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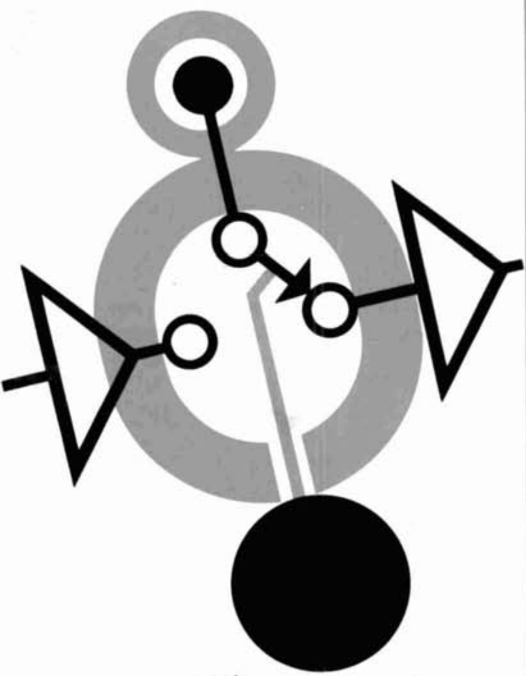
NOVEMBER 1974



low-power single-band cw transceiver

this month

- artificial radio aurora 18
- vhf channel scanner 26
- predicting harmonics 34
- low-frequency converter 49
- solar power 52

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More Details? CHECK—OFF Page 94

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volume 7, number 11

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ham radio magazine is
published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048

Subscription rates, world wide
one year, $7.00, three years, $14.00
Second class postage
paid at Greenville, N.H. 03048
and at additional mailing offices

Foreign subscription agents
United Kingdom
Radio Society of Great Britain
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London WC1N 2AE, England

All European countries
Eskil Persson SM5CJP
Frotunagrand 1
19400 Upplands Vasby, Sweden

African continent
Holland Radio, 143 Greenway
Greenside, Johannesburg
Republic of South Africa

Printed by Wellesley Press, Inc
Framingham, Massachusetts 01701, USA

Microfilm copies of current
and back issues are available
from University Microfilms
Ann Arbor, Michigan 48103

ham radio *magazine*

contents

8 low-power cw transceiver
Wesley H. Hayward, W7ZOI

18 scattering characteristics of artificial radio aurora
Victor R. Frank, WB6KAP

26 vhf fm channel scanner
Gerald F. Vogt, WA2GCF

32 *measuring peak envelope power*
Carl C. Drumeller, W5JJ

34 harmonic prediction
Fin K.L. Utne

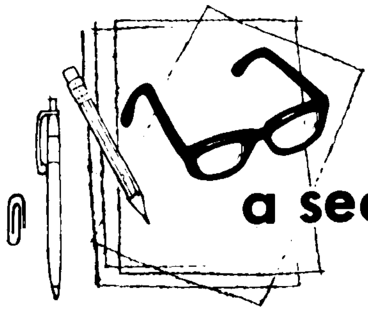
38 the code mill
H.E. Thomas, W6CAB

44 automatic phone controller for repeaters
Robert C. Heptig, K0PHF
Robert D. Shriner, WA0UZO

49 tunable low-frequency converter
Guenter Ruehr, OH2KT

52 solar power
Edward M. Noll, W3FQJ

4 a second look
94 advertisers index
58 *comments*
83 flea market
60 ham notebook
94 reader service
6 *stop press*



a second look

by jim fisk

Here's a new technology that you're going to be hearing a lot more about in the not too distant future: *additive* printed circuit wiring. If you're wondering what's new, the process used to manufacture most of the PC boards used nowadays is called the *subtractive* process: you start with a copper-clad board and etch out the desired conductor pattern. In the additive process the copper is deposited on the board in the desired conductor pattern. The big advantage is that the process uses up to 75% less copper — obviously important to an industry that is facing a real squeeze by the world-wide copper shortage.

Additive printed-circuit wiring has been around for several years, but until recently users have been troubled by peeling and edge lifting, as well as warpage and cracking. Now, however, many of these problems have apparently been solved, and it appears that the additive technique is about to produce a substantial portion of the PC boards now manufactured by the conventional subtractive process. In fact, some industry spokesmen indicate that the additive process will account for up to 50% of the market by the end of 1975.

In addition to the conservation of copper, the additive printed-circuit process eliminates undercutting due to etching and offers improved circuit design capabilities such as closer line-width tolerances. For example, in standard etched printed-circuit boards with 1-ounce copper foil, the minimum line width is about 10 mils; with the additive process 5-mil line widths are routine. Closer manufacturing tolerances in the near future are expected to allow line widths down to 2 mils.

In addition, when used for high production quantities, the cost of boards produced by the additive technique are considerably less expensive. In mass applications such as television sets and stereo systems, the savings over subtractive PC boards can be as much as 35%. And high production is not the only promising area for additive circuits — because of the smaller line widths and spacings the technique is expected to find wide use in high-density interconnection applications such as computer peripherals.

The principal problem that had to be solved before the additive technique became a viable manufacturing process was finding a way to make the electroless-copper stick to the glass-epoxy laminate. This adhesion (typically 8- to 10-pounds pull) is provided by micropores on the laminate surface. This is no easy task, but one successful approach is to coat the bare laminate with a special adhesive which is activated by a chromic-acid etch. The etchant produces micropores in the adhesive surface and is similar to the treatment used in the plating of plastics.

I don't expect that this technique will replace the various subtractive systems that amateurs have developed to make their own printed circuit boards, not in the near future anyway. However, printed-circuit boards were being used by the manufacturers of television sets long before the process was even considered for one-off amateur projects, so nothing will surprise me. Perhaps one of the chemists among us will come up with a simple additive PC process that we can use in our own home workshops.

Jim Fisk, W1DTY
editor-in-chief

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FCC AMATEUR RADIO ADVISORY COMMITTEE PROJECT, proposed in ARRL/FCC meeting May 10th beginning to warm up as Amateur and CB Division staff lay groundwork for a formal presentation to the Commissioners. A letter has been sent to quite a number of amateurs outlining the purposes and responsibilities of the ARAC.

Letter Was Definitely NOT An Invitation to membership in the ARAC, but instead was sent out as a means of determining whether there would be sufficient support from amateurs to insure ARAC success if the Commissioners were to approve it. Letter proposed basic 12-15 member "Steering Committee" to be chaired by FCC representative, a number of Ad Hoc subcommittees (each chaired by a Steering Committee member) to consider specific problem areas. Steering Committee membership would be by invitation of the Commission, with broad geographical as well as interest-area representation.

ARRL PREPARATIONS FOR 1979 World Telecommunications Conference should be aided by W1RU and K1ZND having participated in a frequency management seminar in Geneva during September.

BLOCKBUSTER TRANSCEIVER FROM HEATHKIT announced this month (see centerfold). The new SB-104 is all solid state, totally broadbanded, digital readout and covers all ham bands from 3.5 to 29.7 MHz! Delivering 100 watts SSB or CW, the SB-104 operates from 13 Vdc, weighs only 20 pounds and measures about 6x14x14 inches; its 6-digit readout provides 100-Hz resolution, 200-Hz accuracy.

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GENAVE GOING TO DIRECT SALES IN THE AMATEUR MARKET, with change becoming official November 1st. Inflationary pressures, determination that 75-80% of Genave's ham gear has been selling by mail order anyway are cited as the reasons. High proportion of mail-order sales has meant most service has been factory responsibility, so change to factory selling is a logical one.

Genave's Pricing will reflect the change -- list on the popular GTX-200 will go from $399.95 to $299.95, and other models will show similar reductions.

HILDA WETHERBEE LEAVES HAM RADIO. Hilda, a key part of HR since its beginning and well known to many amateurs and the ham industry for her participation in many hamfests and highly effective ad selling, has joined a research firm in New Hampshire.

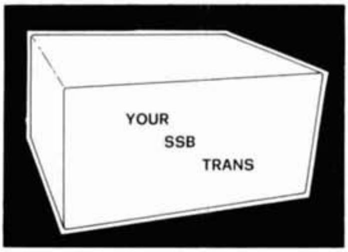
MOBILERS BEWARE: Minnesota has joined Florida and Virginia in forbidding the wearing of headphones while driving.

ANOTHER 10-METER BEACON PLANNED, this one proposed for San Diego by K6HME and WB6KNC. Request for Special Temporary Authority has been returned for justification of proposed 100-watt power and SSB ID, but as soon as questions are resolved it should be operative using K6HME call. Proposed frequency is 29.0 MHz, same as German DLØAR beacon.

"COMMERCIAL" USE OF HAM BANDS subject of concern in Washington, confusion among hams. Recent League Lines item (August, 1974 QST) caused most nets to scrub "swap shop" sessions, but one subsequent interpretation seems to permit such activities providing they are scrupulously "non-commercial" and any given individual limits himself to "infrequent" participation (whatever that means)...

JA STATIONS TO GET NEW 80 METER "WINDOW" within the next month or so. Present JA allocations are 3500-3525 CW, 3525-3575 for phone; the new band is a 10-kHz slice from 3793-3803 kHz, will help JA-Europe and JA-U.S.A. QSOs.

Slow-Scan Enthusiasts can also look for a big increase in JA SSTV activity very shortly. Although over 100 stations have been licensed for slow-scan in Japan since April, 1973, reorganization of the Japanese Ministry of Post and Telecommunications has caused a big backlog to develop. Kinks are now being worked out, and a big influx of JA SSTVers is expected momentarily.



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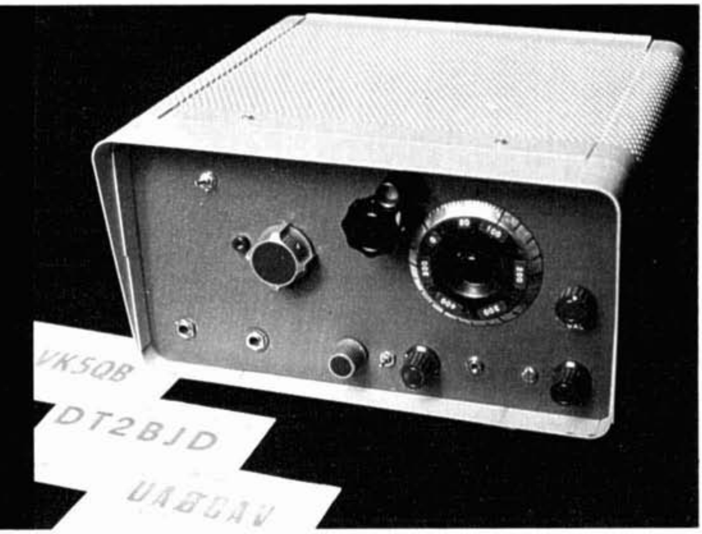


low-power single-band cw transceiver

Design and construction of a deluxe 1.5-watt QRP CW package for 20 meters

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

There is little doubt that low-power (QRP) operation has become one of the more popular amateur radio activities in recent years. I got hooked on the QRP game a little over ten years ago when a need arose for light-weight portable gear for use on mountaineering trips. Although this goal was easily and quickly realized, my interest continues. Even today, after a few thousand QSOs with a power of only a few watts, I still get a thrill when another QRP contact is completed.

While some manufacturers are now seriously aiming products at the QRP market, the area is still ripe for the home experimenter. To many QRP enthusiasts, the only equipment which is considered for construction is that which is as simple as possible. While simplicity certainly has its merits, especially for portable operation in a severe environment,[1,2] home-station operation is greatly enhanced with equipment which is a bit more elaborate. This article describes a transceiver for 20-meter CW which is aimed at this improved performance.

One of the more significant deficiencies of many QRP stations is the lack of selectivity and dynamic range in the receiver. Although it is possible to obtain superb performance from a direct-conversion design, as demonstrated by the recent work of DeMaw,[3] there is still no substitute for a cleanly operating superhet. Hence, this approach was taken in this design.

Some of the simplicity of a direct-conversion design is retained by eliminating virtually all of the gain usually found at the intermediate frequency. The result is a receiver which is more than adequately sensitive and selective, but is still easy to duplicate. Provision is made for receiving both CW and ssb in the unit described, with sideband included mainly for use with vhf converters. Further simplification and reduced cost will result if one of the modes is deleted.

The transmitter portion of the package was designed with a number of objectives

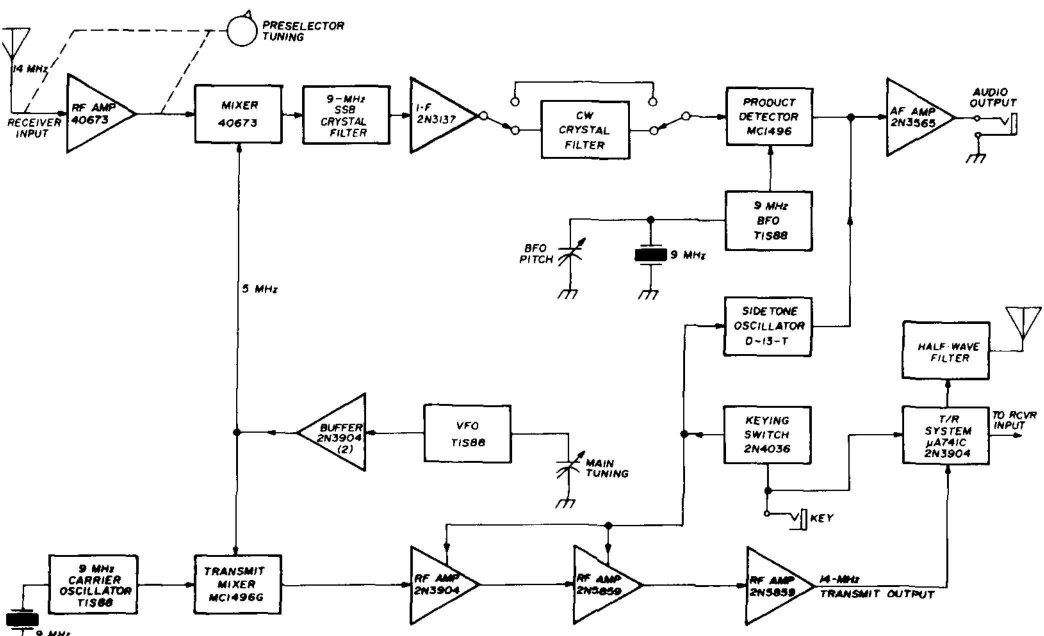


fig. 1. Block diagram of the deluxe QRP CW transceiver for 20 meters.

in mind. First, full transceive operation was desired. However, it was not acceptable to sacrifice the cleanliness of the system output. Experience with an earlier 40-meter transceiver demonstrated that this objective is not easily met with a casual design. The unit described in this article has been evaluated with lab quality test equipment, and all spurious outputs were more than 50-dB below the desired 1.5-watt output.

Additional design criterion for the transmitter included the desire for a built-in sidetone oscillator and a semi-break-in keying system. These features are not absolutely necessary, but are easily realized and add measureably to the operating enjoyment derived from the station.

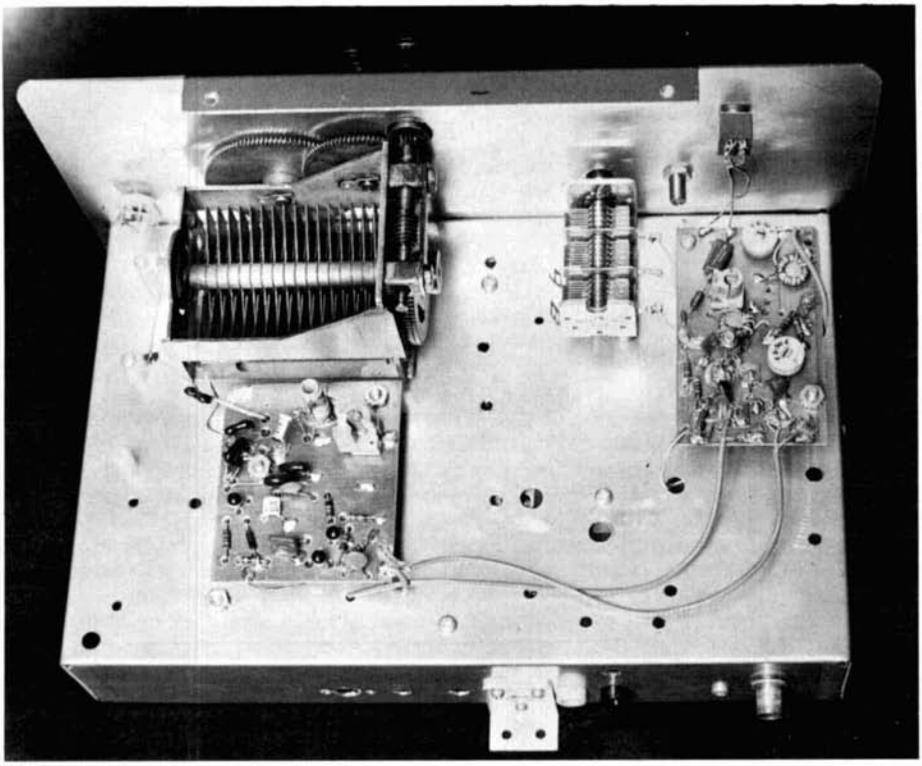
Shown in **fig. 1** is a block diagram of the transceiver. The usual 9-MHz i-f and 5-MHz vfo scheme is used, providing good performance in a single-conversion system. A total of seven individual circuit boards are used, most of them based upon a simplified double-sided PC board technique outlined earlier. The use of double-sided board is highly recommended since it provides the low-impedance ground paths required for clean, spur-free performance.

vfo design

Shown in **fig. 2** is the variable-frequency oscillator which controls both the receiver and the transmitter. The design offered several years ago by Hanchett[4] was used, although the original mosfet was replaced with a junction fet. I keep coming back to this design since it is both stable and reliable. As with most oscillators, the selection of components is a large part of achieving stability. The inductor is wound on a ceramic form which was originally tuned with a powdered-iron slug. However, the slug was removed to enhance stability, and the number of turns on the coil was pruned to obtain the proper inductance. The vfo is tuned with the ever reliable and easy-to-use capacitor from a surplus ARC-5 Command transmitter. Although becoming scarce, these capacitors can still be found in junk boxes and at hamfest flea markets. With the components shown, the oscillator tunes from 5.0 to 5.55 MHz.

A number of commonly available field-effect transistors can be used in this oscillator. I used a TIS88 which is a plastic device very similar to the popular 2N4416. The Motorola MPF102 also performs well in this circuit. The oscillator supply voltage is stabilized with a 6.8-volt zener diode. The voltage rating here is not critical.

The vfo is buffered with a single-stage feedback amplifier using a pair of 2N3904 transistors. Again, transistor type is not critical in this application, although devices with an f_T of at least 250 MHz should be used. The output of the buffer is 3 volts, peak-to-peak and sinusoidal. If it is suspected that significant harmonic energy might be present in the oscillator output, a lowpass filter could be included.[5]



Top view of the 20-meter CW transceiver, showing the vfo (left) and receiver rf amplifier (right).

receiver front-end and i-f filtering

Presented in **fig. 3** is the front-end and filter section of the receiver. For the most part, the design is quite standard and is easily duplicated. Both the rf amplifier and the mixer circuits use RCA 40673 dual-gate mosfets. Since my transceiver tunes a 500-kHz range, a dual-section variable capacitor was included for front-panel preselector tuning. An earlier version of this transceiver tuned only the CW portion of the band, and front-panel tuning was not needed. The toroid cores used in the receiver and in the transmitter section described later are very similar in on the respective coils. The minimal degradation in gain and noise figure should present no problem in typical applications. The dynamic range of this receiver has not yet been measured.

A coarse rf gain control is provided by a front-panel switch which decreases the gate-2 bias from the nominal 4-volt level to ground. This yields a gain reduction of about 15 dB. Since the application of

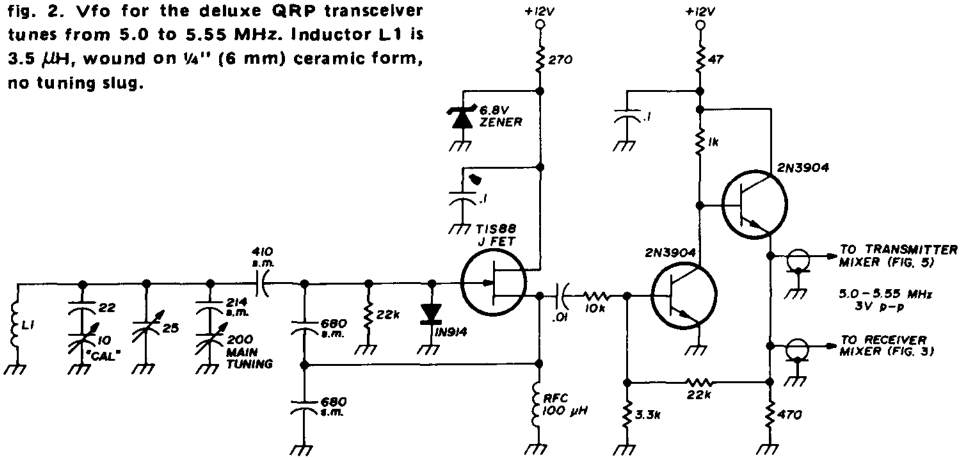


fig. 2. Vfo for the deluxe QRP transceiver tunes from 5.0 to 5.55 MHz. Inductor L1 is 3.5 µH, wound on 1/4" (6 mm) ceramic form, no tuning slug.

inductance and Q to the Amidon T-50-6.* Substitution of Amidon cores should be possible using the same number of turns shown.

Sensitivity measurements revealed a system noise figure of around 7 dB. Hence, the receiver is probably quite a bit hotter than can be used in normal locations. Much of the dynamic range possible with mosfets was retained by terminating the mixer output with a 330-ohm resistor. Since drain nonlinearity is the usual mechanism for blocking in a fet mixer, a decrease in output load impedance results in relative freedom from this problem. Similarly, the load resistance presented to the drain of the rf amplifier is about 1000 ohms. If dynamic range problems are encountered, further improvement would result if the gate of each stage were driven from a tap

gain reduction can often decrease the immunity of an fet rf amplifier to cross-modulation and IMD, passive front-end signal attenuation would be a better means of gain control.

The output of the mixer is routed through a coax cable to the first crystal filter. This filter is always in the signal path and is used for ssb reception. The filter I chose was the WF-8 model manufactured by Wheatlands Electronics.† This eight-pole unit performs well in this application, and probably represents one of the better component buys around.

The output of the first filter is applied to a low-gain amplifier. A junk box 2N3137 was used, although this stage is not critical, and could probably be replaced with a 2N3904 or a 2N4124. Alternately, if some impedance matching is done at the input, a dual-gate mosfet

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

†Wheatlands Electronics, P.O. Box 343, Arkansas City, Kansas 67005.

with a 510-ohm drain load resistor should perform well in this slot. The main function of this stage is to provide a proper driving impedance for the second filter with a minimum gain.

During CW operation, a KVG type XF-9M four-pole crystal filter is switched into the system.* The switching is done with a pair of inexpensive slide switches which are mounted on the circuit board.

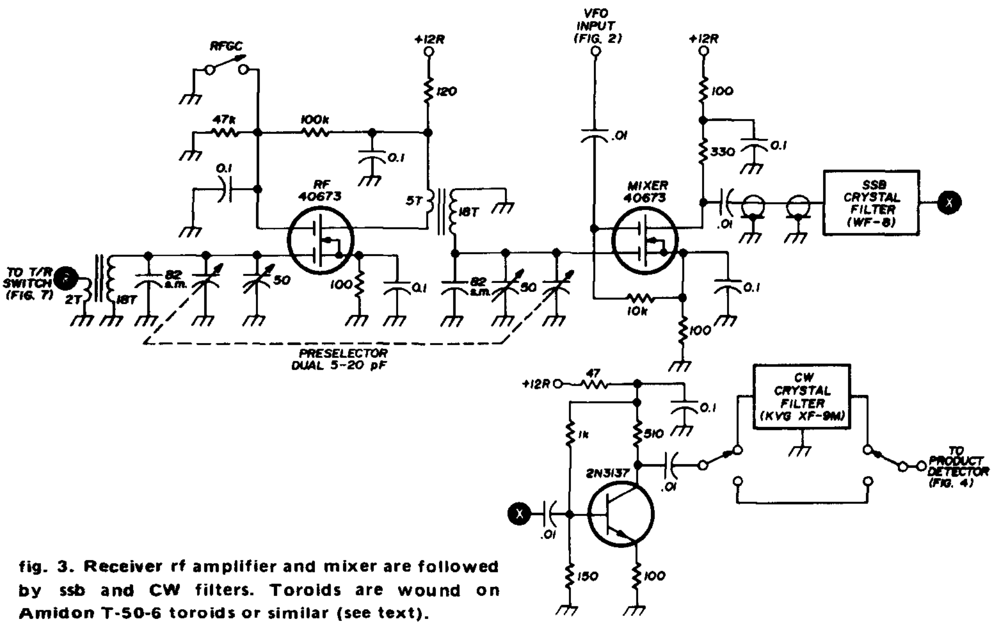


fig. 3. Receiver rf amplifier and mixer are followed by ssb and CW filters. Toroids are wound on Amidon T-50-6 toroids or similar (see text).

The switches are ganged together by drilling small holes in the plastic handles and attaching a strip of scrap PC board. A spade lug is also bolted to the connecting strap. A tapped, ¼-inch (6-mm) spacer is then screwed onto the spade lug. This spacer extends through the front panel where a knob is mounted.

product detector and audio

Shown in **fig. 4** is the product detector and bfo used in the receiver. The detector uses a Motorola MC1496G integrated circuit. The configuration is generally the same as that shown in the Motorola applications literature, except that the input resistors have been chosen to properly terminate the KVG crystal filter. The bfo is a simple fet crystal oscillator which is trimmed from the front panel with an 80-pF variable capacitor.

The audio section consists of a pair of 2N3565 transistors and provides around 70-dB of gain to drive high-impedance headphones. This amplifier is built on the same board that contains the sidetone oscillator and control circuitry and is shown in **fig. 7**.

Although the lack of gain at the intermediate frequency makes the possibility of adding agc a bit difficult, this simplified distribution does have its virtues. The main advantage is that product detection occurs at fairly low signal levels. This minimizes the noise modulation effects which often occur in product detectors. Another problem which is avoided is the effect of bfo leakage. Often, stray bfo energy finds its way into a high gain i-f amplifier, causing both intermodulation and noise modulation to occur. This is avoided in a system of this kind. The overall result is a receiver which sounds exceptionally crisp and clean, a virtue usually limited to direct-conversion

*KVG crystal filters are available from Spectrum International, Box 1084, Concord, Massachusetts 01742.

receivers and some well designed superhets.

transmitter mixer

Shown in **fig. 5** is the transmitting mixer and carrier oscillator for the transceiver. Although it would be possible to replace the carrier oscillator with some energy from the bfo, this would necessitate the introduction of some offset of the main system vfo during transmit. The use of a separate crystal oscillator was considered to be the simpler solution.

The transmitting mixer itself uses another Motorola MC1496G doubly-balanced modulator IC. This chip is ideally suited for this application due to the excellent balance available. This significantly reduces the amplitudes of many of the spurious products in the output from those which would appear in the output of a single-ended mixer. I have also used the RCA CA3028A and diode rings in this application, although the Motorola chip is by far the more satisfactory and is easily applied.

Balance is maintained in the mixer by using a center-tapped tuned circuit in the output. This is realized by putting a bifilar winding of 15 turns on a toroid core and tuning the series combination. The required power-supply voltage is injected on the center tap and output is extracted from the tuned circuit with a 3-turn link. An alternate output network would be realized by shunt-feeding the MC1496G output collectors with rf chokes. Then, the high-impedance end of a tuned circuit could be lightly coupled to one of the output terminals. This method would have the advantage of being more easily bandswitched.

The input levels feeding the mixer IC are not extremely critical, although severe overdriving should be avoided since this will cause some deterioration in the rejection of spurious output products. In the mixer shown, the carrier port is driven

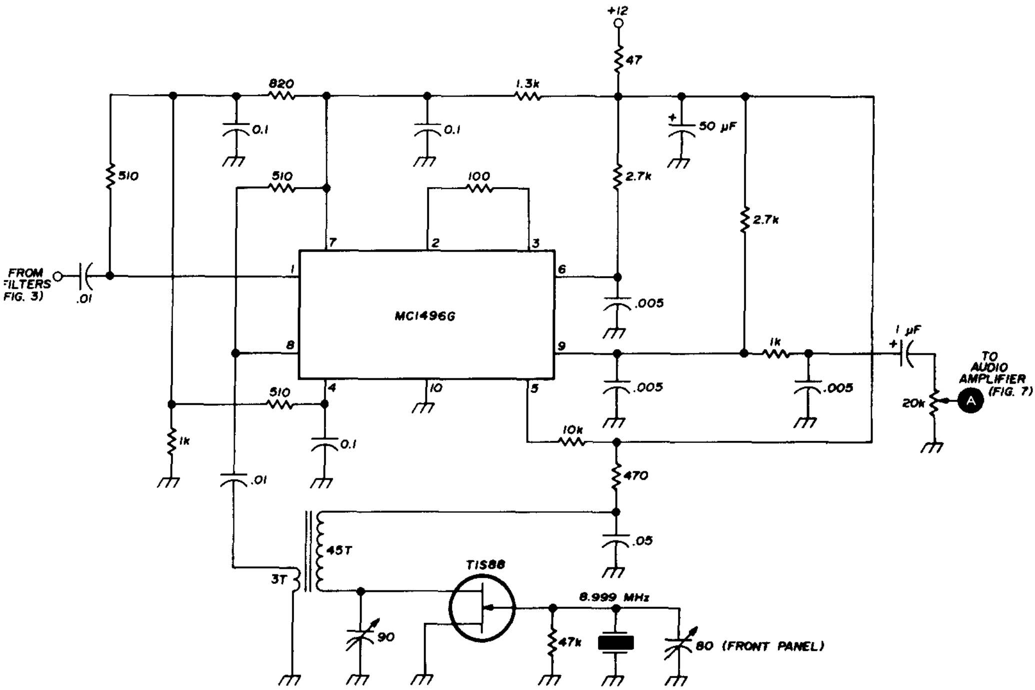


fig. 4. Product detector and bfo for the receiver used in the QRP transceiver. Transformer is wound on Amidon T-50-6 toroid or similar.

with a little over 0.5 volt, rms, and the signal port has about 200 millivolts of drive. During receive periods, the supply voltage is removed from the mixer/carrier-oscillator board.

transmitter power chain

The three-stage power amplifier which completes the transmitter is shown in **fig. 7**. The 2N3904 pre-driver and the 2N5859 driver are keyed. The final amplifier also uses a 2N5859. This Motorola TO-5 device is an excellent, general-purpose QRP device with a price tag of under one dollar. Since it has an f_T of around 250 MHz, it should be usable up as high as the six-meter band, with power output up to about two watts.

Both the pre-driver and the driver stages are operated as class-A amplifiers with emitter degeneration. There are a couple of reasons for this, even though sufficient gain might be provided by a single stage. First, I have found that it is usually worthwhile to add an additional stage with a decreased gain-per-stage to insure stability. The cost increase is minimal, but the unconditional stability is quite assuring.

Class-A operation has the advantage of preserving linearity. This is somewhat important, even in a CW rig. In previous transmitters it was noted that the spurious components in the output would be increased if an amplifier was allowed to saturate. The reason is that saturation would occur for the primary driving frequency, but the amplifier would still behave in a fairly linear fashion for spurious products. The net effect was that much of the filtering prior to the amplifier was negated.

The bandwidth of the amplifier strip of **fig. 6** is rather restricted. With the

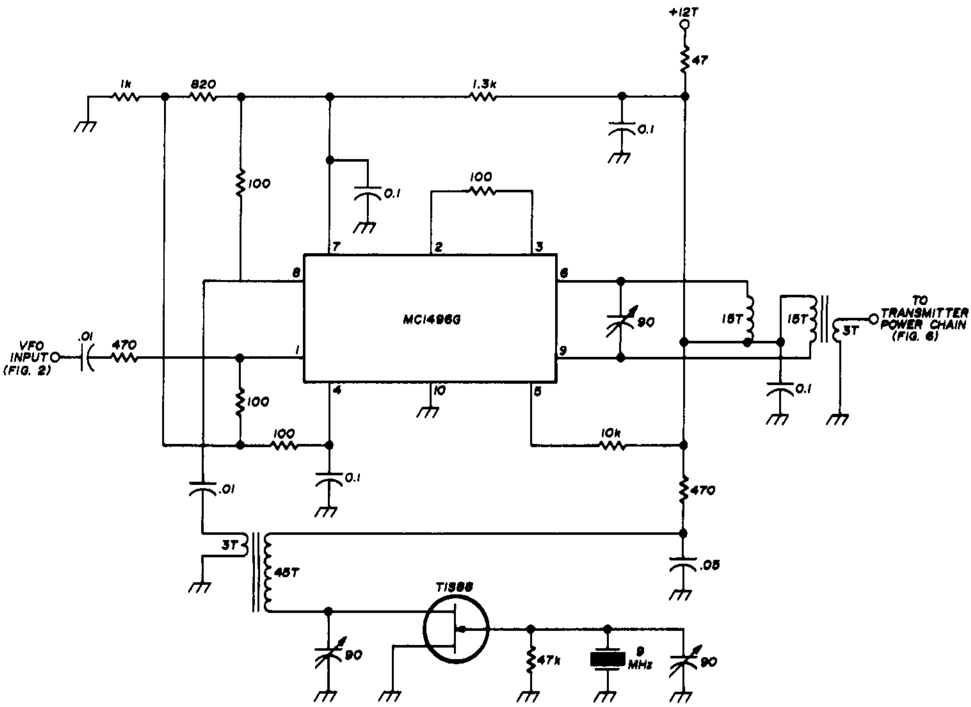


fig. 5. Transmit mixer and carrier-oscillator circuits. The frequency of the 9-MHz crystal oscillator is adjusted to the center of the 9-MHz i-f with the crystal filter (fig. 3) switched in. Transformers are wound on Amidon T-50-6 toroids or similar.

system aligned at 14.065 MHz, the commonly used QRP frequency on 20 meters, the output was down by about 30 dB when the vfo was tuned to the middle of the phone band. Most of the output power could be obtained, however, by retuning the output of the pre-driver. If operation of the transmitter over the total 20-meter band is required, the builder should provide for front-panel relaxation oscillator using a programmable unijunction transistor (PUT). The sawtooth output is attenuated and applied to the input of the audio amplifier. Transmitter keying is accomplished with a series switch using a 2N4036 silicon TO-5 transistor. Almost any silicon pnp device can be used in this slot.

In the circuits discussed above, the power supplies on the schematics have

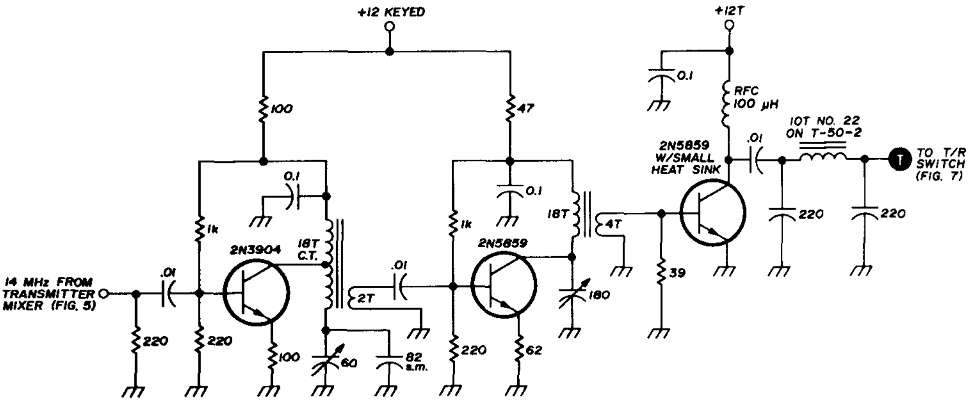


fig. 6. Transmitter power chain provides 1.5-watt output into a 50-ohm load.

adjustment of this tuned circuit. As shown, the system is flat over the CW portion of the band.

The output power of the transmitter was measured at 1.5 watt into a 50-ohm load. When the output of the transmitter board was investigated with a lab-quality spectrum analyzer, it was found that the second harmonic was down by only 27 dB. However, all other spurious outputs were over 60 dB down. Additional harmonic rejection is easily obtained with the half-wave filter at the transceiver antenna terminal. With two class-A stages being keyed, the backwave was more than 70 dB below the usual output.

control circuits

Shown in **fig. 7** is the final board which completes the transceiver. This board contains not only the audio system mentioned earlier but the sidetone oscillator, a keying transistor and the T/R circuitry. The sidetone is obtained from a simple been labeled as +12V, +12R or +12T. All circuits labeled with +12V have voltage applied at all times. However, those with the R or T suffix have power applied only during receive or transmit intervals, respectively. The two voltages are derived from the antenna relay, K1. I used a dpdt relay with an 800-ohm, 12-volt coil, with the second set of contacts switching the antenna.

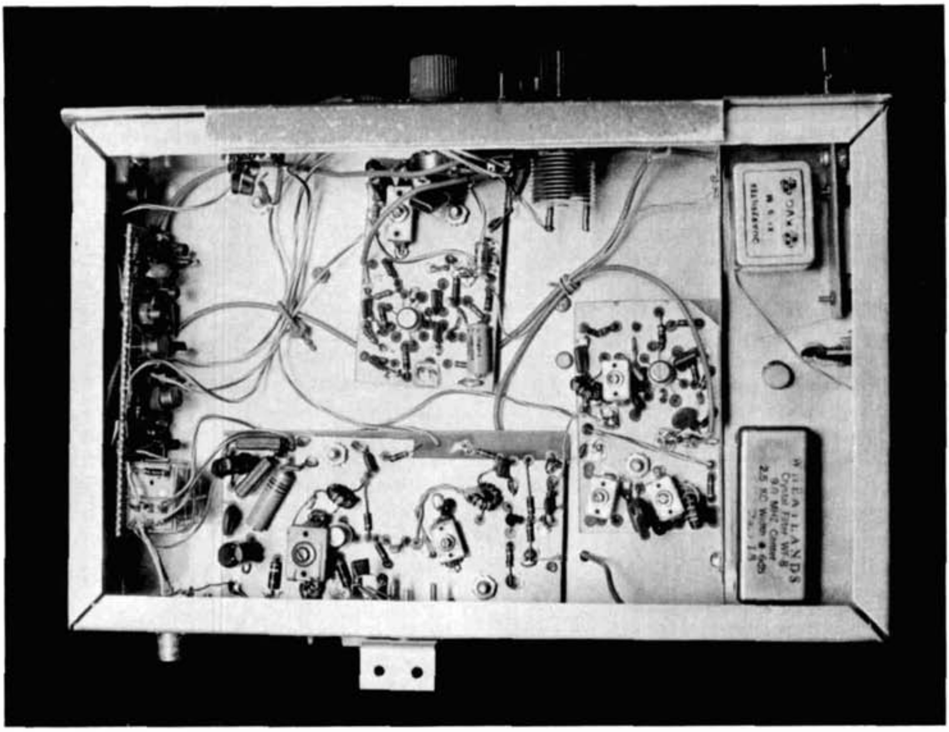
You may have noted that the product detector has power applied at all times, even during transmit intervals. Initially, the detector was muted during transmit. However, an objectionable click appeared when the receiver came on again. This was the result of the large current surge in the 50-μF decoupling capacitor used on the detector board.

The transmit-receiver logic is based upon a μA741C operational amplifier IC used as a differential comparitor. Under key-up conditions, the 2.2-μF capacitor is fully charged. However, when the key is closed, this capacitor is discharged. When this happens, the voltage on the inverting input of the op amp (pin 2) drops below half of the power supply potential. This causes the op amp output to switch to a high state, near the 12-volt supply, and saturates the 2N3904 relay driver. When the key is released again, the timing capacitor begins to charge toward the 12-volt supply through the 220-kilohm resistor. When the capacitor passes the 6-volt point, the op amp changes state again and the relay opens. The hold-in time can be varied from the 0.5 second I used by changing the 220-kilohm timing resistor.

There are two keying inputs shown. One is for the usual hand key. The other uses a *two-circuit* phone jack, with +12 volts available on the second pin to supply power to an external electronic keyer. This keyer[6] is based upon differential comparitor logic similar to that used for the T/R control.

additional thoughts

The transceiver described and shown in the photographs was packaged in an LMB type CO-2 cabinet. The extra holes shown in the chassis are a reminder of an earlier receiver which resided in the same enclosure. Although there is an abundance of extra room left, this will eventually be used for a variety of accessories, including equipment for at least one vhf band.



There's plenty of room under the chassis. Crystal filters are at right, next to transmit mixer and carrier oscillator. Transmit power chain is at bottom with control, keying and audio circuits to the left.

An obvious extension of this design would be the addition of other bands. Of the possible combinations, the easiest would be a 20- and 80-meter CW transceiver. The receiver can be made operative on 80 meters merely by switching the tuned circuits in the front end.

The transmitter mixer is moved to 80 meters by changing the output network in the transmitting mixer as described earlier. The output of the mixer should then be well filtered for the band in use. In this case, a lowpass filter would be suitable for 3.5 MHz with a bandpass filter being switched in for 14-MHz operation.

The transmitter power chain could be replaced with a broadband design with appropriate lowpass filters switched into

beyond the relatively simple systems considered in this article.

The performance of this transceiver has been more than satisfactory. Using only a ground-plane antenna, contacts have been made all over the United States and Canada as well as with VK, JA, UAØ and DM. While the unit does not represent the absolute ultimate in simplicity, the superior performance is well worth the minimal extra effort and expense.

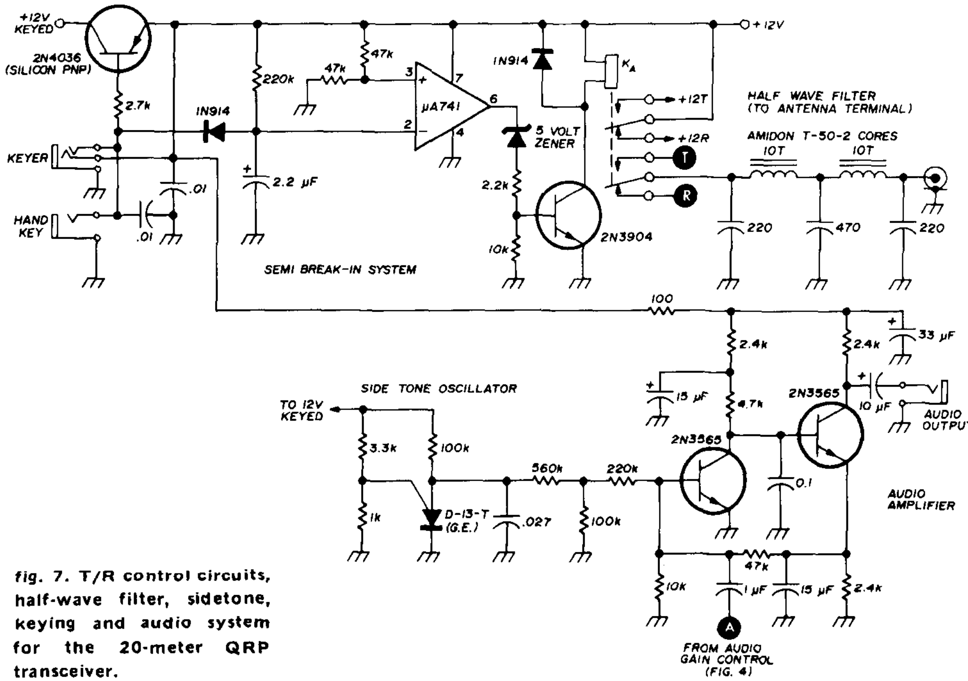


fig. 7. T/R control circuits, half-wave filter, sidetone, keying and audio system for the 20-meter QRP transceiver.

the output. If such an approach is taken, it is important that the filtering between the mixer and power amplifier be sufficient to provide an output free of spurious products. In most cases, a two- or three-section bandpass filter will do the job.

If bands other than 80 and 20 meters are considered, it will be necessary to derive other local-oscillator signals. This could be achieved either by bandswitching the vfo or by premixing the existing 5-MHz signal. Other, more elaborate synthesis techniques would yield superb performance, but they go

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ham radio

scattering characteristics of artificial radio aurora

Artificial aurora, created at will by high-power, high-frequency transmitters, may prove useful for long-distance vhf communications

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The U.S. Department of Defense has recently disclosed that it is now possible for man to create his own artificial radio aurora (ARA) capable of scattering radio waves of frequencies up to 450 MHz. The ARA can be produced in either the ionospheric E- or F-regions over a source of very powerful high-frequency radio waves directed upwards. Don't, however, expect to see this "aurora." The optical effects of ionospheric modification are very weak, corresponding to less than a 30% change in the level of some characteristic lines in the night airglow spectrum.[1]

Research has taken place at three sites. One,[2] near Platteville, Colorado, operated by the Institute for Telecommunications Sciences (ITS), uses a transmitting system with an effective radiated power of 40 megawatts. Two transmitting arrays are used; one, for the frequency range 2.5 to 5 MHz is composed of five crossed dipoles, and the other, usable between 5 and 10 MHz, uses ten crossed dipoles. The crossed-dipole elements are fed in phase quadrature to produce circular polarization. It has been found that the ordinary magnetoionic component[3] produced by right-hand circular polarization (in the northern hemisphere) produces the strongest ARA. The second ionospheric heating facility at Arecibo, Puerto Rico, operated by Cornell University, uses a transmitter with one tenth the power output and an antenna with 10 dB more gain (the 1000-foot dish). A third facility, located at Gorki, 400 km east of Moscow, USSR, uses a transmitter with 60-kW average power and 22-dB antenna gain.[4]

In the ionosphere, just below the reflection height, the high-frequency wave is retarded and energy is imparted to the electron gas, raising its temperature. It is forced to expand along the magnetic field lines, creating field-aligned ionization irregularities. Although the fractional change in ionization density through these irregularities may be only a percent or so, each irregularity scatters coherently. Many of these long, thin irregularities are capable of scattering high-frequency and vhf signals just as the natural aurora does. The primary differences as far as the radio amateur is concerned are that *this is a very localized* aurora, it may be turned on and off within seconds, and its height is determined by the frequency of the heating transmitter.

Since the scattering region is localized and the Doppler spreading is not great, voice signals reflected by ARA are still intelligible although there is rapid and deep flutter fading.

A companion article appearing in *QST*[5] this month describes ARA sounding and communications experiments and the directional properties of the scatterers. Since I will go into further detail here, it may be to the reader's advantage to read the *QST* article first.

F-region scatter observed

Soon after the announcement of the triggering of spread F by the Platteville facility,[6] experimenters noted echoes on high-frequency sounding paths that were determined to be due to reflection from field-aligned irregularities from the F-region over Platteville. **Fig. 1** shows a chart recording made in March, 1971, at an SRI field site in central California of experimental transmissions from a station in Arkansas. The frequency was chosen on the basis of oblique soundings to be a few MHz above the direct path MOF. A clear cause-and-effect relationship is shown, with the 20-MHz signal fading into the noise some 20 seconds after turnoff of the ionospheric modification transmitter. A-m voice modulation on this signal was quite readable.

The fading rate and the lengths of time that the scattered signal remained after turnoff of the heating transmitter varied with the frequency and with the scattering angle. The lower frequencies and the forward scatter paths had lower fading rates and were more persistent than higher frequencies and backscatter paths.

High-frequency soundings had indicated that even higher frequencies might be usable, so equipment for receiving TV video carriers was set up at another field site near Bakersfield, California. The receiving system used an array of eight commercial LPA antennas (**fig. 2**), crystal-controlled converters and narrow-band receivers with recording on chart and tape. **Fig. 3** shows chart recordings of some of the signals observed while the Platteville transmitter was being operated on a one-minute on, one-minute off cycle.

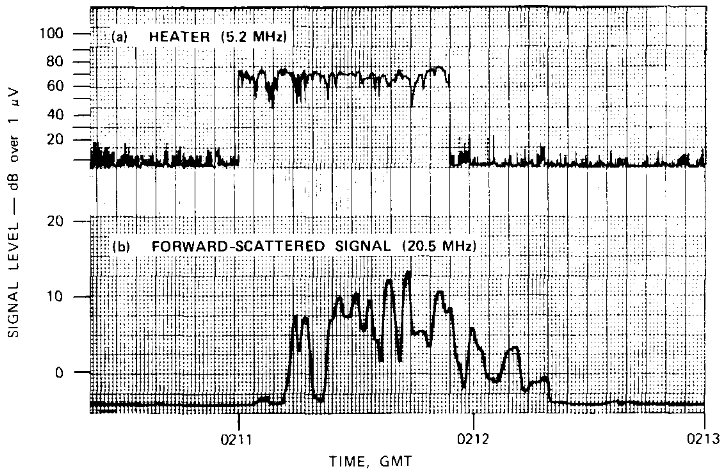

fig. 1. Field-strength recordings made in *central California* of the ITS ionospheric modification transmissions (A) and a transmission from Arkansas at a frequency above the direct path MOF (B).

It is difficult to identify a television station solely on the basis of its video carrier frequency.* Signals were never strong enough to demodulate and, in *addition, too much co-channel interfer-*ence existed from stations in northern and southern California. Based on ray-

*In the United States, Canada and Mexico, vhf TV stations are assigned video carrier frequencies of either 1.24, 1.25 or 1.26 MHz above the lower edge of the channel. These are referred to as minus-, zero- or plus-offset, respectively.

tracing calculations, these TV signals are believed to have originated from stations in northern Mexico and southwest Texas.

E-region scatter

One of the most recent developments in ionospheric modification has been the observation of artificial radio aurora in the E-region over Platteville, Colorado. This development was made possible by the addition of five low-frequency crossed dipoles in an array at Platteville, which allowed operation at frequencies as low as 2.8 MHz. These frequencies are reflected from the E-layer during the daytime.

Actually, most amateur experience with natural field-aligned scattering is that from the E-region, since the aurora rarely penetrates far enough south for echoes to be obtained from the F-region. **Fig. 4** shows the intersection with the earth of scattering cones of various angles to the magnetic field at a height of 110 km over Platteville, Colorado. (A similar picture for F-region reflectors is shown in the companion *QST* article.) Specular scattering may take place between stations located on intersections of supplementary cone angles (their sum equals 180°). These curves were derived for a single field-aligned scatterer. Actually, such scatterers are present in a scattering region of 100-km diameter and 10-km thickness. This distribution will spread the coverage by a similar distance. Soundings were made over a path from a transmitter site south of Albuquerque, New Mexico, to an ITS receiving site near Haswell, Colorado, where there is a 60-foot parabolic dish antenna (**fig. 5**). One of the sweep frequency soundings taken over this path, **fig. 6**, shows time delay, frequency coverage and radar cross section versus frequency. The sounding transmitter was turned off during those portions of the frequency sweep that fell within TV channels 4, 5 and 7.

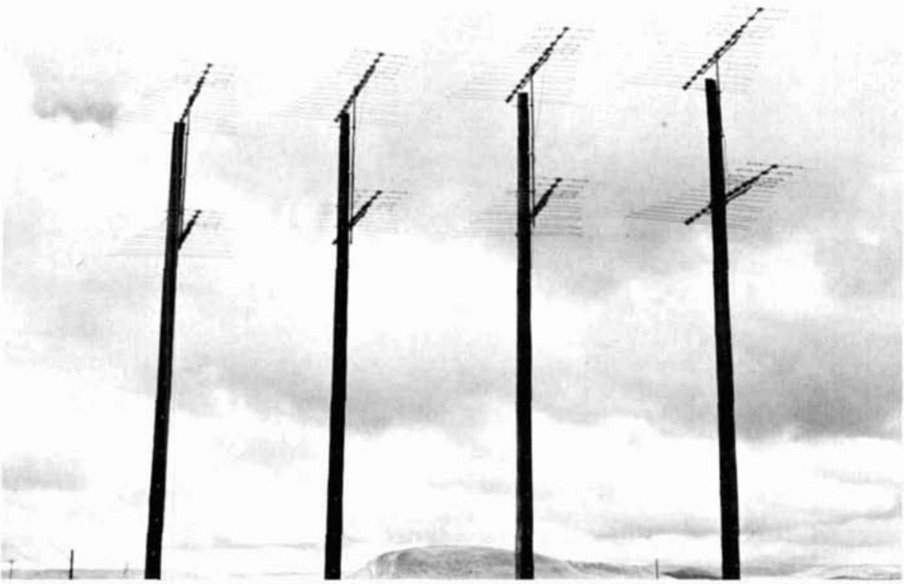

fig. 2. Array of log-periodic TV antennas used at site near Bakersfield, California, to monitor TV video carrier signals scattered by ARA.

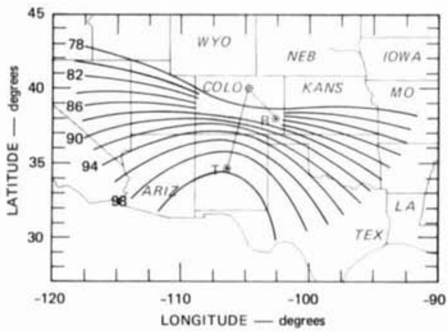

fig. 4. Contours of intersections with the earth of scattering cones from field-aligned scatterer at 110 km over Platteville, Colorado. Communication can be established between stations located on contours of supplementary aspect angles.

Radar cross section may be a new concept to some of you. It is here defined by the radar formula

$$\sigma = \frac{P_r\ (4\pi)^3\ r_1{}^2 r_2{}^2}{P_t G_t G_r \lambda^2}$$

where P_t is the transmitted power output, P_r is the received power input, λ is the wavelength and r_1 and r_2 are the distances from transmitter and receiver to target, respectively, in meters, and G_t and G_r are the transmitting and receiving antenna power gains over an isotropic radiator. This value of received power, P_r, must compete with the noise power, P_n which is

$$P_n = FkTB$$

where F is the ratio of cosmic noise power to that produced by a termination (Johnson noise), k is Boltzmann's constant (1.38×10^{-23}), T is the absolute temperature in degrees Kelvin, and B is the receiver bandwidth in Hz. To spare the reader the exercise of going through the numbers, I have calculated the minimum detectable cross section for an amateur CW station in various amateur bands having the following characteristics:

P_t = 500 watts
r_1 = r_2 = 10^6 meters
B = 100 Hz
P_r/P_n = 1
G_t = G_r = 10 (10 dB) 28 MHz
= 20 (13 dB) 50 MHz
= 40 (16 dB) 144, 220 MHz
= 60 (18 dB) 432 MHz

The minimum observable cross sections are about 10^3 to 10^4 square meters (30 dBsm to 40 dBsm). These are shown in **fig. 7** along with the range of cross sections observed from the Platteville heated region under ideal conditions for both E- and F-region heating. You can see

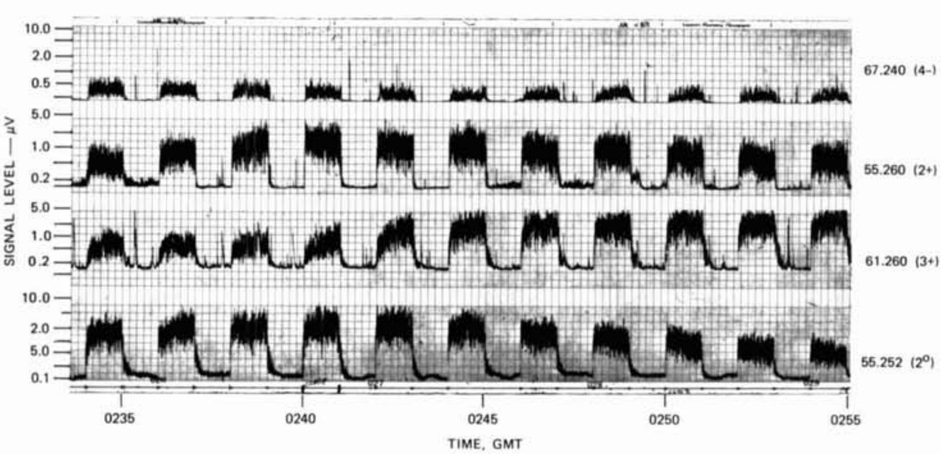

fig. 3. Field-strength recordings of TV signals received during 1-minute on, 1-minute off cycling of ionospheric modification transmitter.

that detection at 50 and 144 MHz is well within the capabilities of the advanced amateur station.

It is quite possible that some amateurs may have already communicated two-way via ARA without realizing they were doing so. During the last test series at Platteville, about 264 hours of heating experiments were conducted (between September 10 and November 2, 1973). Of these, 47 were in the prime evening time interval, 0000-0600 UT. It is hoped that radio amateurs will have a chance to participate in future tests presently scheduled for Platteville, Colorado, some time during the first half of 1975 and for Arecibo, Puerto Rico, during the first weeks of April, 1975.

ionospheric modification at arecibo

Since the Arecibo heating antenna beam is narrower than that at Platteville, the expected cross sections are some 10 dB lower. Also, since Arecibo is not licensed at present for frequencies below 5 MHz, E-region heating is not possible and F-region heating is possible only during daylight and early evening hours. **Fig. 8** shows a map of intersections with the earth of scattering cones from a field-aligned scatterer at 300-km height over Arecibo, Puerto Rico.

Most likely to benefit from Arecibo ARA are the Caribbean, Venezuela, Columbia and possibly Central America—areas with very little amateur vhf activity.

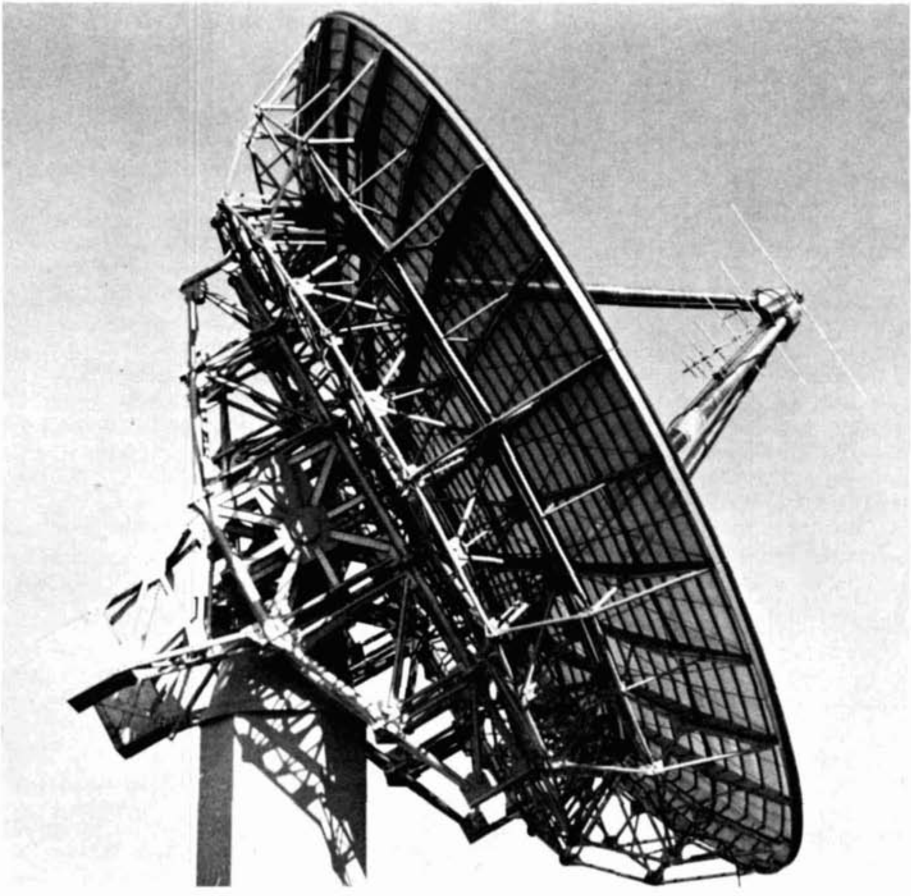

fig. 5. Sixty-foot dish and LPA feed at ITS field site near Haswell, Colorado.

Perhaps some vhf DXpeditions would be in order. These are good areas, and April is a good time of year for transequatorial scatter.

other possibilities

If you are not fortunate enough to be in the ARA coverage areas of Platteville or Arecibo, do not despair. Much remains to be learned about what is really happening during ionospheric modification. There is another form of scattering caused by plasma and ion-acoustic waves[7,8] which, though very weak, may be usable for stations within line-of-site of the ARA. Amateurs, being spread all over the country, are in a unique position to investigate paths not previously available to researchers.

In addition, much of what we have learned during these experiments will be of use to the radio amateur interested in scatter and auroral communication. Field-aligned ionization in nature? Yes, even at mid-latitudes, if you know how to look for it.[9] Further information on coverage areas for scattering by the natural aurora may be found in an article by R.L. Leadabrand.[10] I am convinced that the so-called *X-mode* (50-MHz signals back-scattered from sporadic E-patches) is due to field-aligned ionization in those patches.[11] And transequatorial propaga-

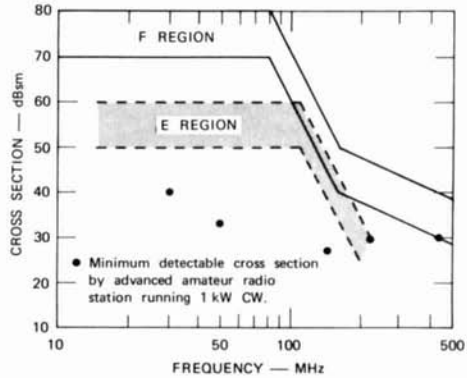

fig. 7. Radar cross-section of the Platteville ARA versus frequency for E- and F-layer echoes.

tion is also probably due to forward scattering from field-aligned irregularities in the equatorial F-region.[12] In addition, experimenters have found that meteor-trail debris becomes field-aligned.[13,14]

Ranging capability, something lacking in most radio amateur stations, would prove invaluable in identifying these propagation modes. It is possible some may have been used for decades without proper identification. Ranging capability implies pulse or linear sweep-frequency CW operation, neither of which is in amateur use in the vhf bands. But here, perhaps, is an opportunity for the radio amateur to use some of those radar

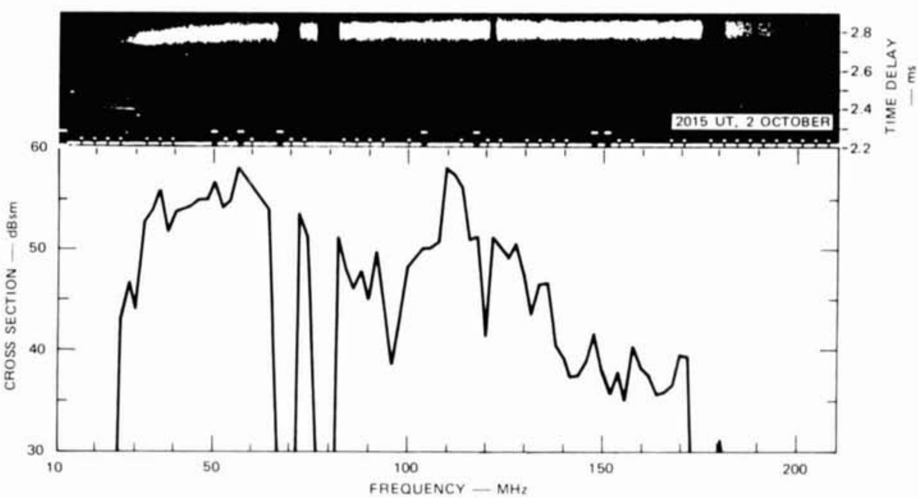

fig. 6. Sweep-frequency sounding received at Haswell, Colorado, showing echo from E-region ARA.

signals in the 220- and 420-MHz bands for scientific investigation.

acknowledgements

I wish to thank W.F. Utlaut and his associates at the Institute of Telecommunications Sciences, Boulder, Colorado, for their cooperation in the conduct of these experiments and for graciously allowing us to use their Haswell facility. Acknowledgment should also be made to the other organizations that were part of the team that explored the scattering properties of ARA. They are Raytheon Corporation, Sudbury, Massachusetts; Riverside Research Institute, New York City, New York; Barry Research, Palo Alto, California; and the Aeronomy Corporation, Champaign, Illinois. This research was sponsored by the Defense Advanced Research Projects Agency through the Office of Naval Research.

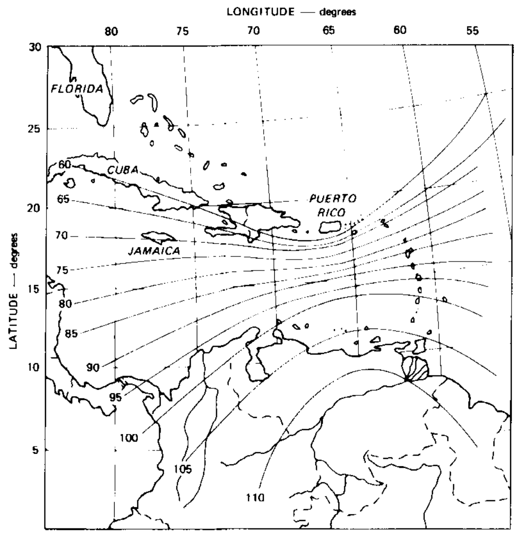

fig. 8. Contours of intersections with the earth of scattering cones from field-aligned scatterers at 300 km over Arecibo, Puerto Rico. The contour values are the angles between the cones and the geomagnetic field.

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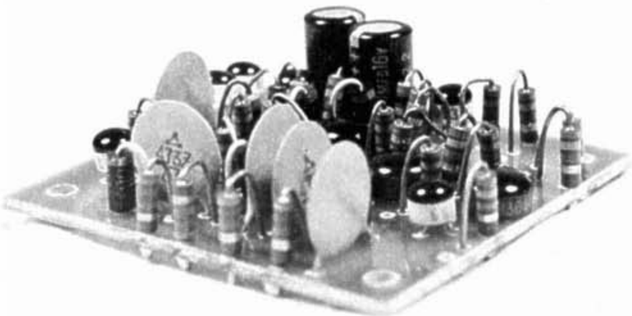

improved channel scanner for vhf fm

A complete four-channel vhf fm scanner that can be built for ten dollars

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If someone were to tell you that an integrated-circuit receiver scanner could be improved by redesigning it with discrete components, you'd say they were crazy! But, forced with difficulties in obtaining TTL logic elements for the K2LZG scanner kit,[1] and having previous experience in simplifying communications circuitry by eliminating ICs, I redesigned the scanner circuit with discrete components. This produced many unexpected side benefits.

The scanner built with discrete components (**fig. 1**) offers the following advantages, in addition to availability and cost of components.

1. Simpler type of readout.

2. Ease of maintenance – troubleshooting is easier with discrete components and spare parts are no problem (almost any silicon transistors provide proper operation).

3. Unit operates directly from a 10- to 15-volt power supply. IC version requires a 5-volt regulator.

4. Less current drain – about 8 times less than the IC version with regulator.

5. Smaller – PC board layout is one-quarter to one-third the size of the IC version. Actual size of board is 2x2-1/8 inches (51x54 mm). This is primarily due to the flexibility of discrete components in crossover-free, small area layout.

6. Flexibility of design – unit can be adapted easily for less than the full four channels without having to skip channels between the active ones. The unit is also adaptable to any desired scan rate with a simple component change.

The new design also takes into account several variable factors which you may run into when applying the scanner to your particular receiver. The unit can be used with either positive or negative logic from your squelch circuit. That is, scanning can be stopped with either a positive or ground signal. The input circuit is very sensitive; therefore, almost any voltage, from a few volts up, may be used to operate the scan-stop circuit. The unit normally puts out +5 volts to turn on (*enable*) the oscillator. However, simple additional inverter transistors used with the oscillator turn-on circuits allow the scanner to operate with oscillator circuits requiring a ground signal to *enable*.

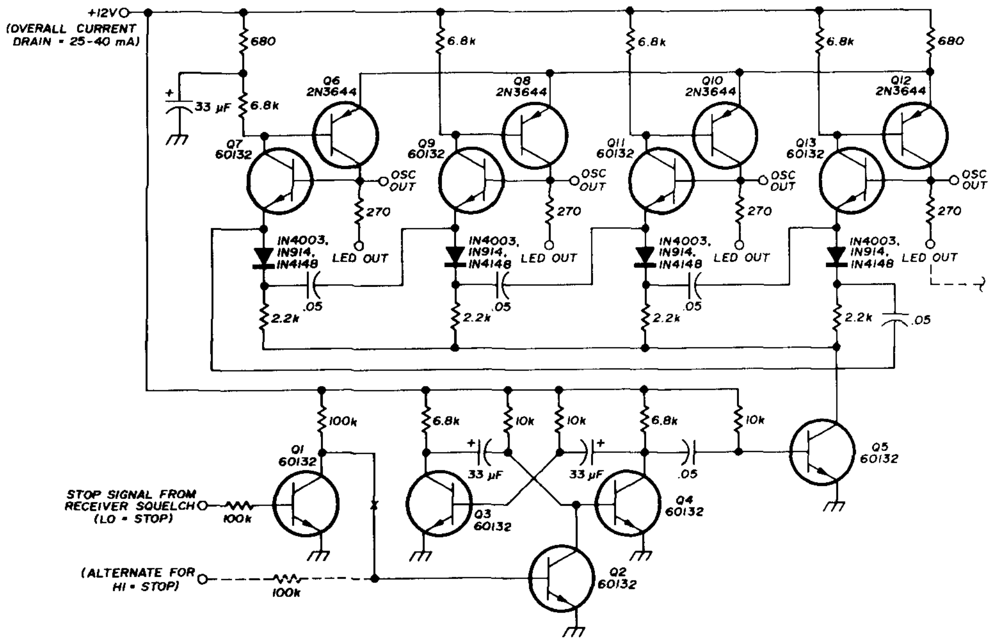

fig. 1. Schematic diagram of the improved fm receiver scanner. Additional channels may be added by connecting additional ring-counter stages, as discussed in the text.

typical test voltages, volts (ground reference)

Q1 collector	0.6/0
Q3/Q4 base	0/0.6
Q3/Q4 collector	0/12
Q5 base	0.6
Q5 collector	0.1
Q6/Q8/Q10/Q12 base	12/4
Q6/Q8/Q10/Q12 collector	0/5
Q7/Q9/Q11/Q13 emitter	0/3.8

theory of operation

Transistors Q3 and Q4 form an astable multivibrator, operating at a pulse repetition rate of approximately 10 pulses per second. Transistor Q2 turns *on* to stop the multivibrator when Q2's base is *high*. Q2 does this by shorting the base of Q4 to ground. Normally, squelch circuits have a *high* output available when closed and a *low* output when open. Therefore, inverter Q1 is used to provide the proper polarity signal to operate Q2. Transistor Q1 is tied to the collector of the squelch switch stage in the receiver through a 100k isolation resistor to turn on Q1 when the squelch is closed. When the squelch is open, Q1 is turned *off*, Q2 is turned *on*, and Q4 is locked *off*. This stops the sequencing action of the scanner. For those squelch circuits operating in the opposite sense, the alternate circuit connections shown on the schematic may be used. Transistor Q1 is deleted, and Q2's base is operated directly from the receiver.

Transistors Q6 to Q13 form a four-stage ring counter. Normally each stage is turned off by the base-emitter biasing on the pnp transistor. When power is first applied, the 33-μF capacitor in the biasing circuit for the base of Q6 ensures that the first stage will turn on by momentarily upsetting the normal bias. Subsequently, due to the return for ring counter stages being pulsed by Q5, the *high* output is passed from one stage to the next in an endless ring pattern. Normally Q5 is turned *on* hard to ground, and it is pulsed to a *high* condition for a short pulse duration each time a negative-going pulse is received from the multivibrator. When the pulsating action ceases due to operation of the squelch circuit, the ring counter stops stepping, holding the output of one ring counter stage *high* until the squelch circuit closes.

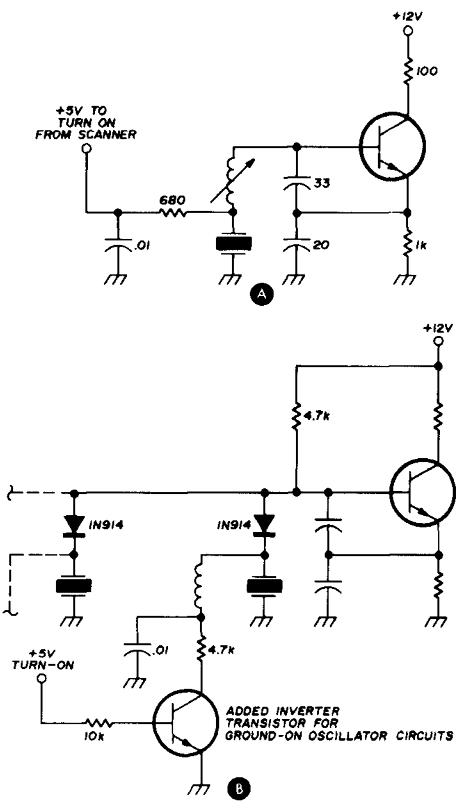

fig. 2. The receiver scanner shown in fig. 1 may be used with oscillators that require +5 volts for turn-on (A), or for oscillators requiring a ground for turn-on (B).

The collector of the pnp transistor in each stage provides a +5 volt output directly to the oscillator to be enabled and also through a 270-ohm current-limiting resistor to the LED readout corresponding to the enabled channel. It is important to note that the 270-ohm resistor in each stage must be connected to the LED for proper biasing to be maintained. If LEDs are not used, adjustments in the circuit or dummy resistor loads must be used to maintain the same current flow in other parts of each ring-counter stage.

The outputs of the scanner can be used to turn on oscillators in various ways, depending on the particular oscillator. Since this subject has been treated in several magazine articles,[1,2] only a few examples will be given. Refer to **fig. 2** for examples of two common types of oscillators which may be encountered. The examples illustrate, in general, how oscillators may be enabled by the scanner.

The first circuit **(fig. 2A)** shows a typical +5 volt turn-on type of oscillator. This circuit is used in a vhf fm receiver described in a previous article.[3] The oscillator has no internal bias; therefore, it cannot oscillate by itself. The base of the transistor, connected through a coil to the crystal and to the *enable* line through a resistor and 0.01-μF bypass capacitor, is biased on by +5 volts from an external source. The resistor and bypass capacitor block the flow of rf on the control line. Application of approximately +5 volts from a scanner circuit to the control line will turn the desired oscillator on.

The second circuit **(fig. 2B)** shows a conventional oscillator circuit adapted to multichannel operation by using diode switching. This circuit was commonly used a few years ago to modify older tube-type equipment for multichannel operation. In this case the crystal is essentially turned on at appropriate times by grounding the corresponding control line. In this case each leg is turned *on* when desired by a *ground* from an added inverter transistor which turns *on* when +5 volts is applied to its base from the scanner through an isolation resistor.

It can be seen from these examples and the referenced articles that almost any fm transceiver or receiver can be modified to provide scanning operation. Application of one or more of the ideas illustrated can be used to convert almost any oscillator circuit.

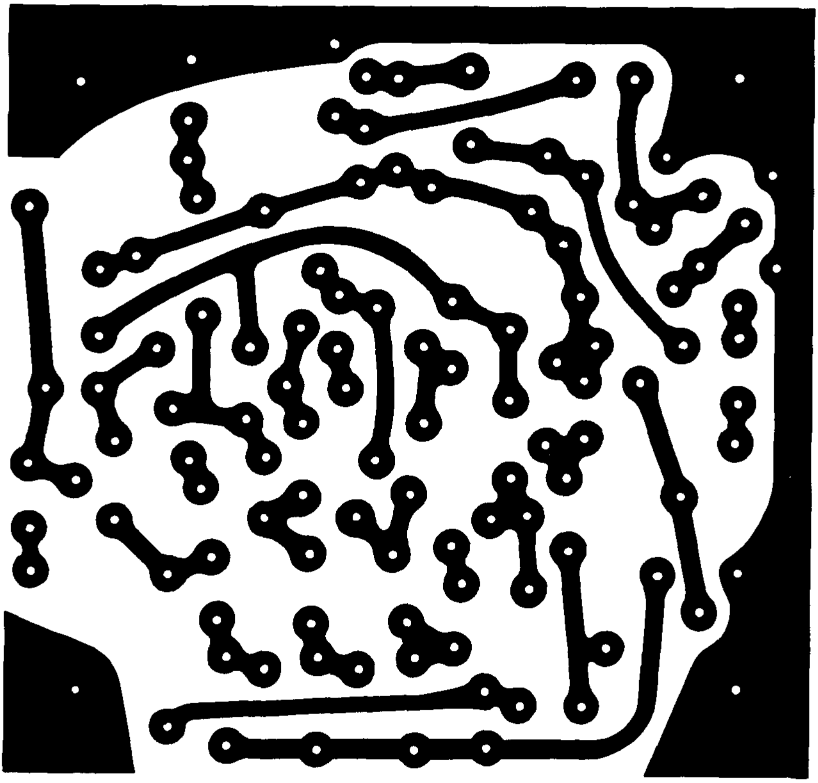

fig. 3. Printed-circuit layout for the improved fm receiver scanner. Component layout is shown in fig. 4.

scanner circuit modifications

There are at least four possible modifications which may be of interest to other system requirements, if necessary.

1. Less than four-channel scan. This may easily be accomplished by reconnecting the 0.05-μF capacitor at the bottom of the last ring-counter stage to be used. It should be rewired to the first stage instead of the next stage. This effectively shortens the ring. The Q6/Q7 stage must always be used since the bias circuit for the base of Q6 includes a capacitor and resistor which ensure that the ring counter operates when power is first applied. On the PC board shown in **fig. 3** this is best done by reconnecting the last 0.05-μF capacitor to the line along the bottom of the board running back to the first stage.

2. Changing scan speed. The speed was set up to be the optimum compromise between proper squelch triggering and ability to catch short transmissions. If the approximate 10 channel/second rate requires change for some reason, different values for the two 33-μF capacitors in the multivibrator will change the oscillation rate of that stage.

3. **Increasing number of channels.** If you can stand the confusion of scanning more than four channels, more than one board may be used. Link the ring counter stages on the two (or more) boards, and use only the triggering and timing circuits of the first board.

4. **Programming channels.** Switching schemes can be used between the scanner outputs and the various oscillators to disable one or more channels at certain times or to select particular channels to be monitored out of a bank of the available channels.

construction

Assembly is straightforward. If the etched circuit board shown in **fig. 3** is used, most of the resistors and all of the diodes are installed standing up on the pads darkened in on the layout drawing, **fig. 4**, and the top leads are looped back down to the pads which are circled. Polarity should be observed on the diodes and electrolytic capacitors. Install the 2N3644 transistors first to avoid any mixup later with the 60132 (general purpose npn) transistors.

Pads which are circled and have an X in them are inputs and outputs and will accept wire such as number-22 hookup wire, for connections to the outside world. The +12-volt terminal should be connected to a source of filtered +10 to +15 Vdc. The input terminal should be connected to the collector of the receiver's squelch switch stage. If the alternate input is used, Q1 and the 100k resistor from +12 volts to the base of Q2 can be eliminated and the 100k resistor from the input terminal can be connected

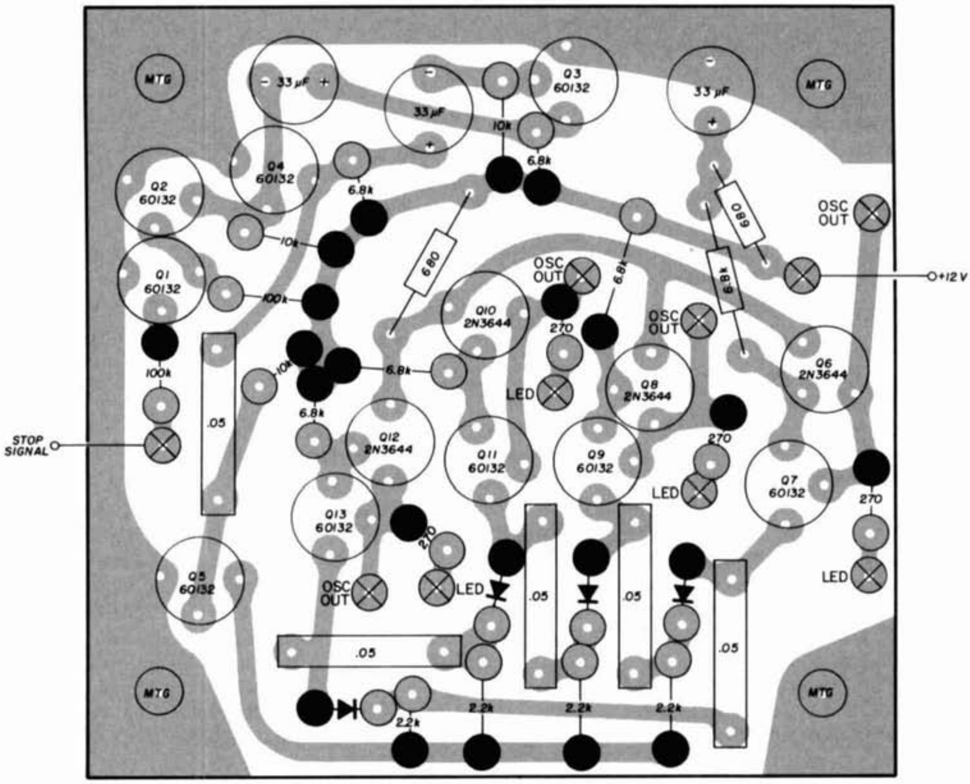

fig. 4. Component layout for the improved fm receiver scanner. Dark circles indicate bottom of resistors (or diodes) installed in vertical position; the other lead is looped over to the open circle.

directly to the base of Q2. The oscillator outputs should be connected to individual oscillator stages, and the LED outputs should be connected to the anodes of the LEDs.

The cathodes of the LEDs should be connected to ground. Using the type of LED supplied in the kit,* the orange dot on the base indicates the cathode terminal. Be sure you observe polarity or the LED won't illuminate. The LEDs supplied with the kit may be mounted by inserting them in 13/64-inch (5-mm) holes drilled in the front panel of the radio. A spot or two of epoxy cement on the base of each LED behind the panel should hold them in place. Alternate LEDs sometimes have one short lead to indicate the cathode. Likewise, alternate 1N4148 diodes used in ring counters have their cathodes identified by a shorter lead.

If you wish to use both the scanner and a regular channel switch, this may be done by first connecting the oscillators to operate with the scanner and then wiring up a rotary channel switch to supply +5 volts to select oscillators when the scanner is turned off. A five-position switch can be used for this purpose with the fifth position turning on the scanner. In positions 1 through 4, the scanner is turned off, and +5 volts is connected through the switch to the oscillators corresponding to the selected channel positions on the switch.

references

1. Bob Reifsteck, K2ZLG, "VHF Receiver Scanner," *ham radio*, February, 1973, page 22.
2. George Allen, W2FPP, "VHF FM Channel Scanner," *ham radio*, August, 1971, page 29.
3. Jerry Vogt, WA2GCF, "VHF FM Receiver," *ham radio*, November, 1972, page 6.

ham radio

*In conjunction with this article a kit is available, complete with undrilled G-10 PC board, all components, the LED indicators and an instruction manual. Price of the kit is $10, including domestic parcel post. If desired, add 50¢ for a number-66 drill bit or 40¢ for air-mail delivery. To order, or to obtain information on other fm kits, write to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

how to measure peak envelope power

Don't depend on your meters for *accurate* measurement of PEP input power — use the technique discussed here

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It's unfortunate that the FCC requires amateur radio stations to compute power input for logging purposes. This is simple for fm or CW or even the almost-obsolete "conventional a-m" (6A3). For the user of single sideband (3A3J), however, the undertaking is far from simple. The purpose of this article is to identify those aspects that contribute to the difficulty of measurement and to present methods whereby the measurement may be made with a reasonable degree of accuracy.

Because the input power to an ssb transmitter continually varies when transmitting voice signals, custom and law stipulates that peak input power must be that which is measured. How do you catch that fleeting, almost ethereal peak of power induced by the upper excursion of a complex waveform? No meter can follow it without unpredictable lag or overshoot. Even if it could, what eye could register a needle flick of less than a thousandth of a second? Being knowledgeable of the instantaneous plate current, assuming you're using a vacuum tube in the metered stage, is, by itself, a formidable undertaking. But power computation also requires knowledge of instantaneous plate voltage. Unless the transmitter's power supply has superb dynamic regulation, far better than you have any right to assume, you're faced with another (but not quite so formidable) task of super-quick information gathering.

We've coasted along for years with the possibly-right, probably-wrong assumption that the peak needle excursion of a plate current meter equals one-half of the actual peak current. The FCC gives a quasi blessing to this assumption by accepting it, but only if the meter is that supplied by the original equipment manufacturer. It's highly doubtful that any FCC engineer has illusions regarding the accuracy of this method of ascertaining peak plate current. It's accepted only because not accepting it would rule off nearly every ssb transmitter in the Amateur Radio Service.

Unless the input power is within 10% of the maximum permitted, the FCC will also accept the nominal *rated* voltage of the plate power supply.

Now, let's review what we have for making our educated guess as to the peak input power of a voice-modulated ssb transmitter:

1. A plate meter indication that is dependent upon the dampening characteristics of the meter as well as upon the voice characteristics of the speaker. These voice characteristics vary not only from person to person, but also as to type of speech used by the individual. If you talk close to the microphone, with the gain turned up, using a monotone normal to confidential conversation, the meter will give one type of deflection. On the other hand, if you're back from the microphone, talking in a quick, excited tone, the deflection will be quite different for a given *peak* excursion.

2. The plate voltage may be measured, or it may be taken from the manufacturer's specifications. If measured, it may have been under *no load* condition. If under loaded conditions, the fall-off under syllabic current loads may possibly be gradual enough to be read on the voltmeter. Unless a huge filter capacitor is used, it's probable that voice-peaks markedly load down the voltage.

The probabilities of truly accurate measurement of peak input power are about the same as those of rolling a seven ten times running at a Las Vagas gambling table! How, then, can you make an accurate measurement?

Assuming, for a start, that you have a child-like trust in the utter trueness of the stated plate voltage, I'll delve into the matter of plate current measurement. Let's hope your transmitter permits an easy access to its negative high-voltage lead so that this lead may be broken for the insertion of a small resistor. Otherwise you'll need, in addition to the normal equipment, an isolation transformer (good for quite high voltage), a well-insulated oscilloscope cart and the type of constitution that permits playing Russian roulette!

But let's say your transmitter is metered in the negative lead, as are most modern designs. You break the lead at a point that does not present any unintentional shunt paths, insert a resistor of a few ohms, connect an oscilloscope across this resistor, and then calibrate the scope in terms of current. It's best to use a scope with a long-persistence CRT. This lets you take a deliberate view of a trace that flicks across the face in less than a thousandth of a second. Now, as you speak into the microphone, you'll be able to see the maximum current drawn, reading it off the calibrated oscilloscope graticule.

For a better idea of your peak input power (if you can scrape up another scope), you can build a resistive voltage divider across the output of your power supply, tapping off a little potential for your scope input. Here, again, you have a simple calibration job to do. With two measuring devices which are not burdened by inertia and yet hold a reading (peak for current, dip for voltage) long enough to permit accurate observation, you're all set to read peak input power. (Provided that you're using a common-cathode triode operating in class A or class AB1.)

Have you noticed the way the FCC is currently grading its operator examinations? Power input to a vacuum tube no longer is measured by just its plate power input. You must also consider the power fed into the screen grid (if used), the suppressor grid (if used) and the control grid! That last is rf power. The others are dc, of course.

Most high-power rf amplifiers use common-grid triode vacuum tubes (also called grounded-grid). If yours does, coming up with the total power input is comparatively easy. All you need to do is to first measure the vswr on the line between the exciter and final. Make it unity. Then measure the instantaneous rf voltage on it with another calibrated oscilloscope. Compute the rf power by E^2/R, where R is the cable impedance. Add this power to the measured dc input power, and you should have fully satisfied all the requirements for ascertaining the peak input power of your ssb transmitter!

ham radio

how to predict harmonic output

Determining drive points for optimum harmonic generation

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Why are certain oscillator circuits better suited for frequency doubling than others? Why are other circuits more suitable for frequency tripling? Questions like these have bothered many amateurs and experimenters. It is the aim of this article to shed some light on the subject of the harmonic content of some common non-sinusoidal waveforms.

A signal of pure sine waves contains only its fundamental frequency. Any departure from this sine-wave pattern, no matter how small, is due to the presence of additional frequencies that are multiples (harmonics) of the fundamental frequency. If this non-sinusoidal wave can be given an exact mathematical description, the amplitude and phase of each harmonic frequency it contains may be calculated.

analysis

For example, assume that we have a series of waves, all alike, each looking like that pictured in **fig. 1**. This is a sine wave

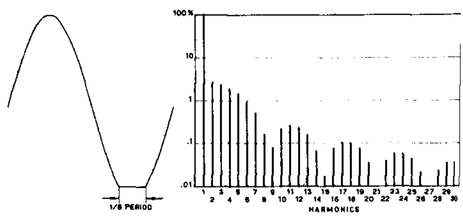


fig. 1. Sine wave with a small amount of negative peak clipping. Graph shows relative amplitudes of the resulting harmonic products.

with the tip of the lower loop sliced off. Such a waveform might be produced when an amplifier is driven slightly beyond the cutoff point.

A Fourier analysis will show that this deformation of the sine wave gives rise to an endless number of harmonics, with amplitudes that tend to decrease with growing frequency. However, this tendency is not uniform.

The chart in **fig. 1** illustrates the relative amplitude of each harmonic frequency up to the 30th. The scale is logarithmic, and the columns show the amplitudes of the harmonics as a percent of the original waveform's amplitude.

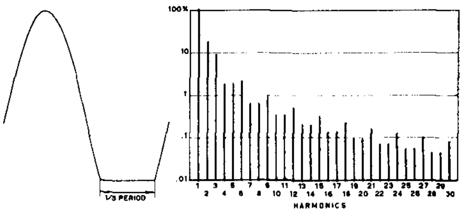


fig. 2. Harmonics generated when the lower portion of a sine wave is sliced off for one-third of the period. Maximum harmonic output occurs at harmonics which are third multiples (3, 6, 9, etc.) of the fundamental frequency.

As mentioned before, the general trend is for the amplitude to decrease with increasing harmonic number, but this tendency is not uniform. A curve joining the tops of the relative amplitude columns makes a kind of wave pattern of its own. with minimum values at the 9th, 15th, 21st, and 27th harmonics.

Fig. 2 shows the situation when a still larger portion of the lower half of a sine wave is sliced off. In this case it would seem that an amplifier is driven into cutoff for a third of the duration of a period. In this case harmonic content is maximum for every harmonic divisible by three. Peculiarly, the two intervening harmonics have amplitudes of equal magnitude.

Another point worth noting is that the fundamental frequency (or the first harmonic) has an amplitude about 7% larger

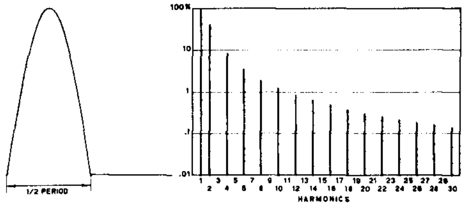


fig. 3. Sine wave with entire negative portion clipped off (rectified half-wave). Note that odd-order harmonics are eliminated entirely.

than the original deformed sine curve. What a temptation to drive an amplifier just a trifle over saturation, to get that little extra gain at the fundamental frequency! Actually, for this series of idealized waveforms the fundamental frequency increases to a maximum of 107.305% when the portion of the sliced-off sine wave represents 31.9% of the period.

half sine waves

Fig. 3 shows the case where the lower excursion of the sine wave has been clipped off entirely—a rectified half wave. Note that every harmonic with an uneven number has disappeared. The fundamental frequency has an amplitude equal to the actual amplitude of the half wave.

Fig. 4 shows a sine wave sliced off even more, so that only a third of the period is left. The envelope over the harmonic columns shows a pattern much like **fig. 2**, but the fundamental frequency has an amplitude of only 78% of the curve's amplitude. The higher harmonics, though, have larger amplitudes than those of **fig. 2**.

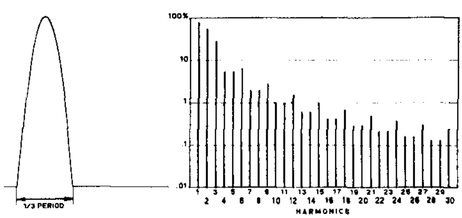


fig. 4. Raising the clipping point above the zero-crossing point sharply increases harmonic output. Note resemblence of this spectrum to that of fig. 2.

In **fig. 5** the sine wave is sliced off so much that only one-sixth of the period remains. The fundamental frequency has an amplitude of only 43% of the curve but the harmonics are quite prominent. As in **fig. 1**, the envelope over the harmonic amplitude columns shows a wave pattern with minima at the 9th, 15th, 21st and 27th harmonic. For more precise harmonic amplitude values for **fig. 1** through **fig. 5**, see **table 1**.

what this means

Now we can return to the questions asked at the beginning of this article. Why are certain circuits better suited for frequency doubling than others? Looking at **table 1**, you can see that the second *harmonic in the column under 2/3* is larger than all the others. This means that slicing off two-thirds of the sine wave would give better second harmonic performance than any of the other clipped waveforms pictured. Actually, for the highest second harmonic content, 66.68% of the sine wave must be clipped off. This represents a conduction angle of 119.9° and provides a second-harmonic amplitude of 55.133% of the basic curve's amplitude.

table 2. Sine waveform clipping required to optimize different harmonics.

harmonic number	clipping required	conduction angle	amplitude (maximum)
1	31.9%	245.1°	107.305%
2	66.7%	119.9°	55.133%
3	77.9%	79.7°	36.908%
4	83.4%	59.8°	27.719%
5	86.7%	47.7°	22.190%
6	89.0%	39.8°	18.498%
7	90.5%	34.1°	15.859%
8	91.7%	29.9°	13.878%
9	92.6%	26.5°	12.337%
10	93.4%	23.9°	11.104%
11	94.0%	21.7°	10.095%
12	94.5%	19.9°	9.254%

table 1. Harmonic amplitudes as a percentage of the basic sinusoidal waveform for various amounts of clipping.

	portion clipped off bottom of sine waveform				
harmonic number	1/6 (fig. 1)	1/3 (fig. 2)	1/2 (fig. 3)	2/3 (fig. 4)	5/6 (fig. 5)
1	104.089	107.267	100.000	78.200	43.045
2	2.843	18.378	42.441	55.133	39.598
3	2.462	9.189	0.000	27.566	34.293
4	1.990	1.838	8.488	5.513	34.293
5	1.477	1.838	0.000	5.513	27.719
6	0.975	2.100	3.638	6.301	20.576
7	0.528	0.656	0.000	1.969	13.577
8	0.169	0.656	2.021	1.969	7.349
9	0.082	0.919	0.000	2.757	2.357
10	0.224	0.334	1.286	1.002	1.143
11	0.269	0.334	0.000	1.002	3.120
12	0.239	0.514	0.890	1.542	3.741
13	0.162	0.202	0.000	0.606	3.323
14	0.069	0.202	0.653	0.606	2.261
15	0.018	0.328	0.000	0.985	0.957
16	0.079	0.135	0.499	0.405	0.245
17	0.109	0.135	0.000	0.405	1.106
18	0.106	0.228	0.394	0.683	1.513
19	0.078	0.097	0.000	0.290	1.471
20	0.036	0.097	0.319	0.290	0.506
21	0.006	0.167	0.000	0.501	0.089
22	0.040	0.073	0.264	0.218	0.559
23	0.058	0.073	0.000	0.218	0.813
24	0.059	0.128	0.221	0.384	0.826
25	0.045	0.057	0.000	0.170	0.633
26	0.022	0.057	0.189	0.170	0.311
27	0.003	0.101	0.000	0.303	0.042
28	0.024	0.045	0.163	0.136	0.336
29	0.036	0.045	0.000	0.136	0.507
30	0.038	0.082	0.142	0.245	0.529

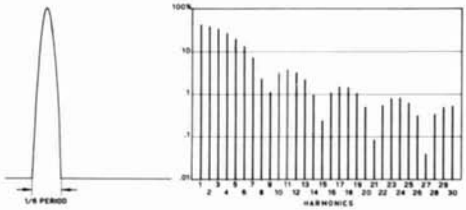


fig. 5. Severely clipping a sine wave so that only the positive peak remains results in lower-order harmonic output very nearly as strong as the fundamental.

In **table 2**, the optimum slice-fractions (conduction angles) are given for harmonics up to the 12th, with their corresponding highest possible amplitude for this set of non-sinusoidal curves.

summary

An oscillator producing an output curve in the form of a horizontally amputated sine wave will provide good second harmonic content if the horizontal part of the curve is approximately two-thirds of the total period. If frequency tripling is wished, the horizontal part of the curve (that is, the duration of the cutoff) should be in the neighborhood of three-quarters of the total period. Conversely, if a specific harmonic must be minimized the data of **table 1** can be used to determine the best operating point.

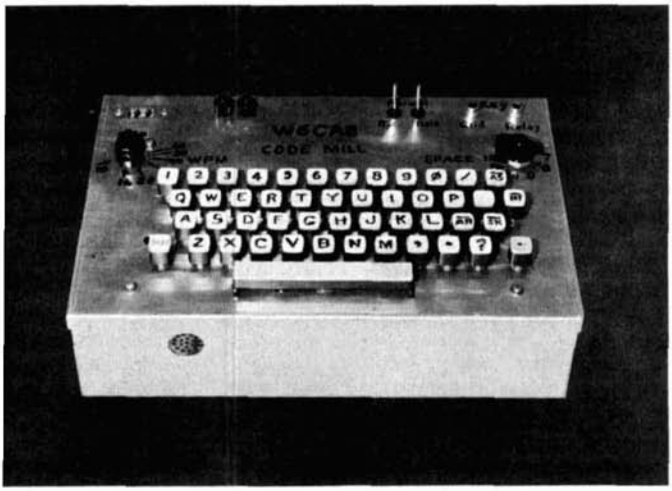
APPENDIX

A Fourier-series of the form $f(X) = \frac{1}{2}a_o + a_1 \cos X + a_2 \cos(2X) + \ldots$ for a sine wave clipped off by a fraction 2V of the total period 2π will have the following coefficients:

$$a_o = [2/(1 + \cos V)]\ [2/\pi]\ {}_V\int^{\pi} (\cos V - \cos X)\, dX$$
$$= [4/\pi(1 + \cos V)]\ [(\pi - V) \cos V + \sin V]$$
$$a_1 = [2/(1 + \cos V)]\ [2/\pi]\ {}_V\int^{\pi} (\cos V - \cos X) \cos X \cdot dX$$
$$= [2/\pi(1 + \cos V)]\ [V - \pi - \sin V \cdot \cos V]$$
$$a_n = [2/(1 + \cos V)]\ [2/\pi]\ {}_V\int^{\pi} (\cos V - \cos X) \cos (nX) \cdot dX$$
$$= [2/n(1 + \cos V)] \left[\frac{\sin (n + 1)V}{n + 1} + \frac{\sin (n - 1)V}{n - 1} - \frac{2 \sin(nV) \cos V}{n}\right]$$

a_o represents the value of the direct current present, and has no significance in this discussion. The absolute values of a_1, a_2, and so on, give the relative amplitudes of the corresponding harmonics.

ham radio



the code mill

A Morse keyboard that provides perfect character generation as well as perfect letter and word spacing, regardless of the operator's keyboard skill

H.E. Thomas, W6CAB, 33020 Almond Street, Lake Elsinore, California 92330

Almost without exception, any person who uses Morse code can attain highest copying speed only when the code is sent perfectly. Because of this, and also in many cases as a matter of pride, most amateurs have tried to improve their sending technique. Many different kinds of equipment have been devised which can assist in this, but none of these, reasonably available to the average amateur, can produce perfect code, despite claims to the contrary.

Perfect International Morse code consists of five parts: the dot, which is one time unit or baud; the dash, which is three bauds; the space between each element of a Morse character, which is one baud; the space between letters, which is three bauds; and the space between words, which is seven bauds. Each and every one of these five parts of International Morse must be produced with machine-like precision – not a single one can be produced by human estimation if *perfect* code is to be the result.

The typical bug can make dots quite well but the other parts of the code are up to the operator. Some types of electronic keyers can produce dots, dashes and the spaces between them in proper form, but the operator is responsible for the letter and the word spaces. The latest device, which goes a step beyond the electronic keyer, is what I call a *code mill*. In radio parlance a typewriter is a *mill*, so what better name for keyboard keyers?

These code mills create perfect dots, perfect dashes, perfect spacing between elements of the characters and terminate each character with a perfect letter space. Whether this letter space remains perfect depends on the availability of the succeeding letter at precisely the right time. This in turn depends on whether the operator is successful in punching the key of the succeeding letter during what might be called its *launch window*. This launch window becomes quite narrow as the code speed increases and is almost intolerably narrow for any character which may follow the letter E (the launch window in this case is less than a tenth of a second at 30 wpm).

Some keyboard keyers provide a one-character storage latch at the input of the circuitry but even this partial solution requires the operator to anticipate various letter combinations. The varying length of the different Morse characters is not compatible with the smooth rhythm of good typing, particularly at the higher speeds. And, of course, the usual code mill does nothing whatsoever about word spacing. Listening on the air will give ample evidence of the inability of such devices to produce perfect code.

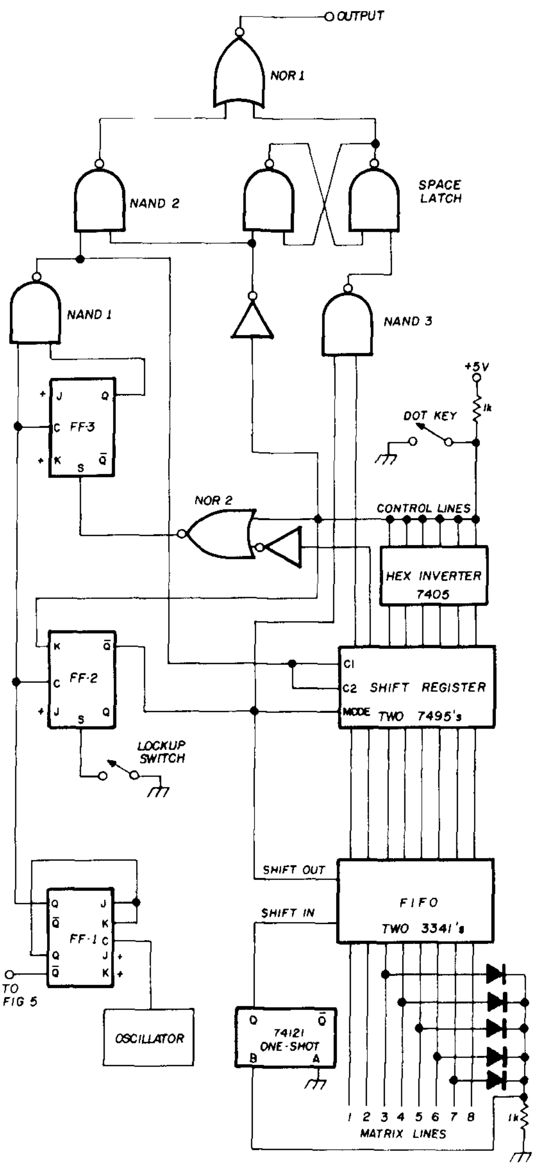


fig. 1 Block diagram of the code mill which uses first-in first-out (FIFO) serial memory to generate perfect Morse code characters and spacing, regardless of the operator's keyboard skill.

the solution

The obvious answer to the problem is to design a code mill which will automatically produce all five parts of the Morse code. Since this includes word spaces, a space bar for the keyboard is required, as in teletype equipment. Furthermore, to maintain exact letter spacing, the availability of successive characters must also be automatic. This means that the operator and his sometimes erratic operation of the keyboard must be isolated from the code transmitting circuits.

The electronic equivalent of a vending machine can best meet this requirement. The keyboard operator dumps the merchandise in at the top in the proper sequence and at whatever rhythm suits, including the "hunt and peck" system. The code transmitting circuit puts the "coin in the slot" and extracts the merchandise in the same sequence but at precisely the correct instant.

the circuit

This and the following paragraphs describe a code mill with an electronic vending machine having a storage capacity of 64 Morse characters. This code mill produces perfect code at any speed between 7 and 70 wpm as long as the operator keeps the storage partially filled and does not exceed its storage capacity at any time.

The block diagram of **fig. 1** shows the basic arrangement which uses two Fairchild 3341 mos ICs for storage. Each 3341 is a 64 by 4-bit FIFO (first-in first-out) serial memory, and is available in a standard ceramic 16-pin dual-inline package. A detailed description of circuit operation will not be given here since the circuit, excluding the FIFOs, is a modernized version of a circuit which was described in fine detail in a previous article.[1] However, a brief description follows.

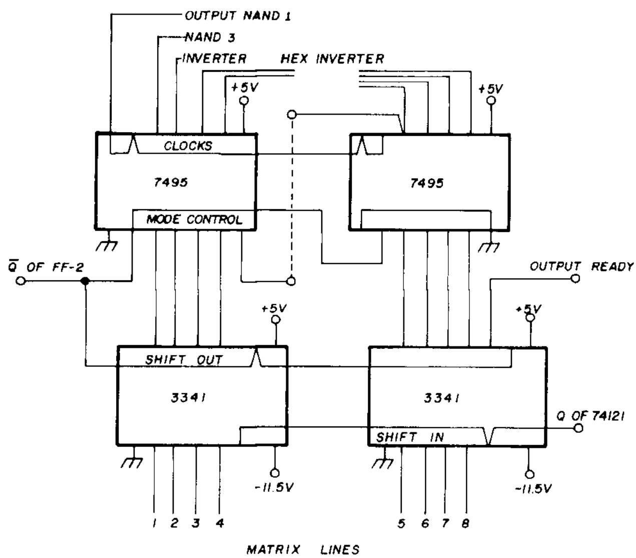


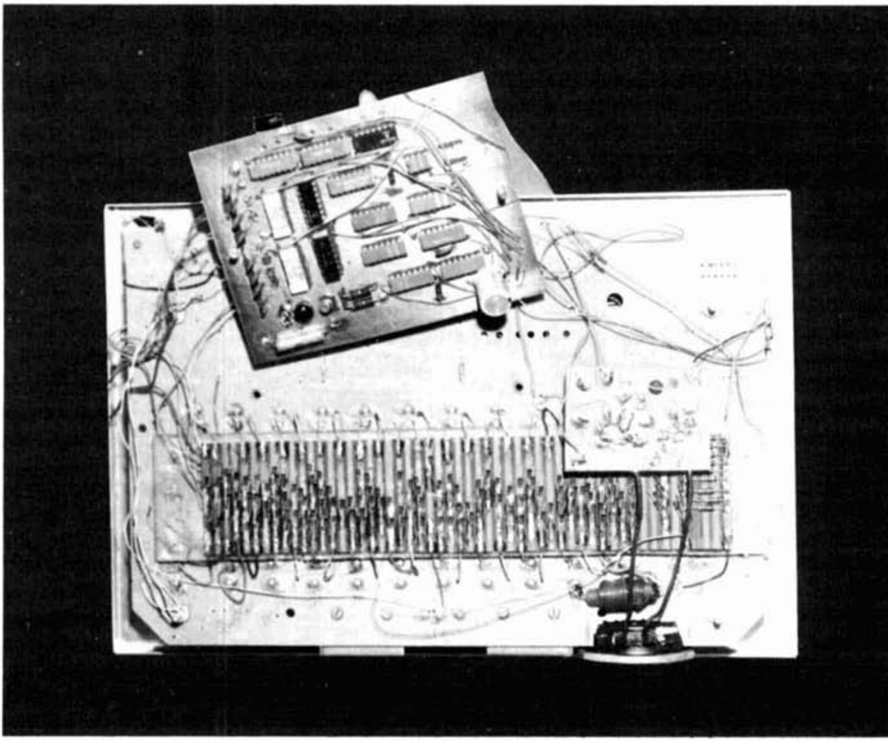
fig. 2. Bottom view, or foil side, of the ICs used in the first-in first-out serial memory and shift register.

The oscillator is a variable-speed pulse generator which runs continuously. It triggers dual flip-flop 1 whose second Q output gives the required square wave. This square wave will produce transmitted dots through gates NAND 1, and NAND 2 and NOR 1 if none of these gates is inhibited. When no character is being processed through the shift register, the control line will be *high* and this inhibits NAND 2 via an inverter. Also, with the control line *high*, the K input of flip-flop 2 is *high* and this flip-flop is then triggered back and forth by the output of flip-flop 1. Whenever the $\bar{Q}$ output of flip-flop 2 is *high*, the FIFOs and the shift registers are ready to process Morse characters for transmission.

At this point it should be mentioned that the mos chips used in this circuit require a bit higher than normal *high* level. For this reason flip-flop 2 is one-half of a Signetics SP321A dual J-K flip-flop. Flip-flop 3 is the other half. Despite the fact that the Fairchild 3341 has internal pull-up circuits to make them compatible with TTL logic, the 3.8 volts minimum *high* of the SP321A looks a lot better than the 2.4 volts minimum *high* for the TTL flip-flops. A 7473 TTL could probably be used to replace the SP321A, but it was not tried.

The *set* input of flip-flop 2 may be grounded with a switch near the keyboard. This blocks the output of the mill while the FIFOs are being filled. With the lockup switch blocking the mill, calls may be typed up while still listening to the transmission from the other station or stations in a two-way or round-table QSO. It is also nice to use when setting up for a repetitive transmission using the recirculating shift registers to be described later.

When any key on the keyboard other than the space bar is depressed, the inputs to the FIFOs are actuated by the 74121 one-shot multivibrator, the character enters the FIFOs and almost instantly appears at the outputs. By the action of flip-flop 2 and the output of NAND 1 on the FIFOs' *shift out* controls and the shift registers' mode controls and clocks, the character enters the shift registers and eventually appears at the inputs of the hex inverter. This drops the control line output of the shift registers, the control line goes *low* because of the E, enabling NAND 2 and the left-hand gate of the space latch. The *high* on line 1, via NAND gate 3, sets the latch, which in turn inhibits NOR gate 1 and prevents transmission of the E. Note that NAND 3 is inhibited by the output of flip-flop 2 except during the loading of the shift



Construction of the code mill showing the matrix board mounted on the back of the keyboard switches. Oscillator is on small board to the right. Main circuit board, including memory, is at top.

to a *low* which enables NAND 2 (via the inverter) and the character is transmitted. NOR gate 2 and flip-flop 3 serve with NAND 1 to bridge the gap between two successive dots to produce a dash when required.

word space

When the space bar is depressed the letter E is formed and matrix line 1 is also activated (raised to a *high*). When the E and the *high* on line 1 appear at the registers. This prevents line 1 from setting the space latch when it goes *high* due to the shifting of *highs* along the shift registers. At the completion of every character the control line goes *high*, resetting the space latch through an inverter.

The letter E is used for the space character because it's one-baud length plus the three-baud automatic letter space adds to the previous three-baud letter space to become a seven-baud word

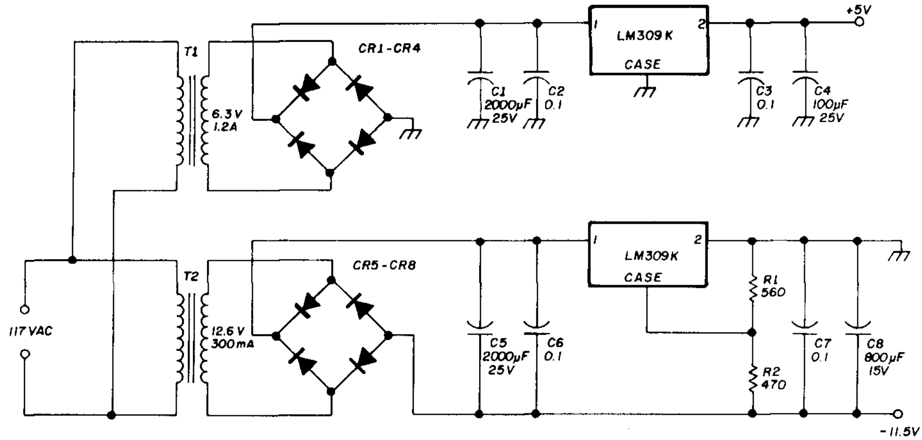


fig. 3. Regulated power supply for the code mill uses LM309K voltage-regulator ICs.

space. The letters I, S, H or the numeral 5 may be used if wider word spaces are desired. A switch is provided for this purpose, although it has seldom been used except while teaching code to beginners.

Fig. 2 shows the bottom view, or foil side, of the ICs comprising the FIFO and the register. **Fig. 3** shows the circuit for the power supply. Both the +5 volt and the -12 volt supplies are regulated by LM309K voltage regulators. The values of the resistors in the -12 volt supply were chosen to give an output of -11.5 volts.

last-character elimination

Fig. 4 shows a circuit which was added some time after the code mill was built. Without touching the keyboard, this circuit will trigger the FIFO inputs at the conclusion of a character (control line goes *high*) and when the *output ready* terminal of one of the 3341s indicates that the FIFO is empty. Unfortunately, the 3341 will never completely empty itself. The last character will remain at the output of the FIFO until another character enters and "falls through" to the output. This causes continuous repetition of the last character. The operator must either learn to punch the space bar as a last character or a circuit must be designed to prevent this repetition. By triggering the input without punching a key, a complete blank is entered and passes through, and in turn takes the place of the offending character.

Since the Fairchild 3341 memories require +5 volts and a -12 volts in addition to the usual ground connection, it was a simple matter to add the recirculating shift registers shown in **fig. 5**. The reason flip-flop 1 is a dual unit is now apparent: it permits triggering the recircu-

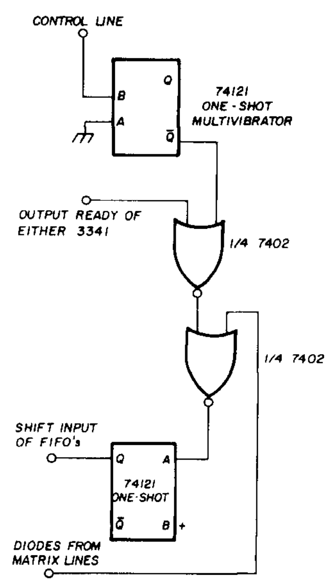


fig. 4. Circuit for last-character elimination.

lating shift registers with the output from the first $\overline{Q}$ which is equivalent to a trigger for every baud in the mill output.

The inverters at the input and output of the 2522 chips are used as buffers. The two 2522 chips will store every transmission from the code mill up to a limit of 528 bauds. When the spdt switch is thrown, the information stored in the 2522s will repeat over and over, and at any speed. Although not a necessary part of the code mill, this circuit is handy for preliminary callups prior to broadcasts, directional CQs or whatever, while you sip your coffee.

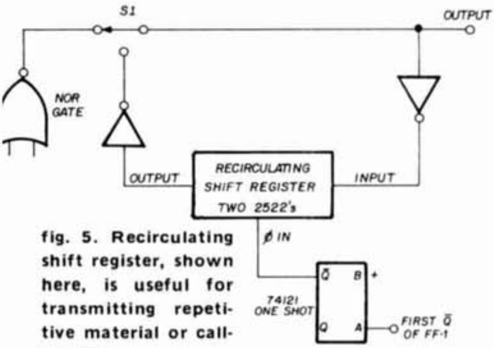


fig. 5. Recirculating shift register, shown here, is useful for transmitting repetitive material or calling CQ.

The code mill described in this article was built for just over one hundred dollars, and all chips were purchased at full list price. My keyboard itself is a surplus unit and similar types are available at several outlets. The price was not too high considering the very great pleasure and satisfaction derived from building the unit and operating it on the air. Without mentioning the code mill, almost every QSO elicits favorable comments on the keying. The code mill is just great for code practice, of course, but probably the best feature of all is the ability to send absolutely perfect code at fairly high speeds, hour after hour if need be, and with no great effort.

reference

1. J.A. Bryant, W4UX, "Touchcoder II," *QST*, July, 1969, page 12.

ham radio

automatic telephone controller for your repeater

A simple, reliable and foolproof control system for remote control of your repeater

Robert C. Heptig, KØPHF, Robert D. Shriner, WAØUZO

The problem of building a practical, positive remote-control system has plagued amateurs for years. Some of the suggested schemes used have been very complex while others have been simple, but not foolproof. The technique applied in the design described here uses the telephone line. Positive and automatic control in the event of line failure are the basic assets of this unique system. Don't be alarmed at the apparent complexity of the schematic, **fig. 1**, as all the IC gates are contained in five, inexpensive SN7400 IC packages. Total cost of the circuit for ICs is less than $5.00 and a well stocked junk box will supply the balance. Although the circuit can be built on peg board, a custom made printed-circuit board is available and is recommended for simplicity.* See **fig. 3** for a suggested parts layout.

construction

A small chassis was made from sheet aluminum and bent in a vise to form the chassis shown in **fig. 2**. The power supply and relays are mounted on the chassis for convenience. A small equipment box could have been used but one was not available at the time I built the system. It is suggested when you choose your chassis that you keep the top and bottom open to aid in troubleshooting. A coat of white paint on the front panel and instant press-on lettering may be used to dress it up.

The model illustrated is set up for a

*Printed-circuit boards and nonlatching relays are available from Circuit Board Specialists, 3011 Norwich Avenue, Pueblo, Colorado 81008. Circuit boards are $5.50 each; nonlatching relays, $2.00 each.

latching function—it turns our repeater off. If you happen to have a stepping relay, this is fine, but I didn't, so another SN7473 was mounted in a socket off to the side and wired in as shown in the alternate circuit, **fig. 4**.

operation

An operating function, such as shutdown or turn-on of a remote station, requires a specific number of rings on the telephone, hang up, wait a specific length of time, then another specific number of rings and the function will be carried out.

In this discussion I will use 3 rings, hang up, wait 20 seconds, another 3 rings and hang up. Any combination of number of rings may be used so long as the total is less than nine. By examining the decoder (U2) you can see that it can be very easily programmed by the moving of only two jumper wires to the various outputs of U2.

Depending upon the type of relay used at relay K2, a momentary, latching or stepping function can be obtained. Relay K1 is used for validating the phone line. The remote station keying voltage can be taken through the contacts of this relay. If the phone line is interrupted, the transmitter cannot be activated.

circuit description

The phone line must be connected with the polarity as shown in the schematic. To assure the phone line being operational, the negative voltage is passed by CR1 and R1, charging C1, which will store the voltage during brief interruptions such as when the phone is ringing. The negative voltage applied to the gate of Q1 causes it to cease conducting. A high plus voltage will appear on the drain of Q1. This voltage is passed to the Q3 base by R4, causing Q3 to conduct and close relay K1. For repeater control K1 is connected in the push-to-talk or transmitter keying line so that it must be closed in order for the repeater to operate.

Although an explanation of the ring-counting circuit is a bit complicated, operation is actually very simple. In the ring circuit you are only interested in the ac signal. Capacitor C2 is used to block the dc and pass the ac signal to the rectifiers. The negative dc voltage is smoothed by C3 and C4 and applied to the gate of Q1, causing a rise in voltage on the drain of Q1. As soon as the ring signal disappears, R7 pulls the gate of Q1 to ground, resulting in full conduction of Q1, causing pin 14 of U1 low. Each shift from high to low on pin 14 will cause U1 to toggle once.

Therefore, the first ring will step the decade counter (U1) one count. The decoder-driver (U2) has now moved off zero position to its first count. When U2 leaves zero, pin 16 goes to +5 volts. This high is fed to pins 1 and 2 of gate 17 (U8A), and pin 3 goes low, causing the ready light to go out. Also, pins 4 and 5 of gate 18 are low, causing pin 6 to go high, turning on the start light indicating that a function has started. The high from gate 18 pin 6, is also fed to U7, pin 14, providing a set signal for flip-flop 1 (U7A), as well as providing the clear voltage to both flip-flops at pins 2 and 6.

To back up a little, to U2 pin 16, +5 volts is also fed to the emitter of Q4. This is the timing circuit of the telephone controller which provides the 20-second timing necessary for all phases of the function. The time is controlled by the RC network R11 and C5.

This all happens on the first ring of the telephone. The second ring simply advances the counter to count two. The third ring, the one that does the business, advances the decoder driver, U2, to count 3 position. The resultant low on pin 9 (count 3) of U2 is seen at pins 1 and 2 of gate 3. Gate 3, pin 3, will provide a high enabling voltage to pin 12, gate 1. Now, assuming that the telephone quit ringing on the third time, pin 12, gate 1 will remain high. After 20 seconds the unijunction transistor, Q4, will have charged up and fired a voltage spike out B1. This spike is fed to pin 13, gate 1. The output of this gate goes low, clocking FF1 (U7A) at pin 1. The Q output of FF1 goes high, applying a set voltage to FF2. At the same time the phase 1 lamp comes on indicating that the first requirement of a function has been accomplished. Pin 11, gate 1, is also fed to gate 16, pin 2, which is used to reset the whole system if the telephone fails to ring again within the next twenty seconds.

When phase 1 has been successfully completed, the telephone must ring three more times to complete the function. The fourth and fifth rings will simply step the ring counter two more counts. On the sixth ring the phone must be hung up, leaving U2, pin 11, (count 6) low. This low is seen on pins 4 and 5 of gate 4, providing a high to pin 9 of gate 2. Twenty seconds after the last clock pulse from Q4 another clock pulse is sent to pin 10, gate 2. Gate 2, pin 8, goes low, causing FF2 to toggle. Pin 9 of FF2 goes high, causing the phase 2 lamp to turn on. This action also turns on the relay driver, Q5, closing the function relay, K2. Transistor Q5 will remain turned on until the third pulse is received from Q4. This time the clock pulse is fed to pin 2, gate 14. Pin 1 is now high since it is controlled by the $\overline{Q}$ output of FF2. The pulse is fed through gate 14 and 15 to reset U1. This restores U2 to zero, clearing FF1, FF2 and turning on the ready light. Thus, one complete function has been performed.

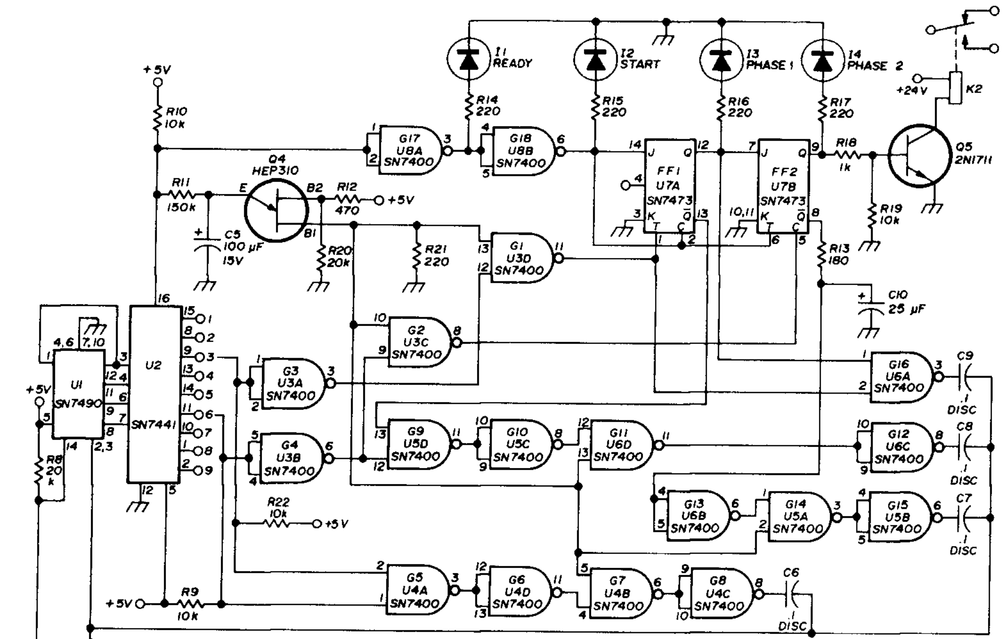
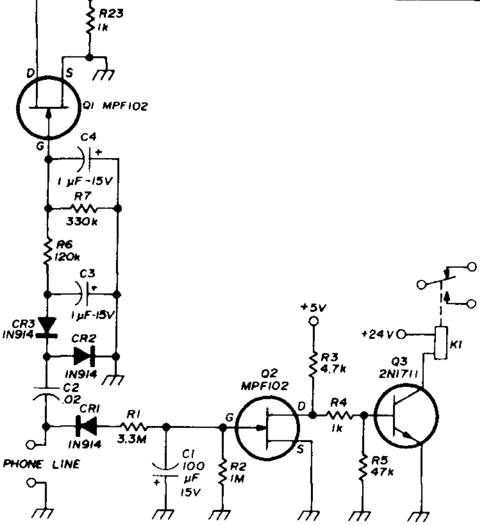



fig. 1. Schematic diagram of the automatic phone controller. All gates are SN7400 or equivalent. Relays K1 and K2 are sensitive dpdt relays with 8000-ohm coils. Resistor R11 is selected for the desired time setting. Resistors R14, R15, R16 and R17, in series with the LEDs, are adjusted for proper light output.

At this point you may be asking what the rest of the gates are for? Well, there must be a means of discriminating between a valid function and an incoming phone call to your wife or teen-age daughter. A few instances of an invalid function would be:

1. Ring three times and hang up. The phone, not ringing again for the next 20 seconds, would allow the system to reset through gates 1 and 16.

2. The phone ringing any number of times except 3 or 6 then hung up, the system will be reset through gates 5, 6, 7 and 8.

3. The last possibility is a remote one but nevertheless it must be considered. In this case someone lets the phone ring three times and hangs up. By sheer stroke of luck or accident they call back twenty seconds later and let it ring any number of times other than three or six. Then the entire system will be reset through gates 13, 14 and 15.

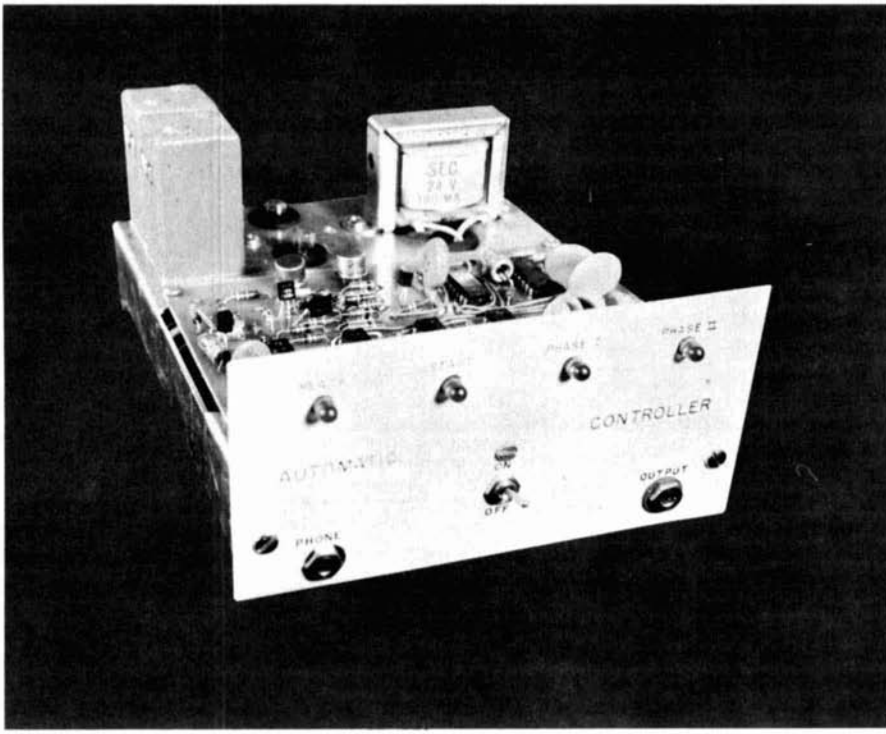


fig. 2. Overall view of the completed controller. Small jacks on the front panel are used for connections to phone and output. (Photo by Russell McGee, WBØGSU.)

This concludes the operation of the controller. As far as I know I have thought of all possibilities of mis-control and have provided protection against it. If, by accident, someone does get your function-control code it is a very simple matter to change the sequence to another code, such as ring 7, hang up and ring 2 more times. As you can see, there are 36 different combinations of sequences. If that isn't enough you can change the value of R11 and C5 for a different time duration. Needless to say, it isn't necessary to describe the multitude of combinations you can come up with using this arrangement.

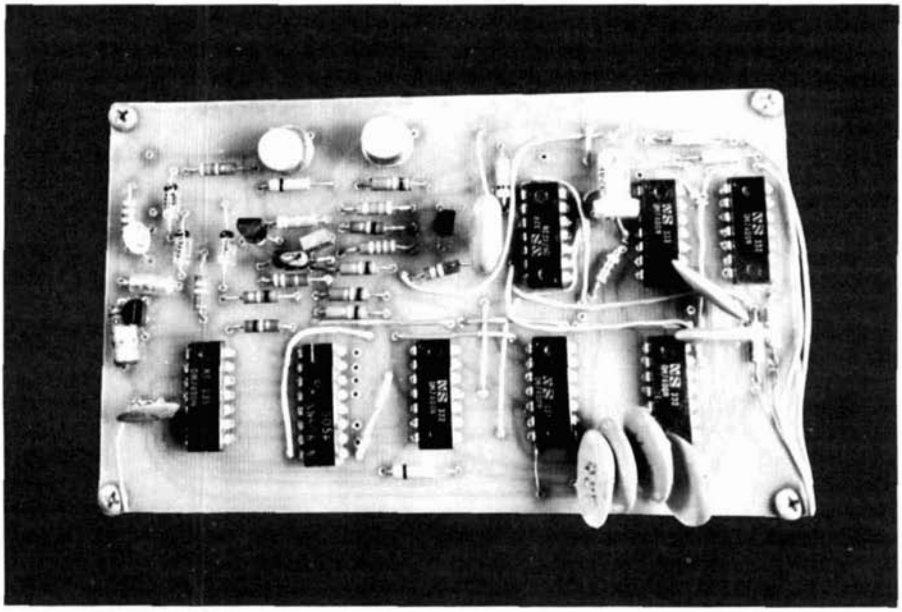


fig. 3. Component layout. The two transistors at the top are Q3 and Q4. The input and output connections are on the bottom; wires run to the light-emitting diodes located on the front panel. (Photo by Russell McGee, WBØGSU.)

troubleshooting

A high-impedance scope or a vacuum-tube voltmeter will be required in the event of trouble. However, the light-emitting diodes connected to each stage will generally suffice for troubleshooting.

After construction is completed and voltage is applied to the circuit check for the following: the ready light, I1, should be on and all others off. Pin 16, U2, should be low, pins 9 and 12, U7, should be low and pins 8 and 13, U7, should be high. Connect the positive terminal of a small 9-volt battery to ground and the negative side to the input. Relay K1 should close and remain closed for about 30 seconds after removing the 9-volt battery.

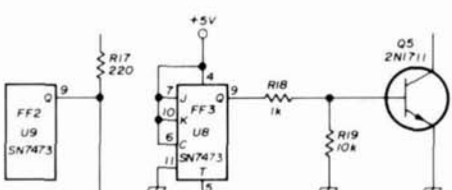


fig. 4. Alternate circuit for use when a latching function is required.

Next, apply the 9-volt battery lead momentarily to the junction of CR3 and R6. Approximately 2 seconds after removing the battery lead I2 should light, indicating a function has started. Make and break this connection, allowing about 2 seconds between pulses for the correct number that corresponds to the first number of rings that you have programmed for. About 15 seconds later I3 should light, indicating that the first phase of the system has been satisfied. Now make and break the battery connection in the same manner for the second combination of rings; in approximately 5 seconds I4 will light, indicating that the sequence has been completed and the function has been carried out.

In the event of difficulty check base 1 of the unijunction transistor with a scope for the pulse that controls the system. If the pulse is being generated, then trace the pulse through the system to determine which gate or flip-flop is not responding.

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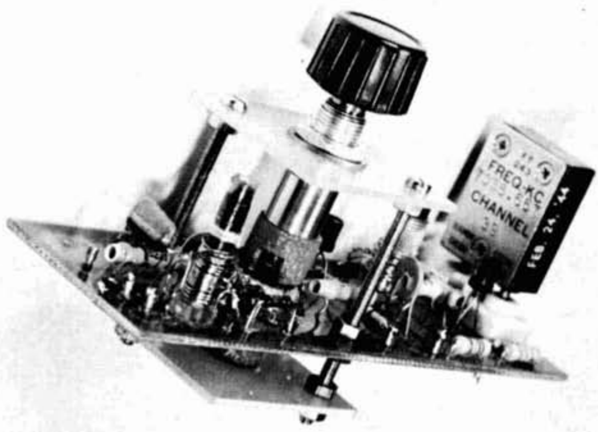


tuned very low-frequency converter

Novel tunable inductance provides basis for tuned VLF converter that covers wide frequency range without bandswitching

Guenter Ruehr, OH2KT, Porintie 1 C 35, SF - 00350 Helsinki 35, Finland

There is considerable interest in the very-low frequencies, particularly for the WWV transmissions on 20 and 60 kHz, and for operation on the no-license band around 1750 meters.[1] One of the big problems in building receivers or converters for this part of the spectrum is in constructing a variable tuned circuit which will cover a substantial portion of the desired frequency range. Assuming that the desired band extends from 10 to 150 kHz, with a ratio of the corner frequencies of 150:10 = 15, the tunable component must have a variation of $15^2 = 225$. Since this cannot be accomplished with conventional variable capacitors or inductors, the frequency range has to be divided into a number of sub-bands or the tuned circuit is eliminated altogether. The latter is done in most VLF converters – they are untuned.

tuned circuit

There is, however, a novel method of inductive tuning which will cover the required range.[2] This method makes use of a toroidal ferrite core which is magnetically biased by a pair of small permanent magnets as shown in **fig. 2**. By rotating one of the magnets with respect to the other, the amount of flux penetrating the toroid is varied, changing the ferrite's permeability and thus, the inductance. It is interesting to note that maximum flux penetration and minimum inductance occur when like poles are opposite one another.

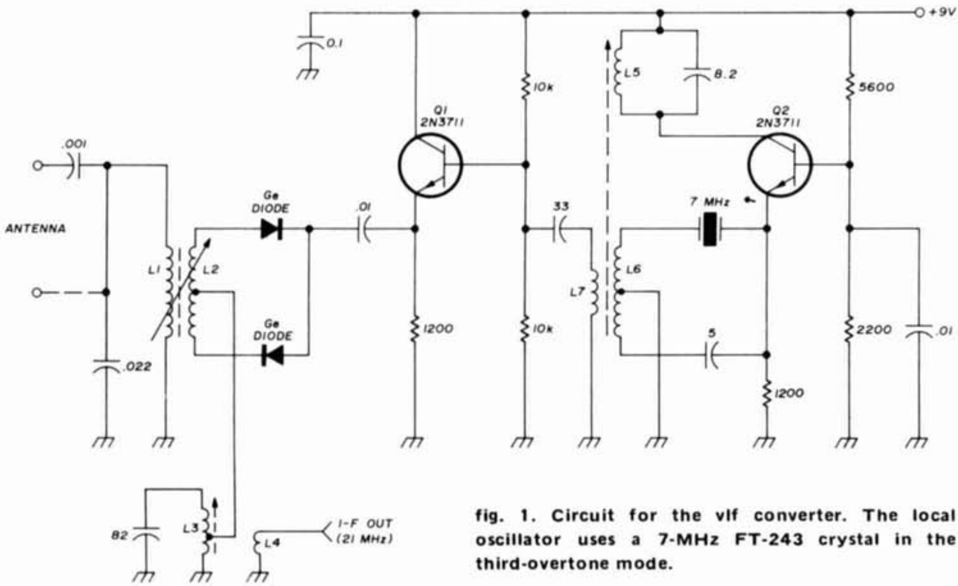
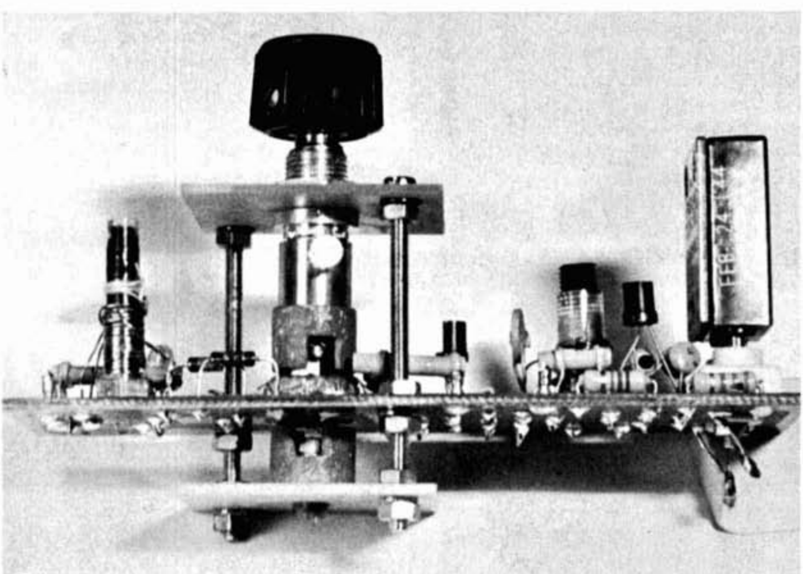


fig. 1. Circuit for the vlf converter. The local oscillator uses a 7-MHz FT-243 crystal in the third-overtone mode.

L1,L2	magnetically tuned inductor (see text)
L3	10 turns no. 20 on ¼" (6-mm) slug-tuned form, tapped 5 turns from cold end
L4	2 turns no. 20 around cold end of L3
L5	15 turns no. 20 on ¼" (6-mm) slug-tuned form
L6	4 turns no. 20, center tapped, around cold end of L5
L7	2 turns no. 20 around cold end of L5

The two magnets used to bias the toroid inductor are of the *button* type with a half-inch (13-mm) outside diameter. The outside diameter of the toroid is also ½-inch. The whole tuning assembly is built around the bushing and shaft of a discarded potentiometer. The particular toroid core I used required 100 turns of



Closeup of the tunable vlf converter showing magnetically-biased tuning inductor.

stranded wire for an inductance variation of 100 μH to 12 mH (a 120:1 range). However, ferrite cores with higher permeability would require fewer turns. Measured Q values for my inductor were around 50 for frequencies between 10 and 150 kHz.

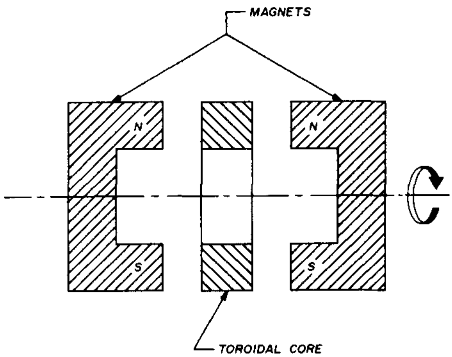


fig. 2. Method for magnetically biasing a toroidal ferrite core with two small button magnets. This technique provides an extremely wide inductance tuning range.

converter circuit

The circuit for my VLF converter, which has an output on 15 meters for use with a communications receiver, is fairly conventional as shown in **fig. 1**. The antenna is coupled directly to the hot end of the tuned circuit (or through a capacitor to provide a degree of matching to long antennas). The mixer uses a matched pair of germanium diodes and the local oscillator uses a FT-243-variety crystal in the third-overtone mode. To obtain third-overtone oscillation at 21.000 MHz, choose a crystal with a fundamental frequency a few kHz above 7 MHz. By simply changing the crystal frequency and the oscillator and i-f output coils, output can be changed to any desired frequency band.

references

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2. N.H. Brown, "Miniature Wide-Range VLF Tuner," *Electronics World*, July, 1971.

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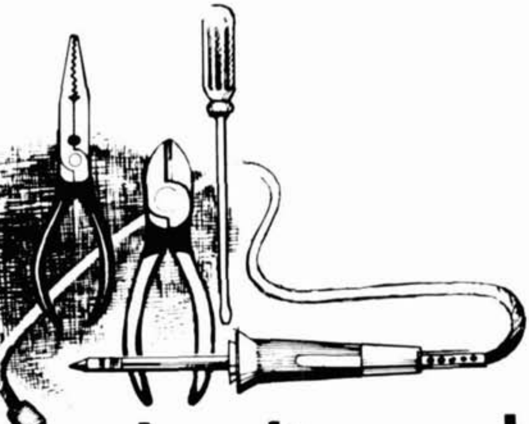


circuits and techniques

ed noll, W3FQJ

W3FQJ goes solar power

On April 15, 1974, the solar-powered QRP station at W3FQJ made an initial radio contact with Hank Brazeal, WB4ZXJ, in Birmingham, Alabama on 15 meters. The transceiver was the five-watt PEP Ten-Tec Argonaut. The same evening W3FQJ checked into the RF Hill Amateur Radio Club's 10-meter net.

The essential units of the solar-powered station are the Argonaut transceiver, charging panel and 5.5-ampere-hour motorcycle battery **(fig. 1)** and a roof-mounted light-energy converter, **(fig. 2)**. The 5x2½x5½-inch (12.7x6.6x14-cm) battery is positioned behind the charging panel for normal operation. Such batteries can be purchased at local motorcycle shops or by mail from one of the auto accessory houses or from Sears. The Sears batteries are shipped with a dry electrolyte which must be added to the battery along with water according to the instructions. In most shops the electrolyte is added when the battery is purchased.

The Spectrolab light-energy converter has a rating of 12-volts at 0.3 amperes. When the solar panel is so operating it is supplying 3.6 watts. This is an average figure. At dawn and dusk and during dark, overcast days the power level is significantly lower. When the rays of the sun are striking the panel directly, the delivered power is slightly more. Nevertheless, it will make available more than enough electrical energy for a very busy five-watt QRP station.

The solar panel can be operated as a continuous float-voltage charger or it can be operated whenever you wish to recharge the battery. As I am writing this column in the early morning hours of a high-overcast day the float-voltage connection is supplying 50 milliamperes.

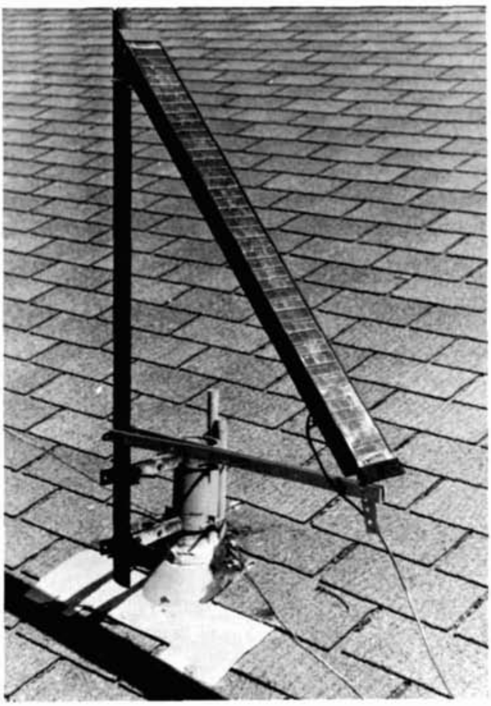


fig. 1. Solar power converter used by W3FQJ is mounted on the roof. Details of the mount are discussed in the text.

Yesterday in bright sunlight a series resistor was inserted to keep the trickle charge current below the 100 mA level. As set up now the station plan is very conservative and would provide continuous operation for normal back-and-forth amateur chatter. Over a period of time I will learn the limits of the system, pressing high-powered transmitters into operation. Such is not conducive to the furthering of corporate empires.

charging panel

Several components are needed for the charging panel. Refer to the schematic diagram and parts list shown in **fig. 3**. About 50-feet (15-meters) of two-conductor cable (number-12 or -14 con-

fig. 2. Basic solar-powered QRP station used by W3FQJ uses Ten-Tec Argonaut transceiver, solar power converter (fig. 1), solar control panel (fig. 4) and 12-volt motorcycle battery.

ductors) connects the solar polar converter to the charging panel. The panel itself supports a dc milliammeter, a 1-ampere diode, CR1, and a number of binding posts that permit ease in monitoring and experimentation.

The two binding posts at the top right of the front panel, **fig. 4**, can be used for monitoring the voltage delivered by the solar panel. The meter to the left reads the actual current being delivered by the solar cells. The two binding posts underneath the meter permit the insertion of a jumper if you want to take the meter out of the circuit for those experiments where you want to draw maximum current under the condition of bright sunlight directly striking the cell surface.

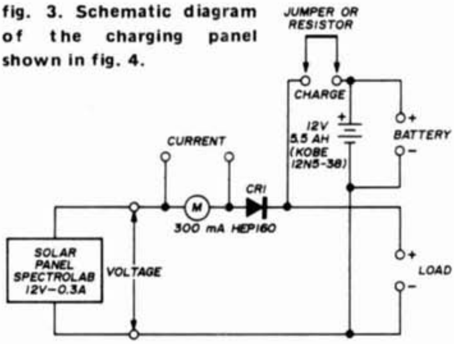
The battery terminals are at the lower right. In a continuous service application

Reports will appear in this column from time to time.

solar power cost

Today solar electric power is expensive. The small 3x39-inch (7.6x99.1-cm) unit I am using lists for more than 150 dollars. Much about it is handcrafted and as yet systems have not been adapted to mass production. Unfortunately, it must be the American citizen that is obliged to set up a widespread clamor for solar power. One cannot anticipate that the oil institute, the electric power industry or the atomic energy bureaucracy will do too much to further the cause of this non-polluting method of power generation. Such power would offer a degree of independence for many homes, small businesses and small industrial plants.

fig. 3. Schematic diagram of the charging panel shown in fig. 4.



the 12-volt device being powered is connected across these two terminals. When the battery is to be charged, a jumper is connected between the two terminals labeled *charge*. If you wish to limit the charge to a specified current level under bright conditions, a resistor can be connected between these two terminals. Often, under very bright conditions, I shunt a 100-ohm, 5-watt resistor between the terminals.

The two load terminals at the bottom center of the panel permit you to supply energy directly from the solar panel to a 12-volt device. In this application the battery is disconnected completely by making certain there is no jumper or resistor connected across the two charge terminals.

Diode CR1 is an important part of the charging system. The actual charging current declines as the battery reaches full charge. This is to be anticipated because the charging battery attains a voltage ever closer to the voltage of the charging source. However, it is possible, especially when the battery climbs to full charge voltage, that the impinging light will not be great enough to maintain the charging source voltage above the level of the battery voltage. Without the diode in the circuit, the battery would then discharge into the solar source. This is avoided because under the condition of high battery voltage and low charging source voltage the diode is reverse biased (cath-

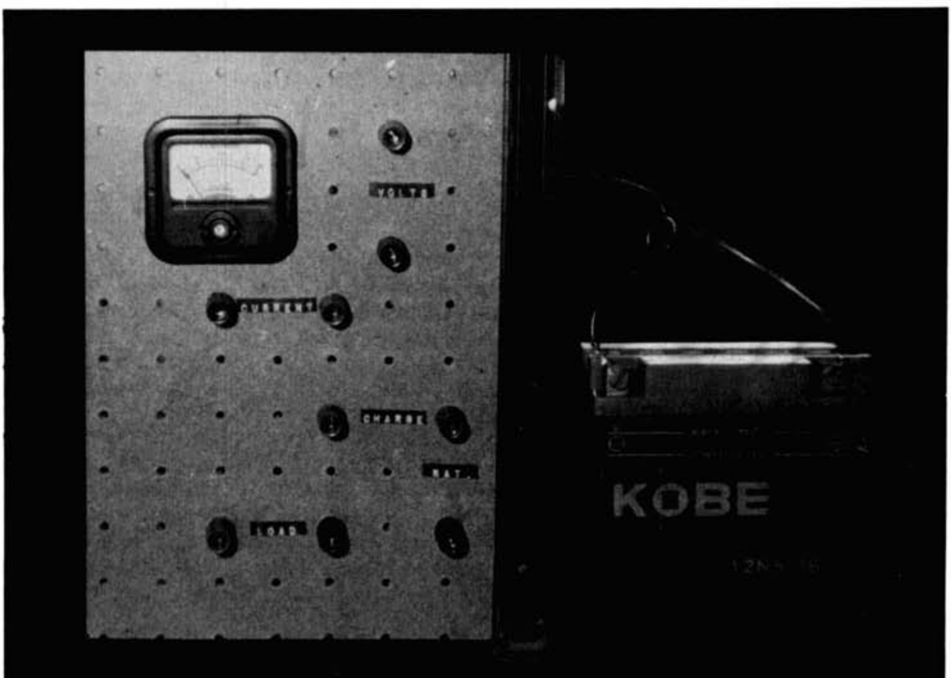


fig. 4. Solar control panel and 12-volt battery used by W3FQJ. Control panel is mounted on 8x11-inch (20.3x27.9-cm) piece of Masonite peg-board. Circuit is shown in fig. 3.

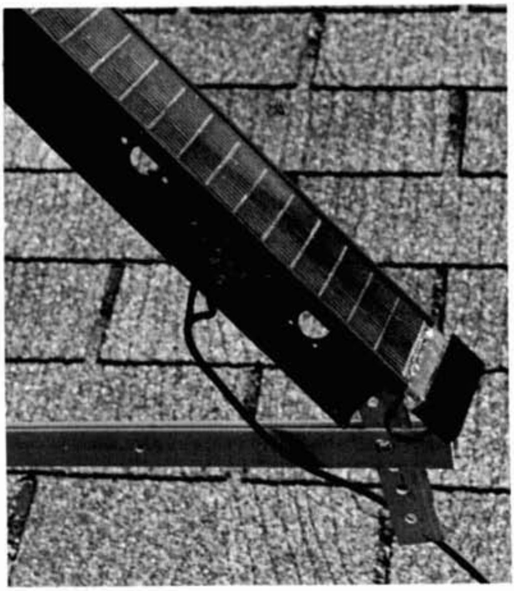


fig. 5. Closeup, showing tilt angle adjusting mechanism for the solar energy panel.

ode connected to positive side of battery and anode connected to the positive side of the charging voltage).

roof mount

The basic physical support for the solar panel is the vent pipe on the roof of my house, **fig. 1**. At one time the two standoff brackets supported an antenna mast. A shortened 5-foot mast section and a homemade bracket arrangement supports the solar panel. It was arranged to permit ease in experimenting with the tilt of the panel. A U-bolt permits the top of the panel to be moved up and down the mast. At the bottom of the panel a flat piece of aluminum with a series of holes permits easy accommodation of various tilt angles.

The recommended tilt for the panel corresponds to the latitude of your station (degrees north [or south] from the equator). My station is reasonably near the 40° north latitude line. This figure refers to the angle of tilt away from the horizontal as shown in **fig. 6**. The nearer you approach the equator, the nearer the optimum mounting angle approaches a horizontal position.

In setting up the mounting arrangement the solar panel becomes the hypotenuse of a right triangle, **fig. 7**. Sine and cosine functions can then be used to determine the vertical and horizontal sides. The overall panel length is 39 inches (99.1 cm); this becomes the length of the hypotenuse. Therefore, the horizontal and vertical side dimensions become:

b = h sin a	a = h cos A
b = 39 inches (99.1 cm) x sin 50°	a = 39 inches (99.1 cm) x cos 50°
b = 30 inches (76.2 cm)	a = 25 inches (63.5 cm)

The vertical side is an appropriate section of the 1¼-inch (32-mm) mast section. The horizontal support is two 1¼-inch (32-mm) aluminum strips bolted together but separated where they wrap around the mast section and where they connect to the short length of aluminum strip that permits the adjustment of tilt angle.

battery data

A 5.5-ampere-hour capacity battery provides a conservative and well-regulated source for the 5-watt Argonaut transceiver. Transceiver specifications suggest a 12-volt, 1-ampere source although peak current demand by the transceiver is less than this value. At any rate, such a battery would supply the unit for continuous overnight operation. In fact, the battery could supply almost a continuous demand of 1 ampere for 4 to 5 hours.

Based on a 20-hour, 5.5-ampere-hour rating, the continuous current demand for this period of time would be:

$$I = \frac{5.5}{20} = 275 \text{ milliamperes}$$

A charging current of the same value would recharge the battery in the same amount of time plus additional time, depending upon the charging efficiency of the battery.

In normal amateur applications deep discharge of the battery and continuous high-current charging are not necessary. Once the battery is fully charged a float charge arrangement is quite adequate. If you assume a charge rate of 1/4 to 1/6 of the rated current value, you are considering a minimum charge current in the 45- to 70-mA range. In the set-up described this level of current and higher is readily available for hazy-bright days. A generous, up to 300-mA, current is available during bright sunny days. Even on a high overcast day the solar panel supplies current at the low end of the range. Charging current is limited for dark days but the battery capacity is adequate, even for the very active QRP operator.

solar power QSO record

Correspondence from Edgar Janes, G2FWA, disclosed two solar-powered CW contacts made on October 27, 1954. Initial contact was made between G3HMO (solar-powered station) Buckingham, England and G5RZ, Leighton Buzzard, England. Here is an account from the December, 1954, issue of *Short Wave Magazine*:

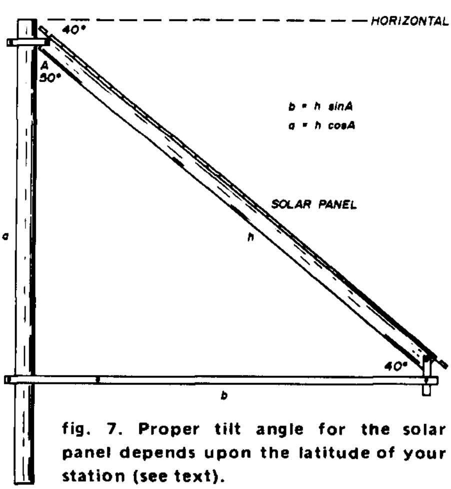


fig. 7. Proper tilt angle for the solar panel depends upon the latitude of your station (see text).

"Fortunately, the sun was shining fairly strongly, and the photo-electric cell battery was giving ample output – about 2 mA at 4 volts – to energize the transmitter . . . Two-way CW contacts (were made) with G5RZ (Leighton Buzzard, 15 miles) at 1505 gmt and with G3IYX (Bradwell, Bucks, 7½ miles) at 1515 gmt. The daylight powered transmitter signals on 1820 kHz were reported as RST 559 in Leighton Buzzard and RST 569 in Bradwell.

These contacts, though pre-arranged, were initiated on the transistor transmitter running on the photo-cell battery alone and were carried through under normal band conditions, with quite troublesome interference on the frequency."

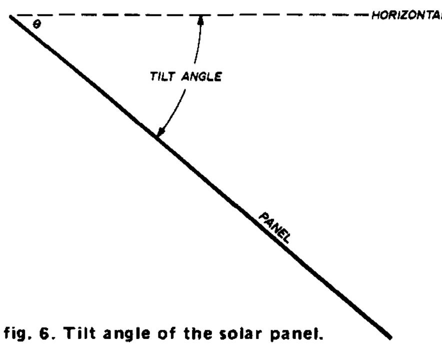


fig. 6. Tilt angle of the solar panel.

Thank you, Edgar, and Austin Forsyth, G6FO, Managing Editor of *Short Wave Magazine*.

My contact with WB4ZXJ on April 15, 1974, is degraded to the first solar-powered ssb contact, not arranged. Maybe? The paramount question, however, is why we have waited so long (except for satellite communications) to take advantage of this limitless source of energy.

new applications

More small companies are now producing small solar energy converters. Zurn Industries* sells two models with peak ratings of 1.5 and 6 watts. These two 12-volt models are covered with a transparent waterproof and detergent-proof coating. A major application of the Zurn models is for marine batteries. Under the condition of a week of normal daylight, sufficient charging power is available that

*Zurn Industries Incorporated, 5533 Perry Highway, Erie, Pennsylvania 16509.

a bilge pump can pump 600 gallons of water with the small energizer and 3000 gallons with the larger model. Such a device is particularly useful for maintaining battery charge during long idle periods, as during winter storage of a small boat.

Another active organization, Solar Energy Company,† sells solar power converters as well as wind-powered electrical sources. Their devices find application in vhf/uhf repeaters, microwave relays, wire and wireless telephone systems, TV translators, monitors, offshore platforms, data buoys, railroad signals and controls, plus traffic and security systems. They also make a unit that maintains the charge on batteries used to power farm electric fences.

Why haven't such devices been used to supply a portion of the power needed in the average household? Needed is a solar heater for the home with its electrical blowers, fans and circulators powered by a solar energy converter. At least there should be a home furnace capable of using a variety of fuels with its electrical accessories powered completely by battery and solar power converters.

experimental approaches

There are several avenues of experimentation. What are the operating limits using direct drive to 12-volt devices and no battery? Some voltage-regulator device would be required. How bright would it have to be to provide operation of a 12-volt zener diode? How much longer a period of operation would be feasible using the solar source for 9-volt regulated operation? What are the limits of a particular solar panel in terms of higher power demand and keeping a higher capacity battery fully charged relative to your normal operating schedule?

Connecting solar panels in parallel increases current capability. How many panels would be required to provide adequate charging current for a high capacity battery system? How many panels and what battery capacity would provide enough power to match normal operating time for your 200-watt sideband transceiver? What advantages are to be obtained from series-parallel groupings of solar panels?

†Solar Energy Company, 810 18th Street, NW, Washington, D.C. 20006.

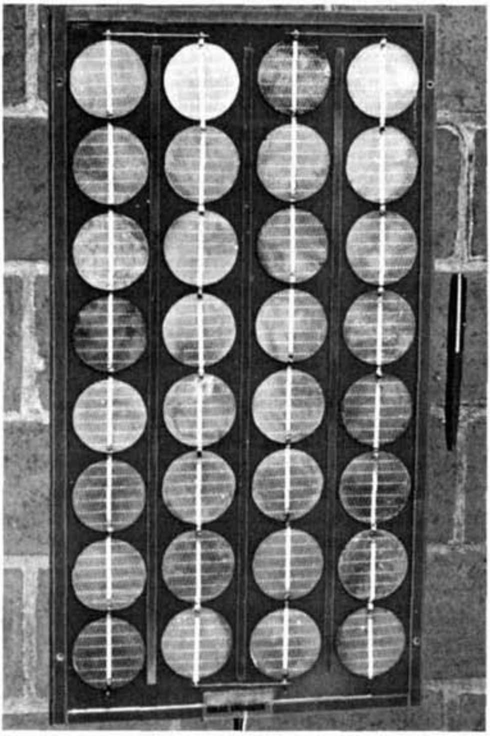
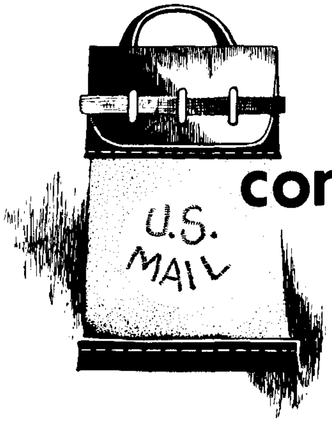


fig. 8. 12-volt, 500-mA solar panel (courtesy Solar Energy Company).

What additional capacity can be gained by using a mechanical mount that would permit you to chase the sun across the sky? Would an equatorial mount and clock-drive be practical? How would you hold the weight of the assembly down? What percentage of the derived power would be needed to power the drive system? Perhaps drive power could be conserved by changing positions only once each hour, or half-hour. I will be discussing these and similar subjects in the months ahead.

ham radio



comments

coherent fsk RTTY

Dear HR:

I would first like to commend Steve Maas, K3WJQ, for his article bringing modern communication theory into the area of amateur RTTY in the June, 1974, issue of *ham radio*. However, I feel that a few areas need comment. First, the numbers in **table 1** (page 32) for coherent FSK probability of error are slightly in error. The published data is based on a shift between the mark and space frequencies of $f_s \approx 0.7/T$, where T is a baud time.[1] For 60 wpm RTTY, T = 0.022 second, resulting in an f_s of only 32 Hz. It can be shown that this shift results in the lowest probability of error for coherent demodulation of FSK. However, for many practical reasons, this is too narrow a shift, so for standard 170- or 850-Hz shifts, that column should be amended to

input snr	probability of error coherent FSK
0 dB	0.16
2 dB	0.10
3 dB	0.08
6 dB	0.02
9 dB	0.002

The probability of error, as well as the Rayleigh and Rician probability density functions, are derived assuming no fading, with white noise having a Gaussian density function at the inputs of the mark and space filters. This is sometimes a good assumption, but primarily at vhf and above. Since most RTTY activity is in the high-frequency spectrum, this white Gaussian noise assumption needs examination.

The National Bureau of Standards has attempted to model atmospheric noise.[2] They have found that the performance of an FSK demodulator can be several orders of magnitude worse than calculated by the white Gaussian noise assumptions, and this study didn't include QRM, which seems to be the most prevalent type of "noise" in our crowded bands.

The high-frequency spectrum is time varying with frequency-selective fading (QSB). This fading results in a tremendous degradation in the probability of error performance. Fading is thought to be due to constructive and destructive combining of various received signals that have been reflected from different layers of the ionosphere, resulting in a received signal with varying amplitude and with a random phase angle. So, while the transmitted phase may be constant, because of the varying time delays through the propagation path, the received signal can have any phase. It appears that demodulation schemes which require phase information, such as coherent FSK, are degraded *more* than noncoherent schemes.[3]

By requiring that a PLL (**fig. 4** in K3WJQ's article) remain close to the previously transmitted phase while the

1. Schilling and Taub, *Principles of Communications Systems,* McGraw-Hill, New York, 1970.
2. Alyce M. Conda, "The Effect of Atmospheric Noise on the Probability of Error for an NCFSK System," IEEE Trans. on Communications Techniques, September, 1965, page 280.
3. G.L. Turin, "Error Probabilities for Binary Symmetric Ideal Reception Through Nonselective Slow Fading and Noise," Proceedings of IRE, September, 1958, page 1603.

other channel is being transmitted would result in a very long time constant in the lowpass filter. This would also have the effect of increasing the time it takes for the PLL to lock up. PLLs in amateur demodulators have usually had the additional problem of locking on to a strong nearby interfering signal. Finally, there is evidence that the use of two oscillators at the transmitter end, which results in discontinuities in the phase of the transmitted signal, will have poorer performance than shifting the frequency of a single oscillator.[4]

Most mathematical analysis of systems like the one described in K3WJQ's article ignore the effect of one bit on the next. This problem is called intersymbol interference, and is found in any receiver having narrow bandpass filters. The degrading effects of intersymbol interference are extremely difficult to analyze, but they are of concern, as evidenced by the multitude of articles on attenuation and delay equalizers in the technical journals.[5]

The major causes of errors in an amateur's RTTY system thus seem to be those caused by fading, high atmospheric noise levels, interfering stations and intersymbol interference introduced by the propagation medium and receiver. **Table 1**, while slightly in error in practice, does point up the fact that a small increase in antenna gain can tremendously improve the performance. Also, for those fortunate hams with forty acres and unlimited resources, space diversity antenna systems reduce the effects of fading. Other known techniques for removing the effects of interference or fading, such as frequency diversity, error correcting codes and spectrum spreading methods, aren't looked upon favorably by the FCC.

4. P.A. Belle and B.D. Nelin, "The Effect of Frequency-Selective Fading on the Binary Error Probability of Incoherent and Differentially Coherent Matched Filter Receivers," IEEE Trans. on Communications Systems, June, 1963, page 170.
5. A. Gersho, "Adaptive Equalization of Highly Dispersive Channels for Data Transmission," Bell System Technical Journal, January, 1969, page 55.

I hope that the above comments will stimulate thinking on the design of FSK demodulators. The articles, such as the ones referenced, should provide indications as to the directions a design should take. I hope that I have shown that while I am enthusiastically for more theoretical articles in the amateur magazines, they should be tempered with the idea that mathematical analysis of various systems can be, at best, extremely difficult. The only way that a majority of amateurs, including myself, will be convinced of a new method's effectiveness is by seeing a working model perform as well or better than existing models.

John Fehlauer, WA2WTL
Gibbstown, New Jersey

bequests amateur gear

Dear HR:

Your March editorial was interesting and to the point regarding the disposition of a deceased ham's equipment. Specifically, it takes into account the fact that there is a big difference between *good* junk and *junk*. Too many of today's hams consider homebuilt equipment to be in the latter category, but I think I have the solution.

My will specifies that all of my equipment which can be categorized as amateur radio equipment and support parts shall be made available to the County Board of Education, with specific instructions that such items be sent to the local high school (which happens to have an amateur station). Further, I have specified that any taxes, fees and/or delivery costs be paid from my estate because I wanted to prevent any failure of delivery because of some cost or expense that the Board of Education might elect not to pay.

There are no strings attached, the high school will receive some valuable equipment and—hopefully—some youngster will learn something from the bequest.

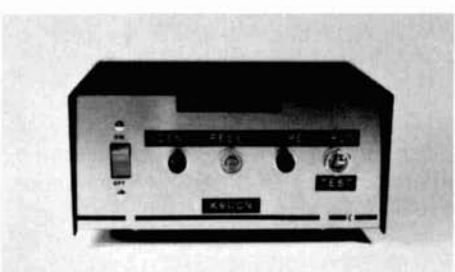
Dean Young, W3FZ
Adelphi, Maryland

the ham notebook

identification timer

Lengthy QSOs and time-consuming phone patches can lead to problems with the FCC if you do not identify your station every ten minutes. After becoming an Official Observer I became aware of times between ID—especially those of ragchewing stations and stations running phone patches. Operators in these two categories tend to let time ramble by without the proper 10-minute identification. To alleviate this problem at my station I set out to build a timer that would accurately reproduce a 10-minute time out, have a minimum number of parts, and could be assembled in one evening. The circuit is shown in **fig. 1**.

This timer was designed around the Signetics NE555 integrated circuit. Any 9- to 12-volt dc supply can be used to run the timer. Once the power supply is turned on, the red lamp (*ident*) will light. The *run-test* switch is placed in the *test* position and the *reset* button should be depressed and released. A ground at pin 2 will cause the green lamp to light and the red lamp to extinguish—indicating that the unit is *timing*. With the values shown the *test* position will allow a time cycle of 2 to 3 seconds after which the lamps will change from green to red (*ident*). Now



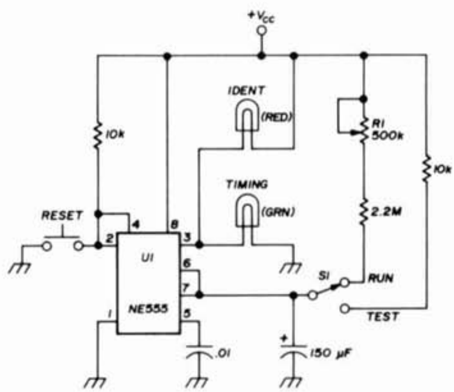


fig. 1. Simple ten-minute ID timer uses low-cost NE555 timer IC. Timing period, adjustable from 7 to 11 minutes, is set by R1.

place switch S1 in the *run* position. Again depress and release the *reset* button. The unit will time out between 7 and 11 minutes, depending on the setting of R1. I have R1 (500k pot) on my unit set to time out at 9 minutes. This allows a one-minute time frame in which to identify.

Even with ±10% variation from nominal supply voltage of 12-volt dc the timer retains a 9 minute, ±1 second, accuracy. If during a timing cycle you wish to identify your station and return to zero time it is only necessary to depress and release the *reset* switch. It is recommended that the indicator lamps draw no more than 100 mA each so as not to strain the current-sinking ability of the NE555 (200 mA max).

The total cost to build this timer is under five dollars. Circuit layout is not critical and component minimization makes the unit very reliable. With a timer like this one next to your phone patch you don't have to worry about being cited for lack of a correct 10-minute ID.

Don Backys, K9UQN

spurious signals

The October, 1973, issue of *ham radio* (page 67) carried details of Yaesu FTdx560/570 vfo frequency ranges, and the incorporation of the 6.358.6-kHz suck-out crystal in recent production. This was summarized in the December, 1972, issue (page 69), based on JA1MP's letter earlier in the year. This modification eliminated a 28-MHz spurious, but does not appear to affect the clean lower-sideband phone signals appearing in the 14-MHz DX band in a manner similar to those reported from the Yaesu FTdx400.

W6PKK added the crystal to his FTdx560, then retrimmed traps L18 and L19 (but not L23). His 14270-kHz upper-sideband transmissions caused lower-sideband spurious on 14080 kHz. Trimming eliminated the spurious completely at my location. In the case of a W7 who installed the suck-out crystal ($5 plus COD postage), the 14197 lower-sideband spurious caused by the 14212 kHz upper-sideband transmission was not reduced; it presumably requires adjustment of the two traps as accomplished by W6PKK.

It appears that all Yaesu amateur equipment may have one or more traps which must be properly adjusted and the performance confirmed by listening tests.

A number of CW signals heard in the phone band may have been spurious from the CW band on 14 MHz, but were not all investigated when heard. One was on about 14345 kHz. Several RTTY signals around 14085/14090 kHz have also been heard in the vicinity of 14220 kHz, these being from HW100, SB101, SB400 and SB401 equipment, according to information received. Some of these sets appear to depend upon a "bandpass coupler" out of the first mixer to suppress any undesired second harmonics of the input frequencies, or of the resulting output frequency. If the bandpass coupler is not sufficient, suitable traps might be applied to the first mixer output or the second mixer input (if two mixers are incorporated) to suppress the undesired products.

Information should be accumulated on the spurious performance of all types of amateur equipment, and the necessary cure for any spurious emissions or responses, to eliminate unnecessary receiver interference.

Bill Conklin, K6KA

full-quieting meter for Clegg 27B

Many amateurs who use the Clegg 27B as a base rig have wanted a visual indication of incoming signal strength for peaking up mobile signals and for keying up repeaters. This can be simply accomplished as follows: Connect one end of a piece of miniature RG-74/U coaxial cable, center conductor to the orange wire on the squelch control and braid to the yellow wire of the squelch control. Run the RG-174/U coax back to the hole near the antenna connector and connect it to a miniature 1/8-inch female connector. For the meter, which is connected to the coax with a matching 1/8-inch male connector, I used a 0-500 microampere meter. A 0.01-μF capacitor across the meter will smooth out any needle movement. Make sure the cable braid is not grounded at any point as this will bypass the squelch circuit and make it inoperative.

Tom Clerk, WA2YUD

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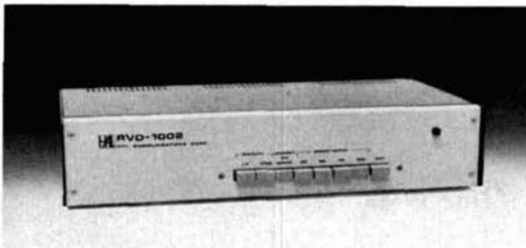

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The DKB-2010 is equally versatile in the CW mode, with complete alphanumeric and punctuation keys, speeds from 8-60 WPM, and a "DE-call sign" key. The DKB-2010 includes a three-character buffer operational in either the RTTY or CW mode. Optional 64 or 128 key buffer also available.

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Price: ST-6 $310 Assembled, $147.50 Kit, ppd USA. 425 Hz Discriminator $40 Assembled, $29 Kit, ppd USA. AK-1 AFSK $40 Assembled, $29 Kit, ppd USA. Air shipment: Assembled ST-6 with any or all options $10, ST-6 Kit $4, 425 Hz Kit $1, AK-1 Kit $1.

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More Details? CHECK—OFF Page 94

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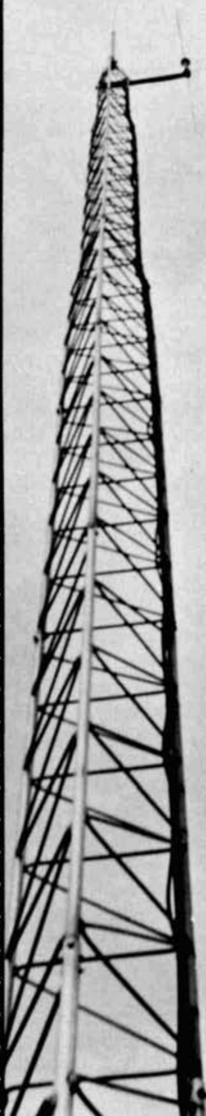

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Panoramic SPA-4 Spec. Anal. Bandswitching. 10MHz-44gHz, 5 or 70MHz sweep width	985
Polarad MSG34-Sig. Gen 4.2-11gHz calib attn. AM-FM-Pulse mod.	495
Polarad R uwave rcvr .4-84gHz with plug-in AM, FM, CW, Pulse - less plug-in	225
Polarad TSA Spec. Anal .01-44gHz with plug-in less plug-in	125
Solitron 200A SCR tester-checks anode, gate volts,current,leakage and holding	165
Stoddart NM10A (URM-6) RF intens mtr 10-250 kHz, complete with acc.	630
Stoddart NM20A (PRM-1) RF intens mtr .15-25MHz, complete with acc.	655
Stoddart NM52A-RFI mtr, .375-1gHz, w/acc.	985
Tek RM 15-DC-15MHz GP scope	265
Tek 181 Time mark generator	95
Tek 190A Const. Ampl. Sig. Gen .35-50MHz	125
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Output Power at Max. i/p. Min. Typical	12 W. 14 W.	10 W. 12 W.	35 W.	20 W.
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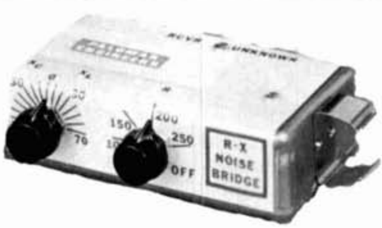

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222.38	223.94	146.12	146.61	52.70	52.60
223.14	223.98	146.16	146.70	52.76	52.64
223.26	224.74	146.19	146.72	52.82	52.68
223.30	224.86	146.22	146.76	52.88	52.72
223.34	224.90	146.25	146.79	52.92	52.79
223.50	224.94	146.28	146.82	52.96	52.80
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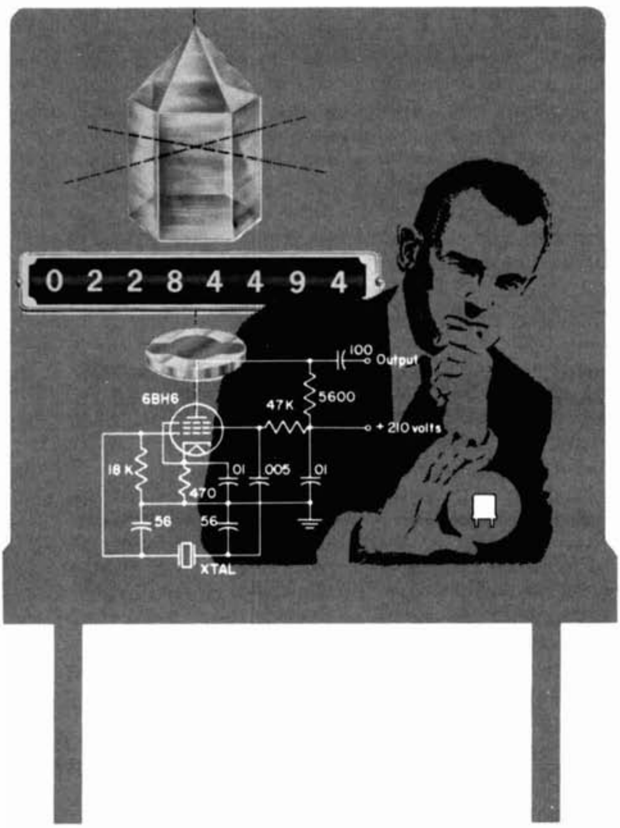

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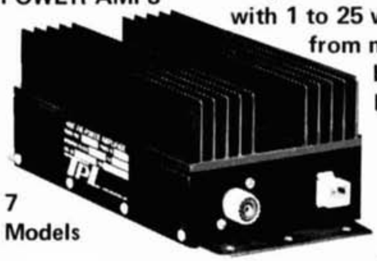

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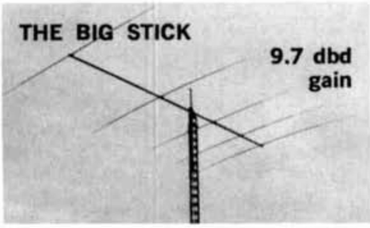

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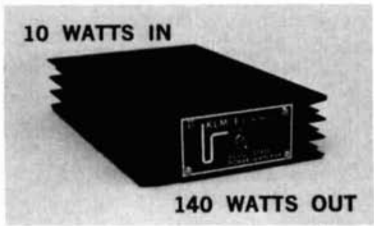

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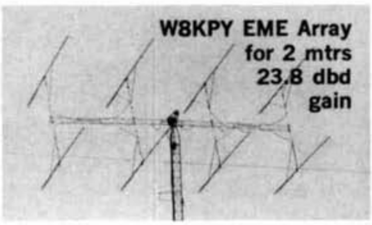

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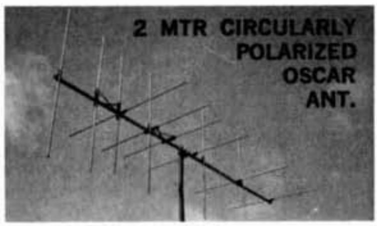

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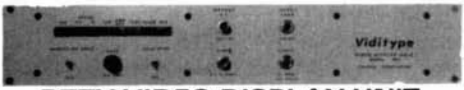

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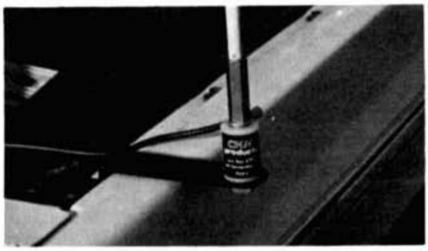

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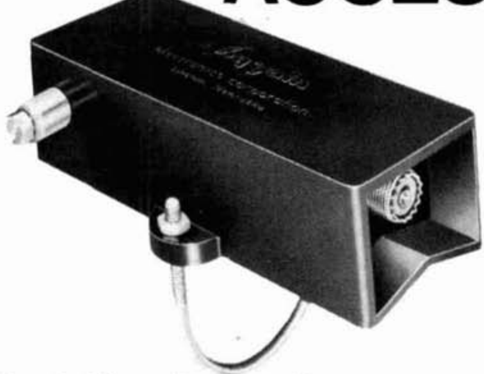

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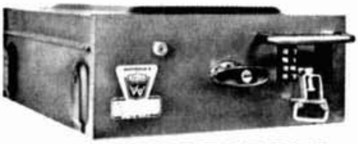

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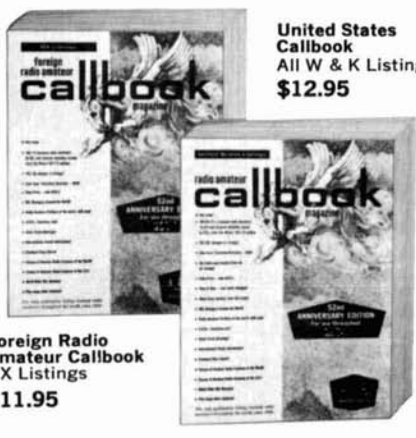

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INDEX

ATV *
Adva ___ 265
Amtech ___ 006
Apollo ___ 011
Barry *
Budwig ___ 233
CHM ___ 273
CFP ___ 022
Caddell ___ 247
Columbia ___ 212
Comm. Prod. & Eng. ___ 268
Craig ___ 177
Curtis ___ 034
Cush Craft ___ 035
Data Signal ___ 270
Dentron ___ 259
Dycomm ___ 040
Dynamic Elect. ___ 041
ECM ___ 190
Eimac ___ 043
Electronics Center ___ 236
Electronic Dist. ___ 044
Epsilon ___ 046
Erickson ___ 047
Fair ___ 048
G & G Radio *
General Aviation ___ 168
Gray ___ 055
Gregory *
Hal ___ 057
Hal-Tronix ___ 254
Hamtronics ___ 246
Ham Radio ___ 150
Heath ___ 060
Heights ___ 061
Henry ___ 062
House of Dipoles ___ 036
Hy-Gain ___ 064
Icom ___ 065
International Crystal ___ 066
Jan ___ 067
Janel ___ 068
K. E. ___ 072
K-Enterprises ___ 071
KLM ___ 073
Larsen ___ 078
Leland ___ 193
Logic ___ 133
Madison ___ 083
Martex ___ 197
Matric ___ 084
Pagel ___ 092
Palomar ___ 093
Poly Paks ___ 096
Porta-Pak ___ 274
RMS ___ 239
RP ___ 098
Callbook ___ 100
Raytheon ___ 101
Regency ___ 102
Robot ___ 103
SGC ___ 275
Saroc ___ 146
Savoy ___ 105
Signal/One ___ 262
Space-Military ___ 107
Spectronics FM ___ 191
Spectrum ___ 108
Sumner Elect. & Eng. ___ 276
Swan ___ 111
Technical Doc. ___ 226
Tri-Tek ___ 117
Tristao ___ 118
VHF Engineering ___ 121
Valu-Pak ___ 264
Webster ___ 255
Weinschenker ___ 122
World QSL ___ 125

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November 1974

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Advertisers index

ATV Research 93
Adva Electronics 84
Amtech 76
Apollo Products 77
Barry 96
Budwig Manufacturing Co. 76
CHM Products 89
CFP Enterprises 87
Caddell Coil Corp. 88
Columbia Electronics 93
Communication Products & Engineering 87
Craig Radio 90
Curtis Electro Devices 85
Cush Craft 95
Data Signal, Inc. 31
Dentron Radio Co. 7
Dycomm 86
Dynamic Electronics, Inc. 93
ECM Corporation 76
Eimac, Div. of Varian Assoc. .... Cover IV
Electronics Center, Inc. 51
Electronic Distributors 85
Epsilon Records 90
Erickson Communications 76
Fair Radio Sales 87
G & G Radio Supply Co. 88
General Aviation 72, 73
Gray Electronics 64
Gregory Electronics 92
Hal Communications Corp. 62, 63
Hal-Tronix 86
Hamtronics, Inc. 89
Ham Radio 85, 93
Heath Company 48A-F
Heights Manufacturing Co. 64
Henry Radio Stores .... Cover III
House of Dipoles 84
Hy-Gain Electronics Corp. 66, 74, 81, 91
Icom 69
International Crystal Mfg. Co. Inc. 75
Jan Crystals 70
Janel Labs 76
K. E. Electronics 89
K-Enterprises 70
KLM Electronics 78
Larsen Electronics 43
Leland Associates 89
Logic Newsletter 86
Madison Electronic Supply 86
Martex Corporation 68
Matric 70
Pagel Electronics 88
Palomar Engineers 68
Poly Paks 82
Porta-Pak 64
RMS Corporation 84
RP Electronics 86
Radio Amateur Callbook, Inc. 84, 92
Raytheon Company 5
Regency Electronics, Inc. 37
Robot Research, Inc. 1
SGC, Inc. 80
Saroc 65
Savoy Electronics 25
Signal/One 2
Space-Military Electronics 90
Spectronics FM 67
Spectrum International 68
Sumner Electronics & Engineering 90
Swan Electronics .... Cover II
Technical Documentation 88
Tri-Tek, Inc. 90
Tristao Tower Co. 93
VHF Engineering, Div. of Brownian Elect. Corp. 79
Valu-Pak 89
Webster Radio 93
Weinschenker, M. 71
World QSL Bureau 84

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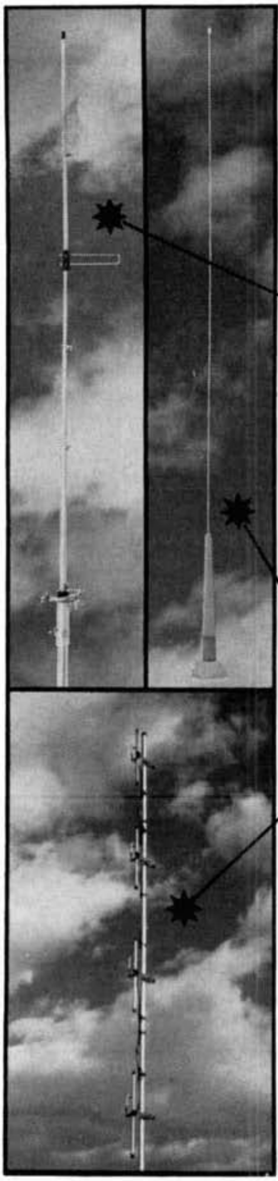

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FM 2 METER ANTENNAS

FROM THE WORLD'S LEADING MANUFACTURER OF VHF/UHF COMMUNICATIONS ANTENNAS

NEW FM GAIN RINGO RANGER – Get extended range with this exciting new antenna. A one eighth wave phasing stub and three half waves in phase combine to concentrate your signal at the horizon where it can do you the most good. Your present AR-2 can be extended with a simply installed RANGER KIT.

ARX-2	100 watts	146-148	MHz
ARX-220	100 watts	220-225	MHz
ARX-450	100 watts	435-450	MHz

ARX-2K RANGER KIT

NEW FM MOBILE – Fiberglass 5/8 wave professional mobile antenna for roof or trunk mount. Superior strength, power handling and performance.

AM-147T 146-175 MHz mobile

NEW 4 POLE – A four dipole gain array with mounting booms and coax harness 52 ohm feed, 360° or 180° pattern.

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More Details? CHECK—OFF Page 94

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Tempo's VHF transceivers offer commercial performance at amateur prices. Compare these transceivers with any other available. Compare their performance, their quality of construction, their ease of maintenance, and then compare prices. Your choice will have to be Tempo.

FMH-MC for Marine & Commercial service also available.

TEMPO/fmh

So much for so little! 2 watt VHF/FM hand held. 6 Channel capability, solid state, 12 VDC, 144-148 MHz (any two MHz), includes 1 pair of crystals, built-in charging terminals for ni-cad cells, S-meter, battery level meter, telescoping whip antenna, internal speaker & microphone. $199.00

TEMPO/CL 146

The CL-146 offers operation on the 146 MHz amateur band. The price includes a microphone power cord, mounting bracket and one pair of crystals. A full line of accessories is also available. • 12 channel capability • 13 watts or a power savings 3 watts • All solid state, 12 VDC • 144 to 148 MHz (any two MHz without retuning) • Supplied with one pair of crystals • RF output meter, S-meter, receiver detector meter • Provisions for external oscillator • Monitor feature • Audio output at front panel •Internal speaker • The Price: $299.00

TEMPO/CL 220

As new as tomorrow! The superb CL-220 embodies the same general specifications as the CL-146, but operates in the frequency range of 220-225 MHz (any two MHz without retuning). At $299.00 it is undoubtedly the best value available today.

TEMPO
VHF/UHF Power Amplifiers

Solid state power amplifiers for use in most land mobile applications. Increase the range, clarity, reliability and speed of two-way communications.

Drive Power	Output	Model No.	Price
VHF (135 to 175 MHz)			
2W	130W	130A02	$199
10W	130W	130A10	$179
30W	130W	130A30	$189
2W	80W	80A02	$169
10W	80W	80A10	$149
30W	80W	80A30	$159
UHF (400 to 512 MHz)			
2W	70W	70D02	$270
10W	70W	70D10	$250
30W	70W	70D30	$210
2W	40W	40D02	$180
10W	40W	40D10	$145
2W	10W	10D02	$125

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SPECIFICATIONS

SB-104 SPECIFICATIONS — TRANSCEIVER SECTION — GENERAL OPERATION: Frequency Coverage: 3.5 MHz through 29.7 MHz amateur bands, 15 MHz WWV receive only. **Frequency Stability:** Less than 100 Hz/hr drift after 30-min. warmup; less than 100 Hz drift for ±10% change in primary voltage. **Modes of Operation:** Selectable upper or lower sideband (suppressed carrier) and CW. **Readout Accuracy:** Within ±200 Hz ±1 count. **Audio Frequency Response:** 350 to 2450 Hz ±75 Hz (6 dB bandwidth). **Dial Backlash:** 50 Hz max. **Phone Patch Impedance:** 4 ohm output to speaker; high impedance output to transmitter. **Power Requirements:** 13.8 VDC nominal (max. 16 VDC) at: Receive: 2 amp. Transmit: low power: 3 amps.; high power: 20 amps. **TRANSMITTER: RF Power Output: High Power** (50 ohm non-reactive load). SSB: 100 watts PEP ±1 dB; CW: 100 watts ±1 dB. **Low Power** SSB: 1 watt PEP (minimum); CW: 1 watt (minimum). **Output Impedance:** 50 ohms, less than 2:1 SWR. **Carrier Suppression and Unwanted Sideband Suppression:** 55 dB down from 100 watt single-tone output at 1000 Hz reference. **Harmonic Radiation:** 45 dB below 100 watt output. **Spurious Radiation:** −50 dB within ±3 MHz of carrier; −60 dB farther than ±3 MHz from carrier, except −40 dB at 3.39 MHz on 80 meter band. **Third Order Distortion:** 30 dB down from two-tone output, reference at 100 watts PEP. **Transmit/Receive Operation:** SSB: PTT or VOX; CW: Keyed-tone VOX or manual. **CW Side-Tone:** Internally switched to speaker or headphones in CW mode. Approximately 700 Hz tone. **Microphone Input:** High impedance with a rating of −45 to −55 dB; approx. 25K ohms to match Heath desk-type microphone. **RECEIVER — Sensitivity:** Less than 1.0 microvolt for 10 dB signal-plus-noise-to-noise ratio for SSB operation. **Selectivity:** 2.1 kHz minimum at 6 dB down, 5 kHz maximum at 60 dB down. (2:1 nominal shape factor). **CW Selectivity:** (with accessory CW filter) 400 Hz at 6 dB down; 2 kHz max. at 60 dB down. **Overall Gain:** Less than 1 microvolt for 0.5 watt audio output. **Audio Output:** 2.5 watts into 4 ohms, 1.25 watts into 8 ohms, at less than 10% THD. Low impedance headphones (4-8 ohm). **AGC:** Less than 1 millisecond attack time; switch selectable 100 μsec. and 1 msec. release, and OFF. **Intermodulation Distortion:** −65 dB min. **Image Rejection:** −60 dB min. **IF Rejection:** −60 dB min. **Internally Generated Spurious:** Below 2 microvolt equivalent antenna input, except at 3.65, 3.74, and 21.2 MHz. **MECHANICAL — Front Panel Controls/Switches:** AGC — Off, Slow, Fast; AF Gain; Microphone Jack; Headphone Jack; Main Tuning; Mic/CW Level; Vox Gain; Vox Delay; Band Switch. **Pushbuttons:** ALC (Meter); 13.8V (Meter); Relative Power (Meter); 100 Hz (Disable); Noise Blanker (On/Off); LSB (Mode); USB (Mode); CW (Mode); Tune; Hi/Lo (Power Select); VOX (On/Off); PWR (On/Off). **Rear Panel Controls/Sockets:** Anti-Trip; Sidetone Level; Linear Amplifier ALC Input; Phone Patch Input; Phone Patch Output; Key (CW) Input; Speaker (4 ohm) Output; Spare (2); Receiver Audio Input; VFO Input; VFO Output; IF Output; Driver Output; Ground Post; Power Plug; Accessory Socket (includes relay output); Antenna Input; Receiver Antenna Input; Common/Separate Antenna Switch. **Dimensions:** 5¾″ H x 14$\frac{15}{32}$″ W x 13⅞″ D. (Less knobs, feet and connectors). **Weight:** 20 lbs.

The SB-104 output board and final transistors are warranted for one full year.

HP-1144 SPECIFICATIONS — Output Voltage: 13.8 VDC regulated (Adjustable from approximately 11 to 16 VDC). **Maximum Output Current:** 20 amperes, Intermittent. 8 amps continuous. **Transistor Integrated Circuit and Diode Complement:** 2N3643 transistor; 2N3055 transistor; 40411 pass transistor (2); MFC6030 regulator IC; 1N4002 silicon diode; MDA990-2 bridge rectifier. **Power Requirements:** 110 to 130 VAC @ 6A or 220 to 260 VAC @ 3A, 50/60 Hz maximum. **Dimensions:** 5½″ H x 9¼″ W x 10¼″ D. **Regulation:** Less than 2% output voltage variation from no load to 20 amperes. **Ripple:** Less than 1% at 20 amperes. **Fuses:** 7-amp, 3AG, slow-blow primary. 20-amp, 3AG, output. **Net Weight:** 23 lbs.

SB-604 SPECIFICATIONS — Speaker Size: 5″ x 7″ oval. **Voice Coil Impedance:** 3.2 ohms. **Frequency Response:** 300 to 3000 Hz. **Magnet Weight:** 3.16 oz. **Cabinet:** Aluminum with gray wrinkle finish. **Dimensions:** 7⅛″ H x 10⅛″ W x 14″ D.

SB-644 SPECIFICATIONS — Frequency Coverage: 5.0 — 5.5 MHz allowing 80, 40, 20, 15, 10 meter operation in the SB-104. **Frequency Stability:** Less than 100 Hz drift per hour after thirty minute warmup. **Modes of Operation:** Remote VFO; Main VFO; Receive Remote/Transmit Main; Receive Main/Transmit Remote; Crystal frequencies (2) (crystals not supplied). **Dial Backlash:** 100 cycles max. **Power Requirements:** 11V and 13.6V at 500 mA from the SB-104. **RF Output:** 0.34 to 0.4V RMS over 5 to 5.5 MHz into a 50 ohm load.

SB-634 SPECIFICATIONS — CLOCK — Display: Six full digits. **Time Base:** 24 hours. **Accuracy:** Determined by accuracy of power line frequency. **TIMER — Display:** Three full digits. **Time Interval:** 10 minutes with automatic reset. Manual reset at any portion of 10-minute period. **Accuracy:** Determined by accuracy of power line frequency. **Signal:** Visual only or both visual and aural; switch selected. **RF POWER/SWR METER — Frequency Range:** 1.8 to 30 MHz. **Wattmeter Accuracy:** ±10% of full-scale reading. **Power Handling Capability:** 2000 watts (maximum). **SWR Sensitivity:** Less than 10 watts. **Impedance:** 50 ohm nominal. **SWR Bridge:** Continuous to 2000 watts P.P. **Connectors:** UHF type SO-239. **PHONE PATCH — Circuit:** Telephone hybrid circuit. Allows voice control or manual operation. **TELEPHONE LINES — Input Impedance:** Approximately 600 ohm. **Null Depth:** At least 30 dB isolation between transmit and receive circuits. **Receiver Impedance:** Effective match from 3 to 16 ohm. **Transmitter Impedance:** 600 ohm or higher impedance output. **GENERAL — Meter:** 100μA movement. VU readings for phone patch monitoring. Null depth indication. RF power output, relative power, and SWR readings. **FRONT PANEL CONTROLS — Timer:** Off, Visual, Aural Visual. **Reset:** Pushbutton switch. **Patch Gain:** Transmitter, Receiver. **SWR:** Sensitivity. **Mode:** SWR, Forward and Reflected. 2000 W and 200 W. Phone Patch. **Rear Panel Controls — Clock:** Time hold, minutes set, seconds set. Null Adjust control; Null-Monitor switch; C adjust control; R adjust control. **Power Requirements:** 120/240 VAC, 50/60 Hz, 15 watts.

SB-614 SPECIFICATIONS — RF SAMPLING SECTION: Frequency Coverage: 80 through 6 meters (3.5 — 54 MHz). **RF Power Limits:** Exciter input (50 — 75 ohm) 10 to 300 watts; Antenna input (50 — 75 ohm) 10 to 1000 watts (up to 1500W PEP). **Insertion Loss:** Negligible. **VERTICAL AMPLIFIER: Input Impedance:** 1 Megohm shunted by 75 pf. **Sensitivity:** 60 mV rms/¼″ vertical deflection. **Attenuator:** 2 position; x1, 2 volts rms max. input; x10, 20 volts rms max. input. **Frequency Response:** 10 Hz to 50 kHz ±3 db. **HORIZONTAL AMPLIFIER: Input Impedance:** 1 Megohm shunted by 50 pf. **Sensitivity:** 50 mV rms/¼″ horizontal deflection. **Frequency Response:** 10 Hz to 3 MHz ±3 dB. **SWEEP GENERATION: Type:** Recurrent, automatic sync. **Frequency Range:** 10 Hz to 10 kHz in three ranges. **GENERAL: CRT:** 3RP1/A flat face, green, medium persistence phosphor. **Graticule:** .250 inch squares 6 x 8 (1.5 x 2.0 inches total viewing area). **Power Supplies:** All solid-state rectifiers. All amplifier supplies regulated. **Power Requirements:** 110-130 or 220-260 VAC, 50/60 Hz, 35 watts. **Front Panel Controls:** Intensity — Off-on; Mode — SSB, TRAP, CROSS; Focus; Vertical Gain; Vertical Position; Horizontal Gain; Horizontal Position; Sweep — variable; Range — 100 Hz, 1 kHz, 10 kHz. **Rear Panel Controls:** Astigmatism; Vertical attenuator —X1, X10. **Rear Panel Connectors:** Antenna: SO-239; Exciter: Phono; Vertical Input: Phono; Horizontal Input: Phono. **DIMENSIONS:** 7¼″ H x 10¼″ W x 15¼″D. **Net Weight:** 12 lbs.

SB-230 SPECIFICATIONS: Band Coverage: 80, 40, 20, 15 and 10 meter amateur bands. **Maximum Power Input:** 1200 W PEP SSB; 1000 W CW; 400 watts RTTY/SSTV. **Duty Cycle: SSB:** continuous voice modulation; **CW:** continuous (max. key-down time 30 seconds); **RTTY/SSTV:** 50% (max. transmit time 10 minutes at 400 watts). **Driving Power Required:** less than 100 W. **Third Order Distortion:** −30 dB or better. **Output Impedance:** 50 ohms at 2:1 SWR max. **Input Impedance:** 52 ohms at 1.5:1 SWR max. **Meter Switch:** Exciter only; Relative Power; Plate Current; Grid Current; High Voltage. **Front Panel:** Load; Tune; Band; Relative Power sensitivity; Power switch; Meter switch. **Rear Panel:** ALC output; Exciter relay; RF input; RF output; Ground lug; Fuse; Line cord. **Tube:** Type 8873. **Zero signal plate current:** 25 mA. **Power Requirements:** 120 VAC, 50/60 Hz, 14 A max. 240 VAC, 50/60 Hz, 7 A max. **Dimensions:** 14¾″ W x 16″ D x 7″ H. **Net Weight:** 33½ lbs.

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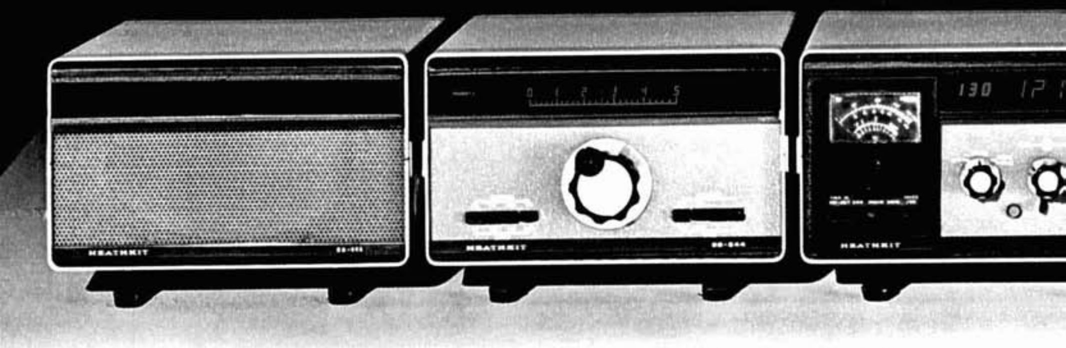
NEW SYSTEM-ENGINEERED

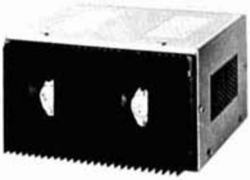


SB-604 Station Speaker

Designed and styled to match the new SB-104 Transceiver, the cabinet is large enough to house the HP-1144 AC Power Supply. The 5" x 7", 3.2 ohm speaker is response-tailored for SSB. Connector cable & plug are included.

Kit SB-604, Speaker & cabinet, 8 lbs., mailable **29.95**

HP-1144 Fixed-Station AC Power Supply



This new 120 V/240 VAC operated supply provides the 13.8 VDC required by the new SB-104 Transceiver. The full-wave bridge circuit has triple Darlington regulation with an integrated circuit which samples, compares, and automatically adjusts transistor bias to maintain a fixed output level. Output is remotely sampled at the load end of the power cable, thereby compensating for voltage drop across fuse and cable, to provide almost no change in voltage from no load to full load conditions. A cable and socket provide output power and a series connection to the SB-104 remote on-off switch. The generous heat sink fits on the back of the supply, and the entire unit may be mounted within the SB-604 speaker cabinet.

Kit HP-1144, fixed-station supply, 28 lbs., mailable **89.95**

SB-634 Station Console

Five station accessories in one!

24-hour digital clock: six half-inch gas discharge digits indicate hours, minutes and seconds. The clock runs continuously, as long as the console is plugged in.

Ten-minute ID timer: Three gas discharge digits indicate minutes and seconds up to 9:59. At ten minutes the timer recycles and provides either a visual alarm or both visual and audible alarms. Pushbutton zero reset.

RF wattmeter: The big meter delivers measuring capability of either 200 watts or 2000 watts full scale. 160 through 10 meters.

SWR bridge: Push a button to measure SWR. Separate front panel SWR sensitivity control.

Phone patch: The hybrid patch can be used either manually or with VOX control without switching connections. VU capability on the meter and separate front panel controls to adjust transmitter and receiver gain independently. Line isolation can be adjusted with a rear panel control.

Kit SB-634, 14 lbs. **179.95**

SB-644 Remote VFO

Designed exclusively for the new SB-104. The new SB-644 provides serious DXers with really useful split transmit/receive capability. With the "104/644" combination, you aren't frequency limited in any way — the transceiver can be at one end of the band, the remote VFO at the other end.

Multi-mode capability. The "644" allows transceive operation on either itself or the "104"... transmit on the "104" and receive on the "644" ... receive on the "104" and transmit on the "644". And you can use either of the two crystal positions in the "644" for fixed-frequency control.

Easy pushbutton operation. Front panel pushbuttons on the "644" control all transceive, transmit and receive modes on both the "104" and the remote VFO. No switching on the "104" is necessary. Status lamps behind the window indicate frequency-control mode.

Digital readout in the SB-104. Although the SB-644 includes a linear dial on its front panel to get you into the right frequency area, actual frequency readout takes place in the "104". The display automatically changes to the correct frequency as you go from transmit to receive.

Kit-built VFO. The "644" uses the same kit VFO as the new SB-104. And thanks to the true digital frequency readout in the "104", concern about dial VFO linearity problems is a thing of the past. If you work serious DX with your new SB-104, you'll want the new "644".

Kit SB-644, 10 lbs. **119.95**

SOLID-STATE ACCESSORIES

SB-614 Station Monitor

How clean is your signal? With the SB-614, you'll know. It monitors transmitted SSB, CW, and AM signals up to 1 kW from 80 – 6 meters. The highly visible 1½ x 2" CRT, with push-pull drive for a keystone-free, sharp, clean trace, indicates a wide variety of common operating problems. non-linearity, insufficient or excessive drive, poor carrier or sideband suppression, regeneration, parasitics and CW key clicks. The manual includes 40 CRT display illustrations and explanations.

Complete controls. All standard scope control functions are available in the "614"...Vertical Gain & Position, Horizontal Gain & Position, Focus, Mode (SSB, Trapezoid & Cross for RTTY Mark/Space adjustments). The improved recurrent, automatic sync-type sweep generator is adjustable in three ranges from 10 Hz to 10 kHz. Front panel control gives 11 steps of attenuation. For limited test applications the "614" can be used as a normal scope, and provides 10 Hz to 50 kHz bandwidth, good sync and high input sensitivity. A rear panel 10:1 vertical attenuator provides extra convenience.

Additional features include all solid-state design; rear panel Astigmatism control; standard horizontal and vertical inputs for use as a scope; exciter and linear inputs/outputs. Circuit board/wiring harness design makes assembly fast and easy. What kind of signal do you have? Order your new SB-614 today and know.

Kit SB-614, 17 lbs. **139.95**

SB-230 Conduction-Cooled 1 KW Linear

Strong and silent. The new "230" uses a husky Eimac 8873 triode in proven, stable, grounded grid circuitry to deliver up to 1200 watts PEP SSB, 1000 watts CW input from less than 100 watts drive. And the "230" is also rated at 400 watts input for slow-scan TV and RTTY. A massive heat sink eliminates the need for a fan.

Complete operating convenience. On the front panel of the new SB-series low profile cabinet you'll find all controls at your fingertips for easy operating. Bandswitching is done with a single knob...Load and Tune controls are clearly marked. Full metering facilities.

A full complement of built-in safety features. The cabinet features microswitch interlocks on both the top and bottom to shut down the primary power when the cabinet shells are removed. Front panel status lights indicate Hi Temp, Exciter and Delay. The heat sink for the 8873 is temperature monitored; if the temperature rises too high, a thermal circuit breaker opens, the linear shuts down and the Hi Temp light goes on. The Exciter light indicates that the linear is running straight through, without amplification. To allow the tube sufficient time to warm up, a delay circuit is built-in. When warm-up is completed, the Delay light goes out. The On-Off switch also includes a built-in circuit breaker for the primary side of the power transformer. And the cathode of the tube is fused for additional protection.

Easy assembly. The new SB-230 goes together in 15 to 20 hours. No alignment is necessary.

The new SB-230, styled to match the SB-104 transceiver, delivers all the features and performance you've come to expect from Heath. We think you will agree it's the greatest value in modern linears.

Kit SB-230, 40 lbs. **319.95**

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Benton Harbor, Michigan 49022

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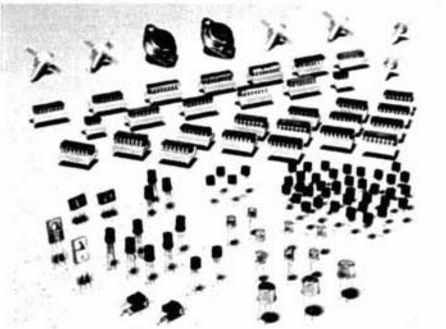
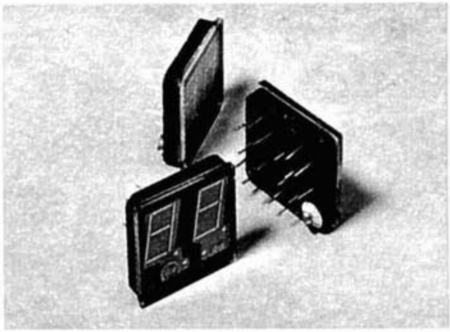
SS-103

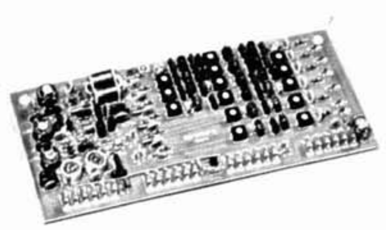
CHECK THESE ADVANCE DESIGN FEATURES:



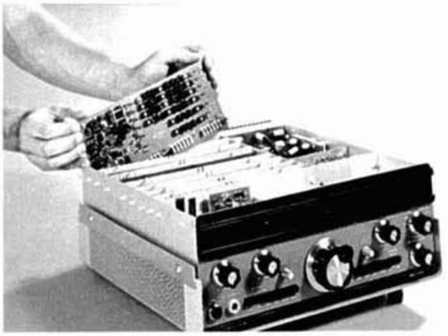
Completely solid-state design...including the finals. Over 275 solid-state devices, including 31 integrated circuits. ***The SB-104 output board and final transistors are warranted for one full year.***



True digital readout. Six ½″ gas-discharge displays deliver resolution down to 100 Hz with across-the-room visibility.

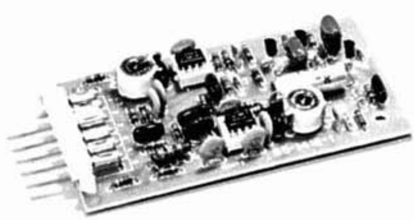


Completely broadbanded. Neither the transmitter nor receiver sections require tuning from 3 to 30 MHz...instant QSY from 80 to 10 meters is a reality.



Circuit board construction. Most components mount on 15 circuit boards for easy assembly. The seven major boards can be extended out of the chassis for adjustment or troubleshooting while rig is operating.

Complete back-panel connections: Phone patch in & out; auxiliary audio input; speaker; key; ALC; VFO in & out; driver out; IF out; accessory plug; power plug; two spare jacks; separate transmit & receive antenna jacks.



New noise blanker plugs into SB-104 & solves the ignition noise problem. Provides up to 50 dB of effective blanking. Rep rate 10 to 2000 pulses/sec.; pulse widths 1 to 250 µsec.