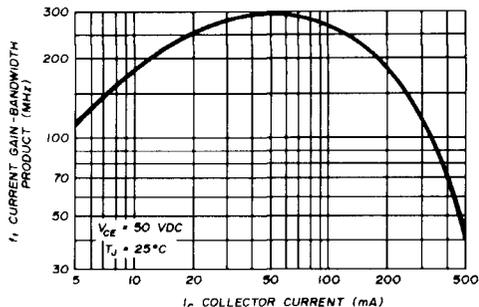


fig. 4. Dc current gain and gain-bandwidth product of the Motorola MPS-UO5 transistor used in the wide band amplifier.



However, putting an 18-dB preamp ahead of most hf receivers will usually cause more problems than it will solve. The extra 18 dB will *reduce* the dynamic range of the total receiving system by nearly 18 dB, unless the receiver has an incredibly bad noise figure. (I have occasionally run into hf receivers with noise figures as high as 15 dB.) The preamp will appear to "add distortion," whereas we know from measurements that it won't

In building up systems of available circuit blocks, the 50-ohm input and output, 18-dB gain block is quite handy. It can be used to deliver the +7 dBm local oscillator input required to drive the L port of a double-balanced hot-carrier diode mixer, for instance; or it can be used to amplify the signal going into the R port or out of the I port of such a mixer.

There are many other uses to which the broadband amplifier is uniquely suited, which involve swept frequency reception and transmission. These techniques are not in general legal for hams to use on the air, but can be useful in test and measurements on the bench.

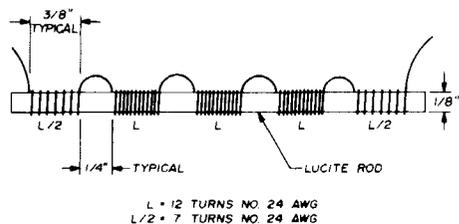


fig. 5. Construction details of the delay-line inductors.

go into nonlinear operation until a signal of over zero dBm is present at its input. The apparent distortion in the preamp is caused by the gain of the preamp increasing signal levels further down the amplification chain *in the receiver*.

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nonresonant antenna impedance measurements

In an earlier article¹ I indicated that a big problem in antenna design is how to measure the impedance of a nonresonant vertical antenna without the use of an expensive rf bridge. Here are some methods using inexpensive components.

There are two parts to the problem: resistance and reactance. Solving for reactance is by far the easiest. All that's necessary is a variable capacitor or variable inductor and a grid-dip meter. Simply tune the grid-dip meter to the desired frequency (check with a calibrated receiver), link it to the antenna with a turn or two, put in the necessary resonating variable device, and adjust it for dip in the meter.

capacitance method

Using the 40-foot (12.2 meter) vertical irrigation pipe described in my previous article, connect a variable capacitor in series with a turn of wire around the grid-dip meter coil to ground (fig. 1). Set dipper to 7.25 MHz and adjust the capacitor for maximum dip. Measure the capacitor with an accurate capacitor checker^{2,3} and solve for $X_C = 1/2 \pi fC$.

inductance method

Using the same antenna for 20 or 75 meters calls for a variable inductor. Assorted roller inductors are available on the surplus market, or you can wind a coil approximately the right size on a slug-tuned form and vary the slug. Or you can wind a coil with taps and tune the dipper until you find a tap that resonates on the low side of the desired frequency. Then start clipping turns, a fraction of a turn at a time, until the dip is at exactly the right frequency. Having achieved a

dip at the proper frequency, the inductance and/or inductive reactance can be determined.

Inductance-measuring devices are a little less common than capacitor bridges. One method is to connect the inductor across a known capacitor and grid dip the circuit. At the dip frequency

$$X_C = X_L = 1/2 \pi fC$$

At the antenna design frequency

$$X_L = X_L \text{ above } \times \frac{\text{ant freq}}{\text{dip freq}}$$

We now have the reactance, X_L or X_C , whichever is required. We will now leave this resonating element in the circuit and measure the resultant R.

antenna resistance

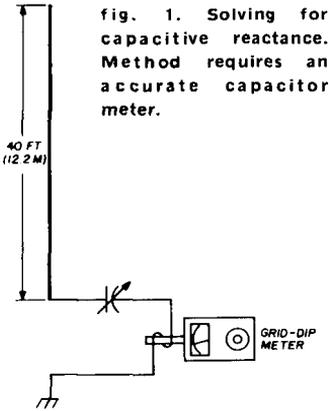
Measuring the resistance of the antenna to a fair approximation without a bridge presents somewhat of a problem. One method within the means of most hams is as follows: We need a voltage-indicating device (not necessarily calibrated), a small exciter, and handful of carbon resistors in the expected neighborhood of the antenna resistance, and last but not least either a thermomilliammeter or some kind of an rf voltmeter. Theoretically, if we maintain constant voltage output from the exciter and add a resistance equal to the antenna resistance in series with the circuit, the current will be one-half, or the voltage across the antenna portion will be one-half.

If you have a thermo-milliammeter, the procedure is simple. (Many thermo-milliammeters were available on surplus a few years ago.) Assuming you have a voltmeter that will read *something* on rf,

Bob Baird, W7CSD, 3740 Summers Lane, Klamath Falls, Oregon 97601

or a scope, feed the antenna system and note the position of the meter needle or amplitude of the scope display. Note the thermo-milliammeter reading. Now insert resistance in series (always keeping the voltage constant) until the current decreases to one-half. The inserted R now equals antenna R .

In the absence of a thermo-milliammeter, use an rf voltmeter across the



antenna. Again, keeping exciter voltage constant, add resistance until the rf voltage across the antenna is one-half. ($R_{ant} = R_{added}$.) We used the field-strength section of an swr meter for our rf voltmeter. We unscrewed the telescoping antenna and connected a 20k resistor in its place then connected the assembly across the antenna to ground. We adjusted for full-scale reading with no added resistance and used a vtm across the exciter. Holding $V_{exciter}$ constant, we added resistance until the meter across the antenna-to-ground circuit read half-scale. In the case of the 40-meter antenna, we came out with a 100-ohm resistor in series with a 33-ohm resistor. The previous bridge measurement was 142 ohms. And that's about as close as you are going to get. (Actual measurement of the two resistors was 131 ohms.) A dc milliammeter, resistor, capacitor, and rf diode will work in either or both positions of the above (see fig. 2).

The accuracy of this kind of measurement hinges on the accuracy of the

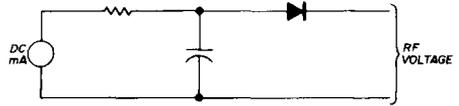


fig. 2. Simple circuit for measuring rf voltage.

instrument used for making the "half" measurement and the added series resistors. Ours measured 131 ohms on dc. Carbon resistors might be slightly higher on 7 MHz. Accuracy of the exciter voltmeter is not important, as it merely has to be kept constant. Fig. 3 shows the setup.

This measurement method will give you a ballpark estimate near enough so that applying the results to the impedance-matching data in reference 1 will give unity swr with a bit of tweaking.

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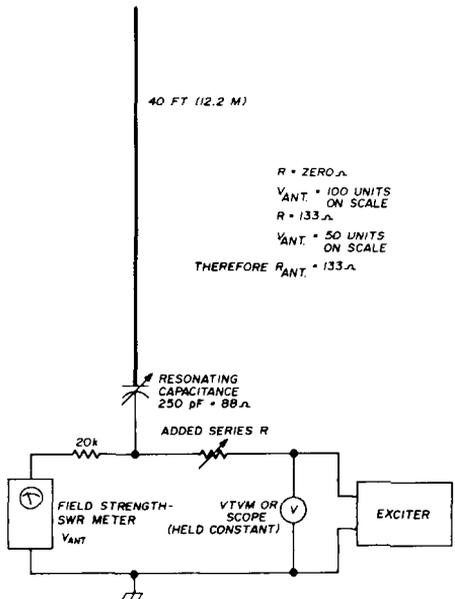


fig. 3. Test equipment for determining antenna resistance (40-meter example).



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vertical antenna radiation patterns

The effect of height on vertical antenna radiation patterns

Various antenna articles have shown that it is possible to use almost any height vertical, from very short ones to ones which are relatively tall. For our purposes this range of heights includes electrical lengths (h/λ) from an h/λ of about 0.1 to 0.3. These ratios correspond to antenna heights from about 30 to 80 feet (9.1 to 24.4 meters) at 3.8 MHz. In this article I will compare the vertical radiation pattern of verticals in this height range. For purposes of this discussion I will assume that the *horizontal* radiation pattern is omni-directional and of no further interest.

Some antenna articles have indicated that the low-angle vertical radiation pattern for tall vertical antennas is much better than for short ones. This is a situation that I'll examine in this article, and will show that on the basis of

comparison usually used, the low-angle vertical radiation patterns for both short and tall vertical antennas are essentially identical. For antennas taller than 0.3λ there is some improvement at certain radiation angles but also degradation in field strength at other radiation angles. Data on this will also be provided.

The vertical radiation angles of interest will be in the range from just above the horizontal (zero degree) through angles of up to 30 degrees. These are the radiation angles of interest for DX work on the 80-meter band. Although the far distant pattern will have a drop in signal strength near zero degree because of ground losses, primarily beyond the antenna radial system, the vertical radiation pattern of signals launched *near* the antenna and radial system is the topic of discussion in this article.

To compare the vertical radiation patterns of verticals of various heights, it is necessary to calculate the signal strength for each antenna at a fixed distance from each antenna at several radiation angles between zero and 30 degrees. It is also necessary at the same time to take into account various other factors, some of which are the same for each antenna, and some which vary with antenna height.

Equation 11-95 from page 314 of John Kraus's book, *Antennas*¹ includes all of the factors necessary to carry out the task just outlined, and his equation applies to the situations being considered here. The equation, with slight changes in notation for clarity, is

Robert E. Leo, W7LR, Electronics Research Laboratory, Montana State University

$$E(a, r) = \frac{60}{r} \sqrt{\frac{W}{R_{loop} + R_{ground}}}$$

$$\frac{\cos(\beta h \sin a) - \cos \beta h}{\cos a} \text{ volts/meter}$$

This equation gives the vertical radiation pattern for a vertical of height h , at an elevation (radiation) angle a , at a distance r meters, for a power W watts, for resistances R_{loop} and R_{ground} ohms. This situation and some of these terms are illustrated in fig. 1. Not shown is the antenna current distribution, a current element, or other items used to derive the equation for E , the field strength in volts per meter.

Now consider the various terms in the equation. The number 60 is a constant, so will be the same for all of the calculations. Since I want to calculate the field strength, E , at some arbitrarily fixed distance, I'll let $r = 1$ mile, which is $(5280/3.28) = 1610$ meters, so this term is also fixed at this value. Antenna books often use a distance of 1 mile and a power output of 1000 watts as a basis for comparison of antennas.

The term W is the power in watts going into the antenna radiation resistance and into the ground resistance. I'll use $W = 600$ watts to be representative of the amateur situation with a power input of 1000 watts. W is the transmitter power output less any losses in your antenna tuner, in the feedline between tuner and antenna matching network, or in the antenna matching network. For this article, I will neglect network losses and ground losses. Their influence on field strength and the radiation pattern will be considered in a later article.

There are two resistance terms involved, R_{loop} and R_{ground} . R_{loop} can be obtained from my previous article² by means of the equation

$$R_{loop} = R_{base} \times \sin^2 \beta h$$

The resistances R_{base} and R_{loop} are small values for short antennas, are both equal to 36.56 ohms for $h/\lambda = 0.25$, and are larger values for $h/\lambda = 0.3$ or greater.

R_{ground} is the earth ground loss resistance. The ground losses are more important for short antennas where R_{loop} is small.

For example, if $R_{loop} = R_{ground}$, then half of the power, W , will go to the

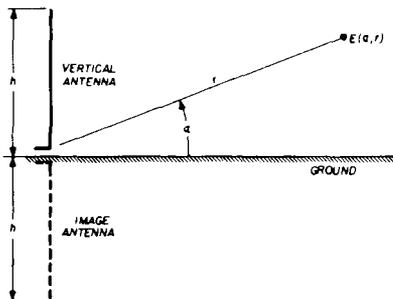


fig. 1. Vertical antenna radiation. For a discussion of the terms and their use in the field-strength formula, see the text.

antenna to be radiated as a useful signal, and half will be used to heat up the ground which, of course, is not our objective and field strength will be affected accordingly. The last term in the equation is

$$\frac{\cos(\beta h \sin a) - \cos \beta h}{\cos a}$$

This term depends on $\beta h = (2\pi/\lambda)h$, and a , the vertical radiation angle. For a given frequency, λ is constant, so βh depends directly on the antenna height, h .

This term is plotted in some antenna articles and is then used as the basis to claim that tall verticals are much better than short verticals, since this portion of the equation does, by itself, favor tall verticals as far as low-angle radiation is concerned. It is, however, not realistic nor correct to consider only this term and to neglect the other terms discussed earlier. Now consider why using only the last term is incorrect.

In more simplified but correct terms the Kraus equation is

$$E \propto \sqrt{\frac{W}{R}} f(h, a)$$

The \propto right after the E says that E is

proportional to both $\sqrt{W/R}$ and to $f(h, a)$. Here $R = R_{100p}$

$$\frac{W}{R} = \sqrt{\frac{I^2 R}{R}} = I \text{ or } E \propto I f(h, a)$$

This form for E , being proportional to the current, I , and $f(h, a)$, is often given in antenna books. If only $f(h, a)$ is plotted, it assumes a constant current, I . But since power, W , is constant, and the resistance, R , is smaller for short antennas, a constant current is not consistent with the equation $W = I^2 R$ for our situation. Perhaps a better way of saying this is that to use $E \propto f(h, a)$ only assumes a constant current, rather than a constant power. For the amateur situation, a constant power is a much more realistic assumption.

For a constant power, $\sqrt{W/R}$ goes up as h/λ becomes smaller since R becomes smaller for short antennas. At the same time $f(h, a)$ becomes smaller, and the two terms tend to offset each other, which is why E does not change much when constant power is assumed.

A simple non-mathematical argument about why E is essentially the same in both cases would be worthwhile. For a short antenna, E at a distance is due to the contribution of a few current elements, each with high current, while the signal from a tall antenna is produced by many current elements, each with lower current.

$E(a, r)$ was calculated in a few minutes time on an HP-35 pocket electronic calculator for h/λ of 0.1, 0.2 and 0.3, and for $a = 0^\circ, 10^\circ, 20^\circ$ and 30° . The results are given in fig. 2. This graph shows that for constant power the height of radiator makes very little differ-

ence in the vertical radiation pattern. This is the essence of the entire article.

The results here are in agreement with those in most antenna books and articles, such as in Kraus, *Antennas*, page 317, figure 11-36,¹ or in Harmon, *Proceedings of the IRE*, January, 1936, pages 42-44.³ Both of these show the sky-wave signal

table 1. Field strength for various height vertical antennas in millivolts per meter at one mile, power output = 1000 watts (multiply these values by 0.78 for 600-watts power output).

height		vertical radiation angle					
degrees	wavelength	0°	10°	20°	30°	40°	50°
360°	1.0	0	0	160	230	225	160
270°	0.75	0	0	0	140	210	225
230°	0.64	276	230	150	0	60	0
190°	0.528	246	230	200	130	70	50
180°	0.50	236	235	200	130	70	50
90°	0.25	195	190	180	160	140	110
	very small	186	190	180	160	140	110

patterns from verticals of various heights for a fixed distance and power as being similar to fig. 2 presented here.

Important assumptions used here were that $W = 600$ watts was a constant, and the $R_{ground} = 0$ ohms. In a later article I'll develop R_{ground} for various radial systems, and will also show the effect of network losses reducing input power to the antenna. The reduced power and non-zero values of R_{ground} will reduce the field strength of a short vertical as compared to a tall one.

What is important is that field strength is not better for a tall vertical due to the $f(h, a)$ term alone, but that the entire Kraus equation and correct assumptions

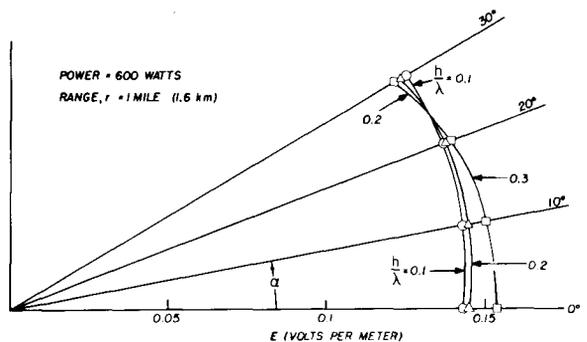


fig. 2. Vertical radiation pattern vs the height of a vertical antenna. Height in this graph is given as a ratio of the operating wavelength, h/λ

and data must be used to make a valid comparison.

This still doesn't answer the basic question of what height antenna to use or which height is better. Since the pattern versus height doesn't really matter, the important factors turn out to be earth and network losses and bandwidth. Once we know about earth losses it will be possible to answer the original set of questions and select an antenna height.

However, from a radiation standpoint, I have shown that a 0.1λ vertical (or less) is essentially the same as one 0.3λ or $\lambda/4$ wavelength tall. Furthermore, a $\lambda/2$ vertical is inferior to a $\lambda/4$ at radiation angles above 30° . It is 0.8-dB better at 10° and 0.4-dB better at 20° .

A $5/8\lambda$ vertical is inferior to a $\lambda/4$ at radiation angles above 20° . It is 0.7-dB better at 10° .

A 1λ vertical has very little radiation below 30° , but is 1.2-dB better than a $\lambda/4$ at 30° . Thus, you would have to put up a 250-foot (76.2-meter) vertical at 3.8 MHz to get about 1-dB gain for medium distances, and with poor DX performance.

There are cases where tall verticals are useful. For example, Ballantine showed back in 1924⁵ that a vertical antenna approximately $5/8\lambda$ tall provided the maximum ground-wave signal, which is important for broadcast stations (most broadcast stations use a vertical somewhat less than $5/8\lambda$ tall to reduce nighttime sky-wave interference problems).

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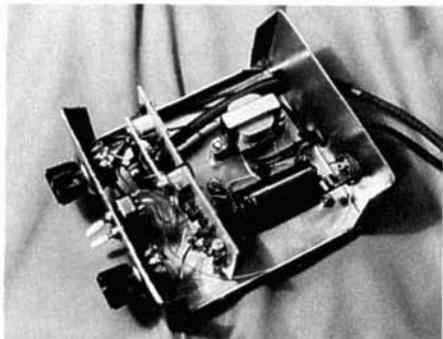
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Here's a slick CW regenerator circuit that uses a phase-locked loop to provide a virtually QRM-free, single-frequency output

C.R. Lewart, R.S. Libenschek, WB2EAX, 54 Fredric Drive, Ocean, New Jersey 07712

The circuit described here should be of interest to all CW enthusiasts. It is a new approach to the old dilemma — how do you listen comfortably to a CW station deeply imbedded in noise, hash and interference? The way this is achieved is to detect one particular CW transmission and key an independent oscillator with the received signal. The input and output frequencies may be different. The result is similar to a very narrow CW filter. The main difference between the regenerator circuit and a passive CW filter is the flexibility of the active circuit and the virtually single-frequency output. Such a clean, interference-free output cannot be achieved with a passive filter.

The circuit presents an improvement over a passive filter in a certain signal-to-



Chassis layout of the CW regenerator. Vector circuit board is mounted vertically, near the front panel. Power supply components are located at the rear.

noise range. Because the active circuit requires a minimum threshold signal voltage to operate, it will fail for extremely poor signal-to-noise ratios. Relative to other means its performance is best for intermediate to good signals. The circuit was tested by the authors with both Allied-Radio Shack SX-190 and Heathkit SB100 receivers. Both receivers provided sufficiently stable and strong CW signals at the earphone or speaker jacks to operate the CW regenerator.

The CW regenerator circuit consists of a signal amplifier, a narrowband frequency detector and trigger, an oscillator, a gate and an output amplifier (fig. 1). The details of the circuit are shown in fig. 2. The incoming signal is amplified by the transistor Q1 and its output coupled to the phase-locked loop U1. This integrated circuit has two states. In the absence of a triggering signal, its output (pin 8) presents a high impedance to ground. In the presence of a triggering frequency, f_o , the output presents a low impedance to ground.

$$f_o = \frac{1.1}{(R_4 + R_5)C_3}$$

The bandwidth of this circuit is determined by capacitor C4 and is approximately 10% of the center frequency with selected values.¹ Integrated circuit U2, another phase-locked loop IC, is used as an independent oscillator; its frequency is determined by R9 and C8. The oscillator output is gated by the output of U1. The gate itself is a p-channel mosfet, Q2. Potentiometer R15 adjusts the dc voltage at the output of the gate to the same value as the dc voltage at the input. This minimizes transients. The output amplifier, U3, is a 1/4-watt audio amplifier IC used to drive earphones or speaker. Resistors R6 and R20 and capacitors C6 and C13 are used to decouple the individual stages of the circuit.

circuit development

The circuit has evolved through a number of stages. Our original idea was to use U1 only and to either trigger its internal oscillator or let it operate a low power buzzer. The results were only partially satisfactory. The dc transients caused considerable breakup of the output signal. The next step was to add a relay and later an fet gate to combat the dc transient problem. This resulted in some improvement but we found that the internal oscillator would cause transients

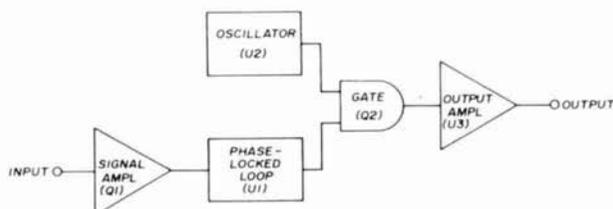
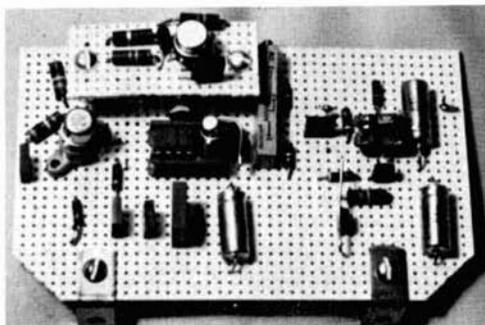


fig. 1. Block diagram of the CW regenerator. A complete schematic is shown in fig. 2.

while locking on the input signal. The use of a separate oscillator helped this situation. As the next step, input and output amplifiers were added to provide simpler interfaces.

construction

Most of the components, with the exception of the power supply, are contained on a 2½x5-inch Vector board. Although we could have used a printed-circuit board, it was found that point-to-



Vector circuit board for the CW regenerator. U2 is located on small subassembly. Q1 is center left with U1 and Q2 to the right.

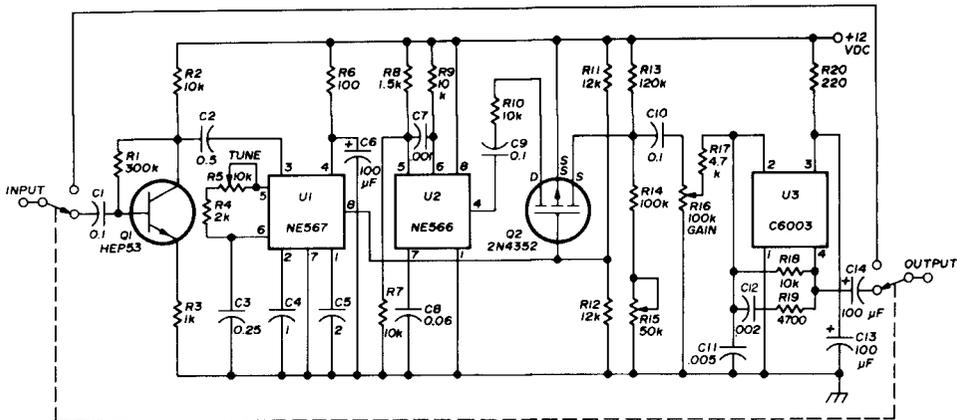


fig. 2. Schematic diagram of the CW regenerator. Integrated circuit U1 is a Signetics NE567 phase-locked loop, U2 is a Signetics NE566 function generator and U3 is a Motorola C6003 audio power amplifier.

point wiring on the Vector board afforded the easiest type of construction. All the transistors and ICs are socket mounted, with U1 and Q2 sharing a 16-pin DIP socket. The trimpot shown in the photograph is R15. This control, although not requiring frequent adjustment, might be better placed on the rear panel of the cabinet. Parts placement on the circuit board is not critical.

The homebuilt cabinet is a simple 5x5½x3-inch aluminum cabinet of U-type construction. Many types of commercially-available cabinets are suitable. The Vector board is mounted vertically toward the front with the power supply components at the rear. The control at the left on the front panel is the *tune* control, R5, while R16, the *gain* control, is at the right. The toggle switch at the bottom center is the power on-off switch. The bypass switch is located at upper center.

adjustment

Adjustment of the CW regenerator is exceedingly simple as there is only one control to adjust, the gate balance control, R15. With no signal input and a high-impedance voltmeter connected be-

tween the gate and source of the mosfet, Q2, adjust R15 for zero voltage. This is the only non-operating adjustment required.

In operation, center the desired CW signal in the passband of the receiver with the function switch in the *bypass* position and the receiver audio gain control in the center position. Place the main switch in the *regenerate* position and adjust the *tune* control, R5, to lock in the signal. This should be an absolutely pure regenerated signal with the quality of a code oscillator. There will be no QRM or

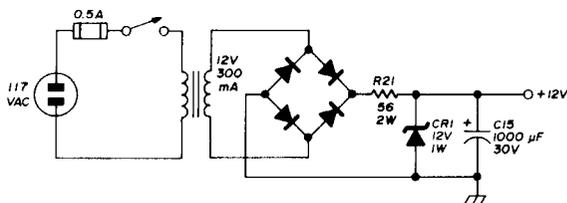


fig. 3. Power supply for the CW regenerator.

noise. With two or three signals in a pileup, the *tune* control will allow you to completely peel off the unwanted QRM. The *gain* control should be adjusted to provide a comfortable signal level.

reference

1. Signetics 567 — Tone Decoder Phase-Locked Loop data sheet and applications note, Signetics Corporation, Sunnyvale, California.

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printed circuit standards

Recently there has been a lot of interesting home-built equipment described in the amateur radio magazines. Frequently a firm offers ready-made printed-circuit boards to aid those readers who wish to duplicate one of these designs. Also, a lot of ready-made and kit-form equipment is now being offered in printed-circuit form, which the buyer can enclose as he chooses. Although these printed circuits come in all different sizes and shapes, the designer probably originally planned to mount the board on spacers inside a Minibox of convenient size. After you have built a few projects of this type you find yourself with a small pyramid of various sizes of Miniboxes, each with a power cord and other dangling wires. The organization and esthetics of the hamshack would be much improved, not to mention the ease of constructing new projects, if a few standards could be adopted for printed-circuit construction.

Interestingly enough, there already exists a *de facto* standard in this area: a card width of 4.5 inches (11.4 cm). There are probably a dozen manufacturers offering card cages which fit into a 5.25x19-inch (13.3x48.3-cm) rack panel space and accept cards up to 4.5-inches (11.4-cm) wide. A number of these card cages are adjustable for smaller widths but in any case this width is by far the most popular. For breadboarding purposes companies like Vector and Vero

offer a wide variety of boards with various hole and etch patterns, with or without edge-connector lands, and all in the 4.5-inch (11.4-cm) width. Hence the first request is that printed-circuit designers try to confine themselves to a 4.5-inch width and that manufacturers of boards center narrower patterns on a 4.5-inch space. If the user wants to mount a smaller pattern in the smallest possible space he can always trim off the excess width.

Most amateur equipment designs require rf shielding so open-card-cage-style construction is not suitable. For a single project the Minibox enclosure is entirely satisfactory for most of us. With a 4.5-inch (11.4-cm) pattern the designer might want to use a 5.5-inch (14-cm) board width, which allows a half-inch (13 mm) on each side of the pattern for the screws and spacers used to hold the board to the cover of the box. For larger installations, both of the manufacturers previously mentioned offer a nice system of cases which mount in something similar to a card cage. An example of this kind of construction is illustrated in WA6JYJ's RTTY speed-converter article in the December, 1971, issue of *ham radio*.

These cases come in several different standard heights, one of which is designed for 4.5-inch (11.4-cm) cards. Of course, if you want to take advantage of this system of mounting multiple units you must standardize on one height. This packaging system is by no means as inexpensive as the Minibox, and it isn't as easily available from the corner radio store, but the cost is still small compared to the cost of what goes into the boxes and the items are readily available from

the factory by mail order. If cost is really an obstacle you can home-make something that looks just as nice and has the same standard dimensions but costs a lot less.

Having suggested that a standard 4.5-inch (11.4-cm) width will help us to build equipment more easily and uniformly, how about the other printed-circuit board dimension? There is really no need to standardize in this direction, for nothing says that cards mounted in an open cage or in cases have to all be the same length. Again, looking at the manufacturers' catalogs we see that several different lengths are offered, but nothing much over nine inches (22.9 cm). Hence, if you can hold the length of a board to less than this, and choose cases or card cages big enough to hold a board at least this long, you can accommodate projects of all different sizes.

A card cage might look funny with different-length cards sticking out of it, but you can cover them up with an attractive, hinged, front door. Then the only problem is how to extract a short card from between two longer ones; that is easily solved with a couple of holes in the front edge and a U-shaped piece of coathanger wire with the tips bent to form hooks to engage holes in the cards.

Finally, what about external connections to the cards or cases? The most popular scheme for cards in an open cage is the familiar card-edge connector that mates with lands etched on the board. These are available in an enormous variety of sizes and features but it is surprising how often you see connectors with 22 contact positions on 0.156-inch (4-mm) centers. These are available with 22 contacts on one side or double-sided with 44.

There are many options open for the external connections to cards mounted in boxes or cases. You could just run cords out the back of the case, or have a terminal strip on the back. This doesn't ease the old tangle of wires, but at least it doesn't show from the front. A more attractive possibility is to have a connec-

tor on the back of the case which mates with a connector attached to the frame holding the cases. Then an encased unit can be removed for inspection with no dangling wires; and it can be operated in the removed position by means of an extension cord running from the plug on the case to the socket in the frame. Or, the socket can be left unattached to the frame so that the case can be pulled out for servicing without the use of an extender. Again, there are many suitable kinds of connectors that are inexpensive, such as the ordinary octal plug. My own preference is for the 22-position card edge connector. This allows encased cards to be easily intermixed in the same frame with open cards. A card in a case has the usual edge-connector lands which project through a slot in the back of the case to engage the connector. This makes it unnecessary to run a lot of wires from some other kind of connector to the card inside the case.

There are lots of ready-made experimental cards designed for this kind of construction that can be used either in cases or in open guides. Unfortunately, these are pretty high priced as compared to plain Vectorboard but they might occasionally be worth their price in the time they save. It's certainly a convenience to be able to pull out a piece of equipment and plug another in its place for testing with no wires or cables to bother with.

In summary, here are three recommendations for uniform packaging:

1. Design printed circuit boards for a standard width of 4.5 inches (11.4 cm).
2. Keep the other dimension of the board to about 9 inches (22.9 cm) or less; if a project is too large for this size, use more than one board.
3. If a printed-circuit edge connector is appropriate, use the kind having 22 contact positions on 0.156-inch (4-mm) spacing.

Jim Haynes, W6JVE

new products

solid-state ssb transceiver



The new Atlas-180 ssb transceiver recently introduced by Atlas Radio offers a number of unusual features in an amateur high-frequency ssb transceiver. Most obvious, of course, is its extremely small size: 9½ inches (24.1 cm) wide, 3½ inches (8.9 cm) high, and 9¼ inches (23.5 cm) deep. Weight is a mere 7 pounds (3.2 kg). Packed into this small package is a complete all solid-state 180-watt PEP ssb and CW transceiver for the 20-, 40-, 80- and 160-meter bands (crystal oscillator accessory available for MARS). The Atlas-180 is ideal for mobile or portable operation since it operates directly from a 12- to 14-Vdc source, negative ground, drawing 200 to 400 mA in the receive mode, 16 amps peak on transmit. An optional ac supply is available for home station use.

The Atlas-180 uses modular construction, including plug-in circuit boards, for ease of service and maintenance. Connectors on the rear of the transceiver are designed to plug into the mobile mount-

ing bracket, or into the AR-117 desk-top ac power supply, making transfer or removal a simple operation.

The receiver in the new Atlas-180 features sensitivity of less than 0.5 micro-volt (typically 0.25 μ V) for 10-dB signal-plus-noise to noise ratio while providing excellent immunity to overload and cross modulation. The input signals are converted directly to a 5520-kHz i-f without any preamplification. Selectivity is provided by a 5520-kHz crystal lattice filter with a 6-dB bandwidth of 2.7 kHz and a shape factor of 1.7. Ultimate filter rejection is 110 dB. Receiver image rejection is greater than 60 dB. Included in the receiver is an internal speaker, S-meter and 100-kHz crystal calibrator.

Receiver and transmitter frequency control is provided by a highly stable vfo circuit. The tuning dial is calibrated in 5-kHz increments and is easily interpolated to 1 kHz. Tuning rate is 15 kHz per revolution.

The transmitter has several interesting features including a broadband design which eliminates transmitter tuning and single conversion from the i-f to the output frequency. Included in the circuit is ALC and infinite vswr protection. Input power is 180-watts PEP on ssb; output is 80 watts PEP minimum (100 watts PEP typical). Unwanted sideband suppression is better than 60 dB at 1000 Hz, and carrier suppression is more than 50 dB below peak power. Intermodulation distortion is approximately 30 dB down and image outputs are more than 40 dB below rated peak output. Harmonic output is more than 35 dB below rated peak power.

Accessories available for the Atlas-180 include the AR-117 table-top 117-volt ac power supply and mobile mounting bracket (deluxe plug-in and non-plug-in models available). Automatic VOX (CW semi-break-in), phone patch and other accessories will be announced in the near future. For more information, write to Atlas Radio Inc., Post Office Box A, Carlsbad, California 92008, or use *check-off* on page 94.

antenna tower protection system

A new system introduced by the Towtec Corporation automatically guards against tower and antenna damage from high winds. Called *Towgard*, the system continuously monitors your local wind velocity. When the wind velocity reaches or exceeds the changeable, preset 35 mph limit (even on gusts) the system will automatically lower your motorized tower. Towtec also manufactures electric hoists for manual crank-up towers.

The *Towgard* system includes a computer/controller, a wind sensor and a bottom limit switch for automatic motor turn-off when the tower is nested. For more information on this automatic system, write to the Towtec Corporation, 118 Rosedale Road, Yonkers, New York 10710, or use *check-off* on page 94.

cambion catalog

The new Cambion XQ Components Catalog illustrating various Cambion (CTC) components now available through retail outlets is now being offered at no charge. Many popular items are illustrated, such as terminals, jacks, plugs, handles, battery holders, IC sockets, IC breadboards, coils and rf chokes.

Cambridge Thermionic Corporation, better known as CTC, has been a manufacturer and supplier to industry of precision electronic components for over thirty years and has just recently made its Cambion products available to the amateur and experimenter. Industrial, commercial, military and aerospace engineers use more than 20,000 different components in the Cambion line. Through this catalog the amateur and experimenter now have available the same quality and reliability.

For further information on the new Cambion XQ Components Catalog, write to Cambion, Department XQ, 145 Concord Avenue, Cambridge, Massachusetts 02138, or use *check-off* on page 94.

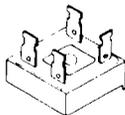
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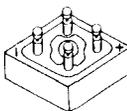
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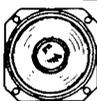
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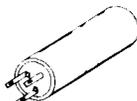
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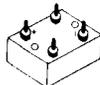
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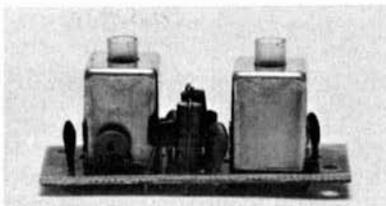
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Hamtronics, Inc. has announced several new products for the vhf buff. Of primary interest is a new improved version of the well-known cascode 6- and 2-meter preamps described in articles in March, 1972, and January, 1973, issues of *ham radio*. The new units are $\frac{3}{4}$ -inch wide x 2-inches long x 1-inch high (19x51x25-mm) so they will fit into almost any receiver or transceiver. New fets used in the preamp provide more gain and better noise figure while retaining the inherent stability of the cascode circuit which has made the Hamtronics preamp so popular. Amateurs like this unit because they can tune it easily without worry about neutralization.

With several thousand preamps now in the field, hams are reporting that even newer transceivers are improved by adding a low-noise preamp ahead of the front end. Units are still \$6.00 in kit form or \$10.00 wired and tested, postpaid in the USA. Club prices are still available. Models are also available now for any frequency from 20 to 240 MHz, including 10-meter Oscar reception.

Other new products to be announced soon include a lower-cost, more-compact scanner device for up to four channels; a receiver kit for a-m reception on aircraft frequencies or 6 or 2 meters for net use; and 450-MHz receivers, preamps, and transmitters for fm. Also, kits are now available for the popular 5-watt audio amplifier, using an integrated circuit, as written up in an article appearing in September, 1972, *ham radio*.

For more information, send a self-addressed, stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612, attention Jerry Vogt, WA2GCF, or use check-off on page 94.

digital voltmeter

Now a low-priced, laboratory-quality digital voltmeter is available from MITS, Inc. The model DVM 1600 measures alternating and direct current in five ranges from 0.1 mA to 1 amp. Ac and dc voltage is measured in four ranges from one volt to 1000 volts. Measurement of resistance is in six ranges from 100 ohms to 10 megohms.

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The DVM 1600 also features auto polarity which automatically displays polarity and magnitude without probe reversal. Other features include a regulated power supply and 100% overrange capability on all ranges. Power requirements are 115/230 Vac, 50/60 Hz, 20 watts.

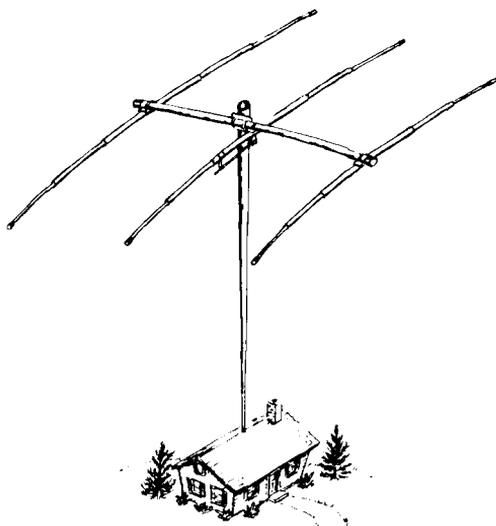
The MITS DVM 1600 digital voltmeter is available in an easy to assemble kit form (\$89.95) or factory assembled (\$129.95). Warranty on the assembled model is one year on parts and labor. Kit warranty is ninety days on parts. For more information, write to MITS, Inc., 6328 Linn Avenue, NE, Albuquerque, New Mexico 87108, or use *check-off* on page 94.

short circuit

transistor curve tracer

There are several errors in the schematic for the transistor curve tracer published in the July, 1973, issue, page 53. There should be a 0.02- μ F capacitor connected from the collector of Q1 to the collector of Q2; the value of C4 should be 0.2 μ F, not 0.02. Also, CR9 should be a 1N457. When building this unit be sure that the circuit ground (the common ground for Q1, Q2, Q3, Q5 and Q7) is separate from the chassis grounds used on the oscilloscope jacks (J1 and J2).

More Details? CHECK-OFF Page 94



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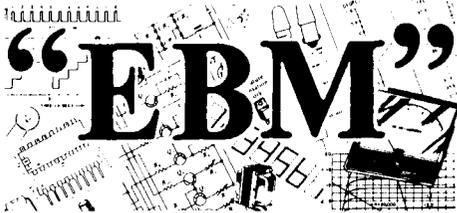
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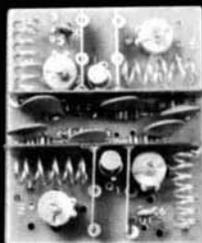
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The RVD-1002 is compatible with signals from any TU at 60, 66, 75 or 100 WPM. The RVD-1002 means an end to the headaches of electromechanical printers — and the beginning of silent, reliable, low-power-consuming, trouble-free RTTY reception.

Whether you're into amateur RTTY, or thinking about it, or just interested in "seeing" what all those RTTY signals you hear are saying, the RVD-1002 is the perfect answer. HAL also offers the RVD-2110 (a complete, all-channel TV set, plus RTTY read-

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If you're looking for better television programming, get into the exciting world of video-displayed RTTY from HAL. Whether you want to ham it up or catch the latest news, or just explore the very wide world of RTTY entertainment, HAL has it. And it's a really big show!

RVD-1002 video unit: \$575.
RVD-2110 monitor/TV: \$140.
ST-6 RTTY terminal unit: \$310.
RKB-1 TTY keyboard: \$250.
RKB-1 TTY keyboard: \$250.



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Telephone: (217) 359-7373

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In the RTTY mode, you can transmit at standard data rates of 60, 66, 75 or 100 WPM, as well as an optional 132 WPM, 100 baud. In addition to the complete alphanumeric keys, you get 17 punctuation marks, 3 carriage control keys, 2 shift keys, a break key, 2 three-character function keys, a "DE-call letters" key and a "Quick brown fox . . ." test key.

In the CW mode, you can send at speeds anywhere between 8 WPM and 60 WPM. You can also adjust dot-to-space weight ratios to your liking. For CW, you have all alphanumeric keys, plus 11 punctuation marks, 5 standard double-character keys, 2 shift keys, a break-for-tuning key, error key, "DE-call letters" key, plus

2 three-character function keys. Output interfacing is compatible with cathode keying or grid-block keying. A side tone oscillator and built-in speaker allow you to monitor your signal — with adjustable volume and pitch controls.

The DKB-2010 also has a three-character memory buffer which operates in either the RTTY or CW mode, allowing you to burst type ahead without losing characters. A 64-character memory buffer is also available as an option. Key function logic in either mode is governed by LSI/MOS circuitry. All key switches are computer grade.

The DKB-2010 is available assembled or in kit form. Should you choose the kit, you'll find construction easy — the unit consists of three assemblies: power supply board, logic PC board, keyswitch PC board, and pre-assembled wiring harness.

Any way you look at it — as an easy-to-build kit, a complete assembly, as a CW keyboard, or an RTTY keyboard, the HAL

DKB-2010 is a real breakthrough for every amateur. It adds a whole new dimension to the exciting world of amateur radio. Once you've used the DKB-2010, you'll wonder how you ever got along without it!

Prices: \$425 Assembled;
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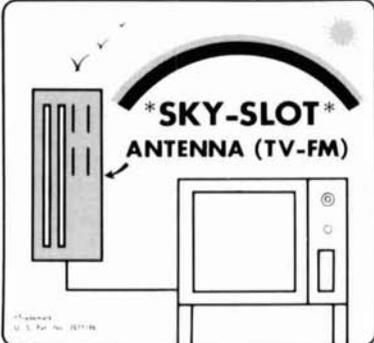
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7400	5	29	7447	5	45	74123	5	15
7401	25	1488	7448	150	74125	69		
7402	25	1450	7449	29	74126	95		
7403	25	1451	7450	32	74141	176		
7404	29	1451	7451	72	74145	176		
7405	27	1454	7452	45	74150	125		
7406	55	1455	7453	32	74151	105		
7407	53	1460	7454	30	74151	145		
7408	29	1461	7455	30	74154	175		
7409	29	1464	7456	49	74155	135		
7410	25	1465	7457	45	74156	150		
7411	35	1470	7458	50	74157	150		
7412	95	1472	7459	45	74161	165		
7413	50	1473	7460	25	74163	180		
7416	50	1474	7461	55	74164	295		
7417	50	1475	7462	95	74165	295		
7420	25	1476	7463	55	74166	195		
7421	32	1478	7464	80	74173	195		
7422	32	1481	7465	125	74175	195		
7423	37	1485	7466	120	74176	95		
7425	39	1486	7467	55	74177	95		
7426	25	1489	7468	125	74180	135		
7427	39	1490	7469	125	74181	425		
7430	25	1491	7470	40	74182	110		
7432	30	1492	7471	105	74190	165		
7433	50	1493	7472	105	74192	165		
7438	55	1494	7473	105	74193	165		
7440	25	1495	7474	105	74194	165		
7441	125	1496	7475	105	74195	165		
7442	115	1498	7476	85	74196	135		
7443	125	1498	7477	115	74197	115		
7444	130	1499	7478	55	74198	250		
7445	125	1499	7479	65	74199	230		
7446	145	1499	7480	55	74199	230		

Low Power TTL

74100	5	40	74151	5	40	74190	5	75
74102	50	1493	74152	50	1494	74191	150	
74103	40	1493	74153	40	1494	74192	175	
74104	40	1493	74154	60	1494	74193	175	
74108	40	1493	74155	60	1494	74194	175	
74109	40	1493	74156	80	1494	74195	195	
74110	40	1493	74157	80	1494	74196	195	
74111	40	1493	74158	80	1494	74197	195	
74120	40	1493	74159	80	1494	74198	195	
74130	40	1493	74160	80	1494	74199	195	
74147	175	1493	74161	85	1494	74200	195	

High Speed TTL

74H1	5	40	74H01	5	45	74H90	5	45
74H01	5	40	74H02	4	47	74H03	45	
74H02	40	1494	74H04	40	1494	74H05	45	
74H04	45	1494	74H06	45	1494	74H07	45	
74H08	45	1494	74H09	45	1494	74H10	45	
74H10	40	1494	74H13	47	1494	74H16	70	
74H20	40	1494	74H55	47	1494	74H76	70	
74H20	40	1494	74H55	47	1494	74H76	70	

8000 Series TTL

8054	5	45	8200	5	295	8554	5	295
8060	30	8210	8210	195	8570	295		
8061	69	8214	195	8600	115			
8062	69	8219	195	8810	95			
8063	69	8220	195	8812	125			
8064	69	8230	195	8822	295			
8121	105	8290	95	8830	60			
8122	105	8298	105	8831	295			
8123	175	8520	145	8832	295			
8130	250	8551	195	8836	69			
8132	175	8552	195	8880	150			

ALL DIP PKGS

Specify spec sheets required with order. Add \$ 50 per spec sheet for items less than \$1.00 ea.

CMOS

74C00	8	85	74C76	1	70	74C163	3	26
74C02	85	74C107	1	70	74C184	3	50	
74C04	95	74C151	2	80	74C173	2	80	
74C10	85	74C154	2	80	74C172	2	80	
74C15	85	74C157	2	80	74C193	3	26	
74C42	2	74C190	3	26	74C195	3	00	
74C43	1	74C191	3	00	80C97	1	50	
74C74	1	74C192	1	30				

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TTL (DIP)

7402	Quad 2-input NOR gate	5/\$1.00
7410	Triple 3-input gate	5/\$1.00
7437	Quad 2-input NAND buffer	3/\$1.00
7438	Quad 2-input NAND buffer O.C.	3/\$1.00
7460	Dual 4-input expander	4/\$1.00
7476	Dual J-K flip-flop	\$ 45 ea.
7490	Decimal counter	1.10 ea.
74121	One shot monostable multivibrator	49 ea.

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75451	Dual peripheral driver	.39 ea.
75453	(351) Dual peripheral driver	49 ea.

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LM305	Positive Voltage Regulator	TO 5	1.25 ea.
LM307	Op Amp (super 741)	TO 5 or MINI DIP	46 ea.
LM308	Micro Power Op Amp	TO 5 or MINI DIP	1.25 ea.
LM309H	5 V Regulator	TO 5	1.25 ea.
LM309K	5 V Regulator	TO 3	1.95 ea.
LM310	Voltage Follower Op Amp	TO 5	1.45 ea.
LM311	Hr perf Voltage Comparator	TO 5 or MINI DIP	1.25 ea.
LM319	Hr Speed Dual Comparator	DIP	1.66 ea.
LM320	5.2 V Negative Regulator	TO 3	1.95 ea.
LM320	12 V Negative Regulator	TO 3	1.95 ea.
LM320	15 V Negative Regulator	TO 3	1.95 ea.
LM339	Quad Comparator	DIP	1.95 ea.
LM340T	Positive Voltage Regulator	TO 220	2.29 ea.
LM370	ACC Squash AMPL	DIP	1.25 ea.
LM372	AF IF Strip-detector	DIP	85 ea.
LM373	AM/FM/SS Strip	DIP	3.60 ea.
LM376	Vox Volt Regulator	MINI DIP	85 ea.
LM380	2 Watt Audio Pre-AMP	DIP or MINI DIP	1.25 ea.
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LM1304	FM Multiplex Stereo Demod	DIP	1.50 ea.
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LM8801	Retriggerable One Shot	DIP	49 ea.
LM75451	Dual Peripheral Driver	MINI DIP	49 ea.
LM75452	Dual Peripheral Driver	MINI DIP	49 ea.
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Specify TO 5, DIP or MINI DIP Package

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74C107	Dual J-K flip-flop with clear	1.25 ea.
74C160	Decade counter with sync. clear	2.75 ea.

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MM1405	512 bit dynamic shift register	DIP TO 5 55 ea.							
MM1918	Dual 256 bit mask prog. shift register	TO 5 25 ea.							
MM1950	Dual 32 bit static shift register	TO 8 35 ea.							
MM1950A	Dual 64/72/80 bit static shift register	DIP 35 ea.							
MOS Shift Registers 7900 Series									
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NE 568	Function Generator	MINI DIP OR TO 5	2.85 ea.
NE 567	Phase Decoder	MINI DIP OR TO 5	2.85 ea.

Specify TO 5, Dip or Mini Dip Package

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CD 4002	65	CD 4013	150	CD 4023	65			
CD 4009	100	CD 4016	150	CD 4025	65			
CD 4010	65	CD 4017	295	CD 4027	135			
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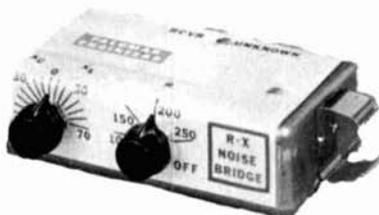
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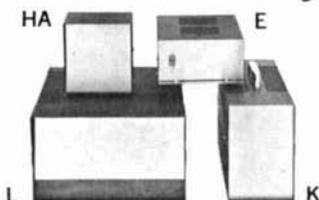
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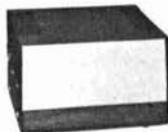
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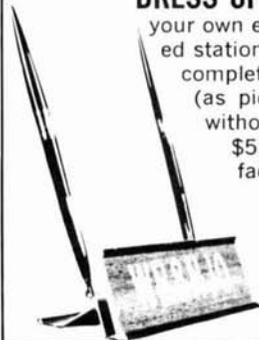
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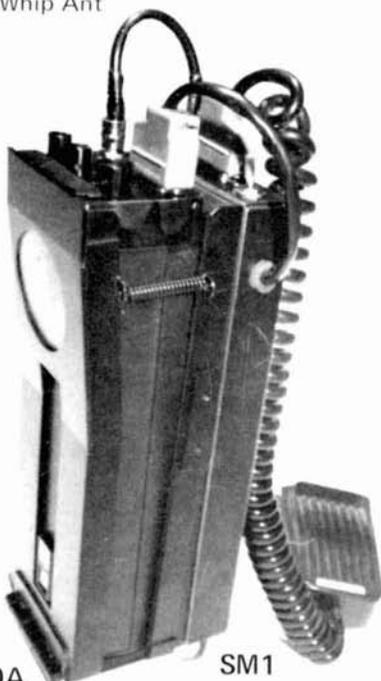
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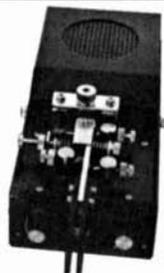


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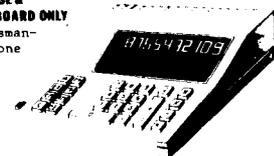
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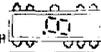
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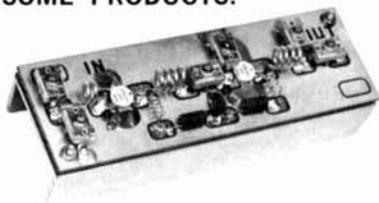
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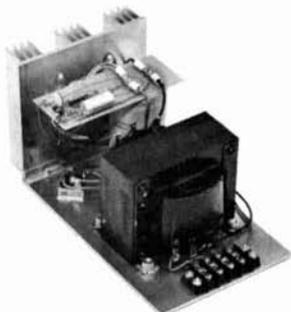
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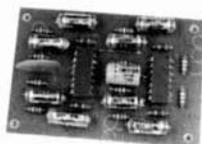
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Build the 2'x3' CWF-2 PC card into your receiver or get the self contained and ready to use CWF-2BX and plug in!

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SKIRT REJECTION	At least 80 db down 1 octave from center frequency for 80 Hz bandwidth
CENTER FREQUENCY	750 Hz
INSERTION LOSS	None. Typical gain 1.2 at 180 Hz BW, 1.5 at 110 Hz BW, 2.4 at 80 Hz BW
INDIVIDUAL STAGE Q	4 (minimizes ringing)
IMPEDANCE LEVELS	No impedance matching required
POWER REQUIRED	CWF-2: 6 volts (2 ma.) to 30 volts (8 ma.), CWF-2BX: standard 9 volt transistor radio battery
DIMENSIONS	CWF-2: 2'x3' PC board, CWF-2BX: 4'x3' 1/4'x2' 3/16' (black wrinkle steel top, white aluminum bottom, rubber feet)

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STANDARD 146-A (1-2) \$238.70, (3-11) \$212.30, Nicard batteries \$1.58, Stubby antenna \$5.00. Standard 826MA (1-2) \$324.50, (3-11) \$306.90. Standard 851T 25 watt mobile \$420.20. Standard RTP-1 repeater \$600.00. HM-175 antenna \$16.00. Base station antenna HM-191 8.25 dB (list \$169.50) net \$119.95. Send check and we'll pay postage or we will ship COD. Electronics Communications Co., P. O. Box 17222, Nashville, Tenn. 37217 (24 hr) 615-834-8999.

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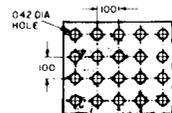


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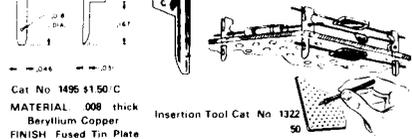
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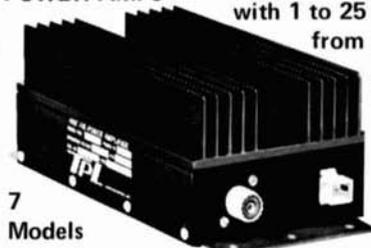
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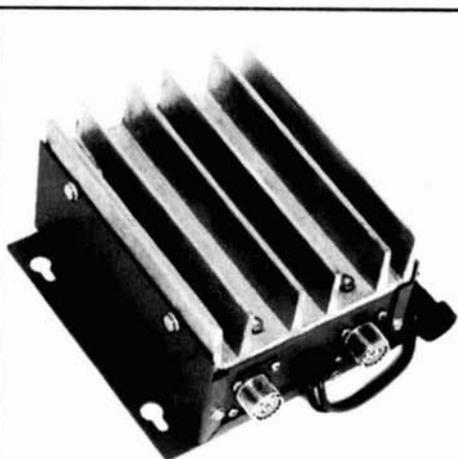
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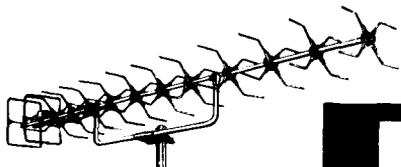
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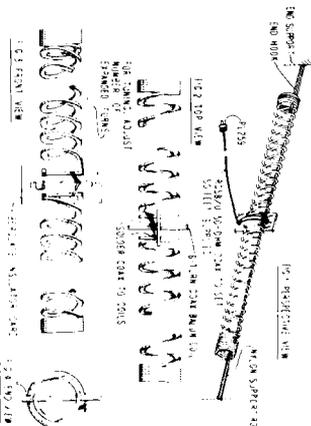
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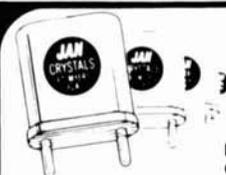
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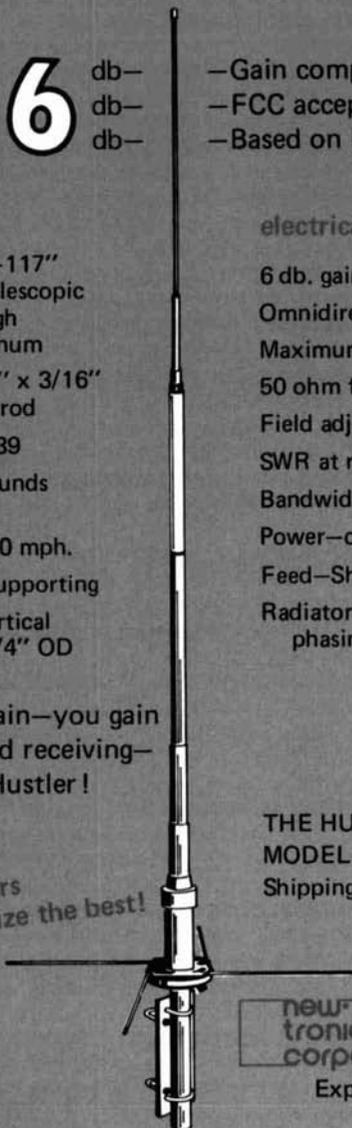
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