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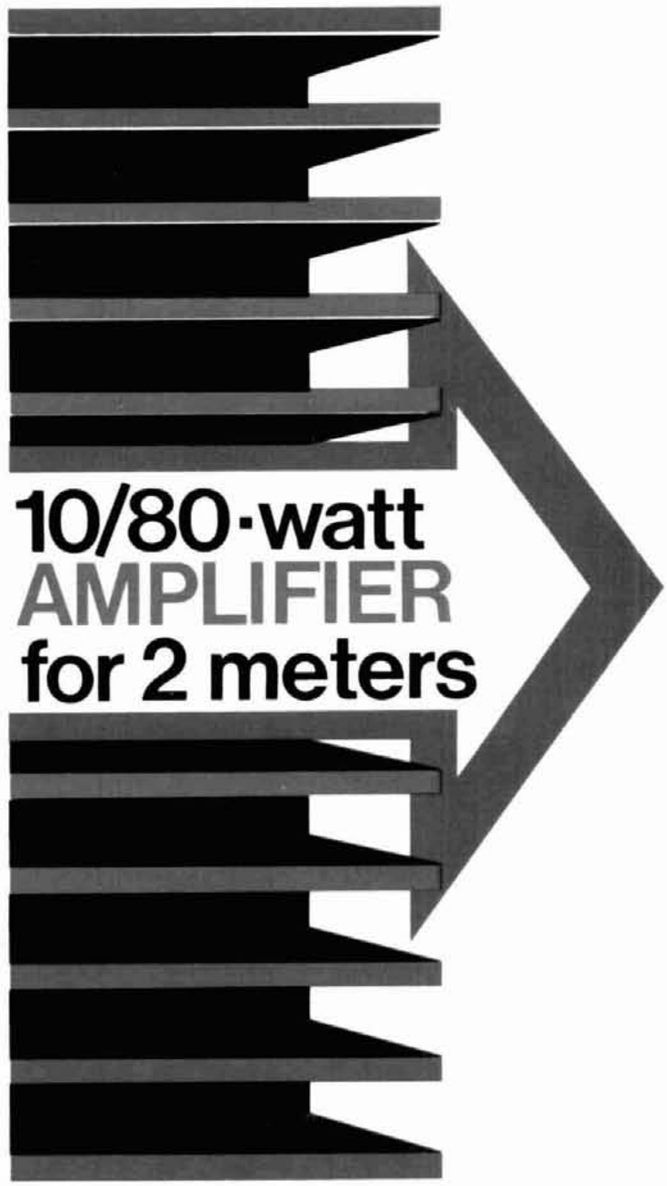
ham radio magazine

FEBRUARY 1979



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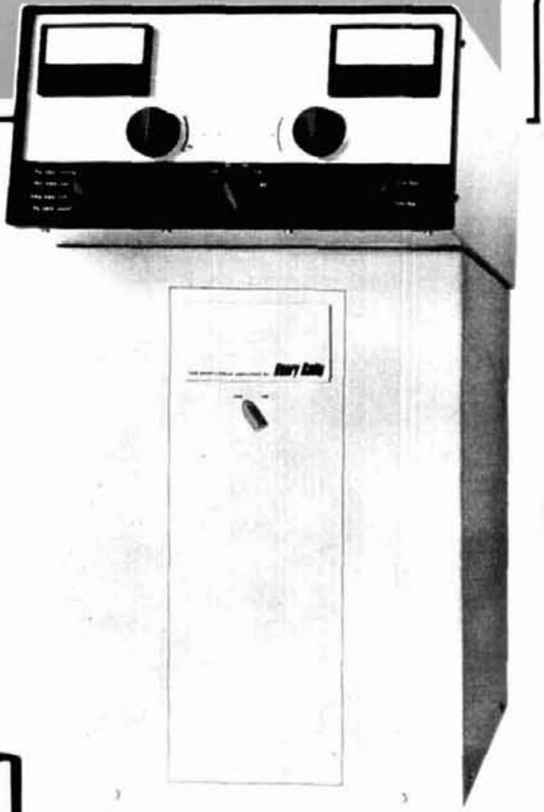

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Transmitter:	
Antenna Impedance:	50 ohm, unbalanced
Carrier Suppression:	Better than -45 dB
Side-Band Suppression:	Better than -55 dB at 1000 Hz
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A SWR and dual range wattmeter (300 and 30 watts full scale) lets you measure RF power output for simplified tuning.

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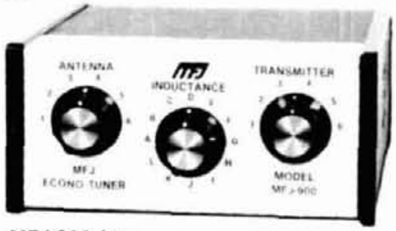

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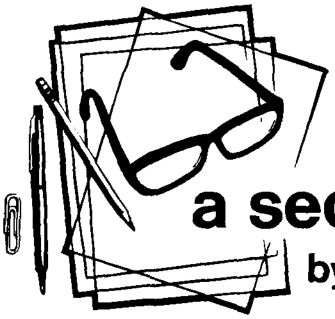

a second look

by Jim Fisk

This year Amateur Radio may be facing one of its biggest challenges in 20 years. What I'm referring to, of course, is the World Administrative Radio Conference which will convene this fall in Geneva, Switzerland. Better known in ham circles as WARC 79, this conference of ITU member nations will consider all the high-frequency allocations, including those of Amateur Radio, broadcasting, maritime mobile, and the other radio services which require operating frequencies.

There are some Amateurs who would like you to believe that the high-frequency Amateur bands will be completely decimated at this conference — ravaged by the greedy big-money interests who covet our bands for their own selfish purposes — but I think that the people who are all too eager to promote that turn of events are either alarmists or poorly informed; probably both. Obviously, it's impossible to forecast the outcome of WARC 79 at this point in time, but the Amateurs I've talked to who are officially involved with preparations for the conference are all cautiously optimistic that the high-frequency Amateur bands after WARC will be pretty much the same as they are now. And they are the ones who should know, not the purveyors of gloom and doom who apparently get their information from the Wizard of Oz — or some other equally unlikely source.

In the past, dozens of magazine articles have been written about the "terrible drubbing the Amateur Radio service has taken at every international frequency allocation conference." If you carefully review the record, however, you'll find that exactly the reverse is true; in every case American Amateurs have actually *gained* more high-frequency spectrum than they lost.

It is generally believed, for example, that at one time Amateurs had exclusive use of all wavelengths below 200 meters. That's a fable which has been quoted so often it's now accepted as fact. Actually, the 1912 regulation in question restricted *all* stations not involved in commercial traffic from going *above* 200 meters. That included *all* private, commercial, and experimental stations not transacting business or developing equipment for business purposes. Amateurs had no exclusive claim on "200 meters and down" — they shared that spectrum with virtually every other radio service. In fact, Amateur Radio stations at that time were required to specify their operating wavelengths, which were invariably 150, 175, or 200 meters — three spot frequencies below 200 meters.

In the early 1920s it became apparent that the 1912 law was hopelessly inadequate for the then existing conditions. More stations were on the air than ever before, the broadcast boom was well underway, and Amateurs had demonstrated the long-distance capabilities of the short waves. The scramble for short-wave territory was on, and every service was pushing for all the high-frequency spectrum it could get. To bring order to the ensuing chaos, a domestic radio conference was held in Washington in 1924; part of the outcome was the establishment of four harmonically-related Amateur bands: 160, 80, 40, and 20 meters. It's important to remember that this was not an international agreement, however, nor in fact did it have the authority of law — it was purely a mutual agreement between the various radio services in the United States.

The 1927 International Radio Conference in Washington saw precious kHz shaved off the American 160, 40, and 20 meter bands, but in return we received an exclusive new band at 10 meters. Amateurs in Europe fared less well, and some will argue that American Amateurs now had to share 40 and 80 meters with the foreign broadcasters, but that was true *before* the 1927 conference convened. Compared with the spot frequencies given to Amateurs in 1912, the new international allocations were a vast improvement.

There was no change in the Amateur bands at the Madrid conference in 1932, nor at Cairo in 1938. The next conference was scheduled for Rome in the spring of 1942, but because of the war, the next International Radio Conference was not held until 1947, in Atlantic City. Amateurs lost some space on 160, 20, and 10 meters at Atlantic City, but we received a nice bonus in return: a brand new band at 15 meters. Hence, there was not a net loss at all, but a gain! Those are the same bands we are still using today.

In reviewing the record of high-frequency Amateur allocations, we have progressed from what was essentially spot-frequency operation in 1912, to 3485 kHz of high-frequency operating space in 1927, to 3500 kHz today. Based upon past performance, and the proven service of Amateur Radio to the public, I believe we have every reason to be optimistic about the future.

Jim Fisk, W1HR
editor-in-chief

Imagine All The Places You Can Tuck ICOM's Remotable IC-280. (Think small.)

The **IC-280** 2 meter mobile comes as one radio to be mounted in the normal manner: but, as an option, the diminutive front one third of the radio detaches and mounts by its optional bracket, while the main body tucks neatly away out of sight. Now you can mount your 2 meter radio in pint-sized places that *seemed far too cramped* before.

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All ICOM radios significantly exceed FCC regulations limiting spurious emissions.

Specifications subject to change without notice.

IC-280 Specifications: ☐ Frequency Coverage: 143.90—148.11 MHz ☐ Operating Conditions: Temperature: −10°C to 60°C (14°F to 140°F), Duty Factor: continuous ☐ Frequency Stability: ±1.5 KHz ☐ Modulation Type: FM (F3) ☐ Antenna Impedance: 50 ohms unbalanced ☐ Power Requirement: DC 13.8V ±15% (negative ground) ☐ Current Drain: Transmitting: 2.5A Hi (10W), 1.2A Lo (1W), Receiving: 0.630A at max audio output, 0.450 at SQL ON with no signal ☐ Size: 58mm(h) x 156mm(w) x 228mm(d) ☐ Weight: approx. 2.2 Kg ☐ Power Output: 10W Hi, 1W Lo ☐ Modulation System: Phase ☐ Max. Frequency Deviation: ±5 KHz ☐ Spurious Output: more than 60 dB below carrier ☐ Microphone Impedance: 600 ohms dynamic or electret condenser type, such as the SM-2 ☐ Receiving System: Double superheterdyne ☐ Intermediate Frequency: 1st: 10.695 MHz, 2nd: 455 KHz ☐ Sensitivity: 1 uv at S+N/N at 30 dB or better, Noise suppression sensitivity 20 dB, 0.6 uv or less ☐ Selectivity: less than ±7.5 KHz at −6 dB, less than ±15 KHz at −60 dB ☐ Audio Output: More than 1.5W ☐ Audio Output Impedance: 8 ohms

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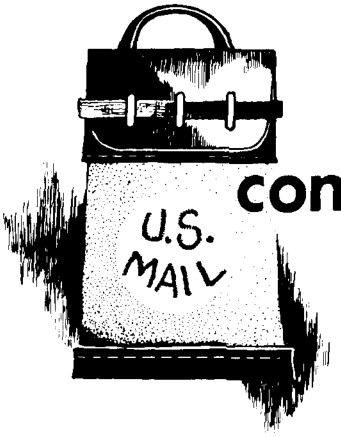

comments

zip-cord feedlines

Dear HR:

The article on "Zip-Cord Feedlines" published in the April 1978 issue prompts me to record, somewhat belatedly, my own experience on the subject. I have, for many years, happily used zip-cord feedlines particularly for temporary or semi-temporary antennas for use up to 21 MHz. I have always used the type with clear insulation, thinking that there was less chance of fillers absorbing the rf energy. I have to admit, though, that I have no concrete reason for this feeling, and I could be quite wrong. I have always used a simple antenna tuner and balun to couple the transmitter to the line, and, without exception, I have found the results to be gratifying. For example, I have quite recently used a 20-meter dipole slung between a tree and the chimney of my apartment building, about 20 feet high, fed by about 30 feet of Radio Shack loudspeaker wire (the heavy-duty type with stranded conductors) and driven by a Heath HW101. I have always been able to raise interesting stations, and have had a good number of enjoyable ragchews with European operators on single sideband.

The characteristic impedance of zip-cord is, of course, unknown. In my particular example I measured and, from standard formulae, estimated its impedance to be around 100 ohms. I think that this impedance presents as good a match to any "real" antenna at modest height with nearby structures as does any well-characterized coaxial cable. The unknown impedance presents a problem, however, when attempting to measure the line loss, for the line must (usually) be terminated. To get some idea of the loss on the 30-foot length, I tried the following experiments:

First, I terminated the line with a 110-ohm carbon resistor combination. I fed the line through my antenna tuner, and, in the coax feed between the transmitter and the tuner, I inserted my SWR bridge. I then loaded the transmitter on 14 MHz and adjusted the antenna tuner for minimum SWR on the coax; I was able to get it down to an indicated value of 1.1:1. Then I removed the terminating resistor and measured the SWR both with the far end of the line open and shorted. I tweaked the tuning for minimum SWR before taking the reading. The reverse power was so high in each case that it was not possible to get reliable readings. I estimated the SWR to be at least 10:1 (neglecting possible error in the bridge — the forward and reverse powers are about the same with an open or short circuit on the output of the device). This indicates that the total loss, one-way, in the tuner and feeder is on the order of 0.9 dB, and this is a worst-case figure. This is not comparable with RG-8/U coax, but for a simple system it is a figure that I can certainly live with.

Tony Garratt-Reed, ex-G3VBZ/W1
Malden, Massachusetts

manned free-aircraft

Dear HR:

The Presstop of the October, 1978 issue is in error; the Double Eagle II was not the first free-aircraft to cross the Atlantic — it was the first *manned* free-aircraft to do so. The first free-aircraft to make the crossing was a high-altitude research balloon flown from Trapani, Sicily, in July, 1975, and brought down near Lexington, Kentucky after about 84 hours. Another balloon was flown from Sicily to Massachusetts in July, 1976. These balloons flew at an altitude of approximately 125,000 feet and had about 20 million cubic feet of capacity; they were launched and flown by National Scientific Balloon Facility personnel from Palestine, Texas for scientists from Italy, England, and Germany.

Spencer Petri, WA5JCI
Palestine, Texas

keyer memory

Dear HR:

I read with interest the Ham Notebook correspondence from K9WGN in the August 1978 issue of *ham radio* concerning my programmable keyer memory accessory. While it may be true that his unit programs the memory chips properly by simply pulling the RW pin to +5 volts, this is not good design practice. If you examine the data sheets for the 1101 (or any other static MOS RAM chip, for that matter) you will see that the RW pin should be pulsed only after the address is stable. The manufacturer's information for the 1101, in fact, recommends both the data and address inputs remain stable for at least 100 ns after the falling edge of the WRITE pulse. If the RW pin is held near +5 volts during programming, there is a chance some undesired bits may be altered while the input memory address is changing. For these reasons I feel the 74121 monostable, U8, is justified.

K9WGN is correct in stating that you can substitute the 7493 for the 74193s I used in that design. However, since the 7493 is a negative edge triggered ripple counter, the inverter, U11B, should be eliminated.

Andrew B. White, K9CW
Urbana, Illinois

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(And 25 watts of power)

Kenwood's exciting 25-watt TR-7625 and optional RM-76 Microprocessor Control Unit...gives you the power you want plus outstanding channel memory and scanning capabilities.

...there's more! The TR-7625 has a lot of versatility for a 2-meter FM transceiver, and when teamed up with the RM-76 gives you a lot of rig. Check these features for both:

TR-7625 (only)

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- Select repeater mode (simplex, transmit frequency offset (± 600 kHz or ± 1 MHz), or one memory transmit frequency).
- Scan for busy or open channel.
- Display indicates frequency (even while scanning) and functions (such as auto-scan, lower scan frequency limit, upper scan limit, and error, i.e. transmitting out of band).


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AMATEUR RADIO FARED QUITE WELL in the FCC's WARC 79 proposal, as further details were made available in mid-December. The biggest gain was the addition of three new HF Amateur bands — 10.1-10.2, 18.068-18.168, and 25.11-25.21 MHz — as exclusively Amateur bands worldwide. A 50-kHz addition to 15 meters, making it 20.95-21.45, is also in the proposed U.S. position, along with a shift 50 kHz downward for 40 meters to make it an exclusively Amateur band worldwide from 6.95-7.25 MHz.

80 Meters Would Become exclusively Amateur from 3.5-3.9 MHz in Region 2, where it's now shared with fixed and mobile services, while 3.9-4.0 MHz would remain shared as at present. The 10-meter band would remain as is, worldwide.

The Proposed Bottom-60-kHz loss on 160 to AM broadcast would be somewhat tempered by a proposed worldwide allocation of 1860-1900 kHz exclusively for Amateurs with 1900-2000 kHz remaining shared with other services in Regions 2 and 3.

With Two Notable Exceptions, the higher frequency Amateur bands also fared well. The big exception was 220-225 MHz, proposed for a worldwide Maritime Mobile band with Amateurs a secondary user in Region 2 only. Radiolocation use of the band would be continued as is through 1990 and beyond. The other significant loss is the lower part of the 1215-1300 MHz band, with Amateurs bumped from the lower 25 MHz in favor of navigational satellites. The 1240-1300 MHz portion would remain as is, shared with Radiolocation. No changes appear to have been proposed for 50, 144, 420, or the other higher frequency Amateur allocations.

CONGRESSMAN MARTIN RUSSO was very pro-Amateur Radio when he was guest of honor at the Chicago FM Club's mid-December meeting. Rep. Russo affirmed his and Rep. Van Deerlin's belief in the importance of Amateur Radio to the United States, and said that if spectrum user fees (as proposed in Van Deerlin's rewrite of the Communications Act) are adopted Amateurs should be exempted. He also feels that the U.S. preparation for WARC is very good, and much of what we proposed will be adopted next summer.

Questioned About Charges that allowing FCC staffers to be Amateurs constitutes "conflict of interest," Russo responded that he felt quite the opposite. It was the "mind-expanding" hobby of Amateur Radio that provided many FCC staff members with the background and knowledge to do their jobs, he said, so that rather than being a detriment an Amateur license was a plus for a member of the FCC. Very encouraging words from an important Illinois Congressman.

AMATEURS HOLDING MORE than one callsign who wish to retain the call assigned to their secondary station when renewing must be sure to renew well before expiration. One West Coast contester lost his prized 1x2 recently when his renewal application, complete with request to reassign his 1x2 to his primary (surviving) station license, arrived at Gettysburg a few days after the license expiration date. As his no longer renewable secondary license had expired, its 1x2 callsign was no longer his and thus could not, under the rules, be assigned to him.

A SERIOUS DROP IN BEGINNER interest in Amateur Radio has been reported in a number of parts of the country. Dealers in some areas say sales of code practice equipment and entry-level study materials have dropped quite significantly though this is certainly not universal and demand for higher grade study guides by already licensed Amateurs has seen little change.

The Drastic Drop In CB interest is the reason most cited for the downturn. When CB was hot many potential Amateurs discovered two-way radio as a hobby; moving into Amateur Radio when CB did not meet their expectations. At the same time the influx of new 27-MHz operators drove many previously satisfied CB hobbyists to Amateur Radio, simply to find room to operate. With those pressures gone, most of today's CBers seem content where they are.

WWV'S 20-MHZ TRANSMITTER is back on the air after an absence of several years. Improving propagation is cited as the reason for its return to the air, with promises to remain on 20 MHz "as long as propagation conditions warrant."

UNUSUAL REPEATER INTERFERENCE in the Wyandotte, Michigan, 147.84-24 system was traced to a super-regenerative receiver in a nearby garage door opener. Quench frequency of the receiver detector was 600 kHz, and any strong nearby signal would cause the receiver to transmit signals of ±600 kHz from that signal. They could be heard for several blocks. The door-opener manufacturer has been very cooperative, working with area Amateurs to determine what's causing the problem.

HENRY KANE, WA2KBI, OF CLIFTON, N.J. WON FIRST PRIZE in *Ham Radio Horizons'* just-ended sweepstakes, taking home a complete station that includes a TR-7, Tri-Ex Tower, CushCraft beam, Hy-Gain vertical, and a Wilson 2-meter hand-held.

The 200 Second-Place winners each won their choice of an MFJ accessory, while 250 copies of *Radio Angels* went to the third-place winners.

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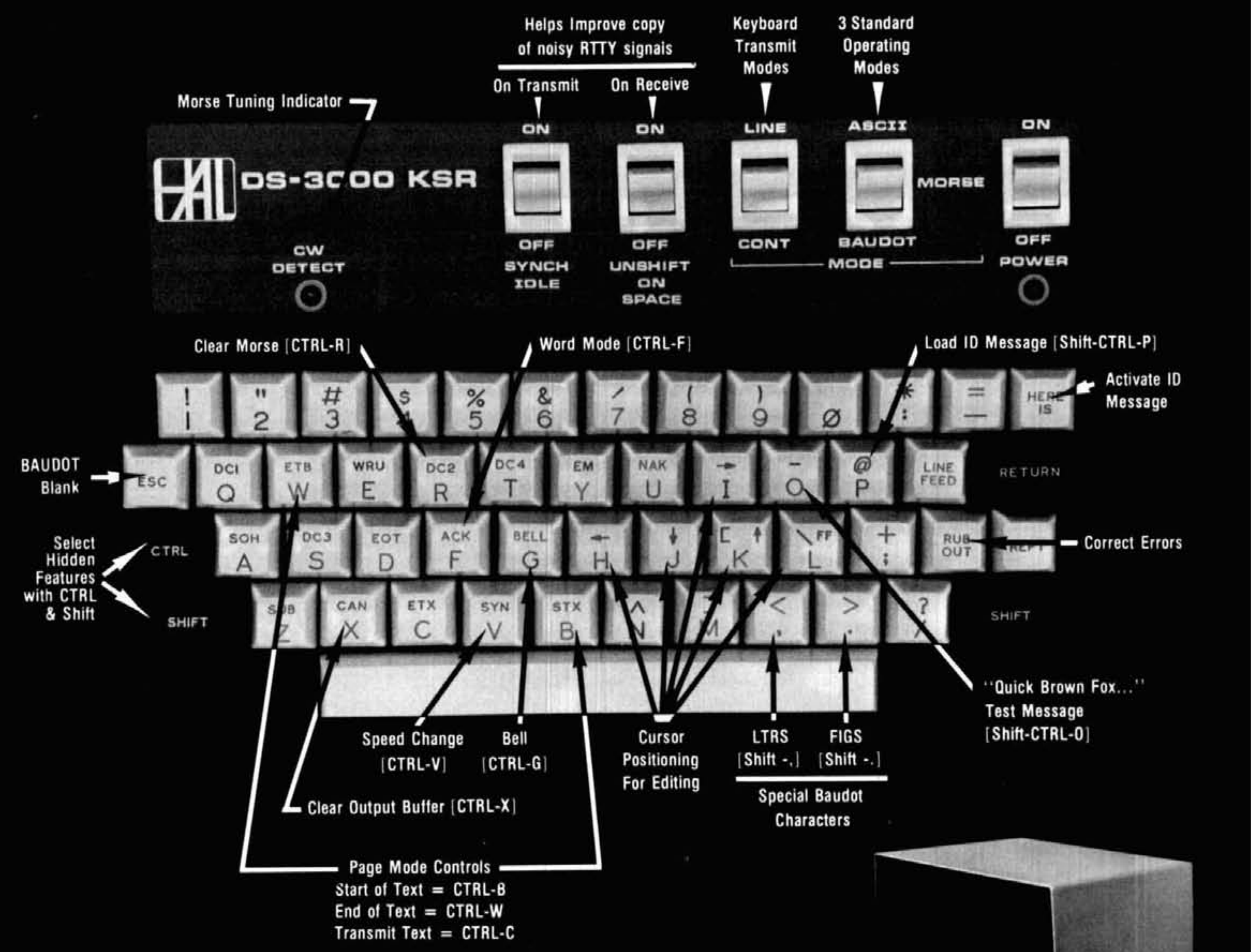

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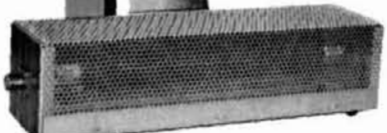

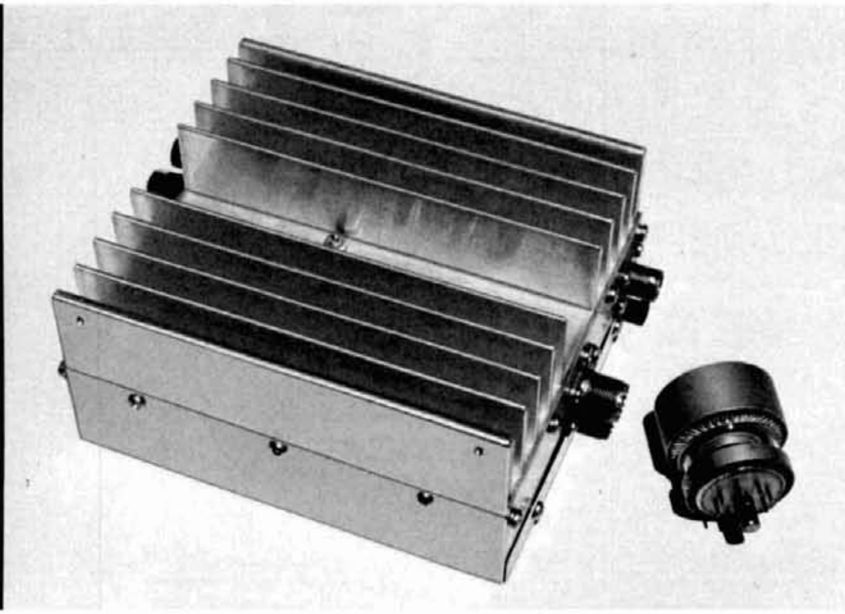

10/80-watt amplifier

for 2 meters

Design of an rf-activated, power-selectable, 2-meter amplifier intended for use with a 2-watt fm handie-talkie

It all began rather innocently. My good friend Jim Fisk, WB6YED (no relation to *ham radio*'s editor), won a pair of Motorola HEP S3041 40-watt vhf power transistors at a prize drawing during the 1975 Dayton Hamvention. Several of us who were with him envied his good fortune, and we wondered what he would do with his new treasures. Months later, during a visit to Jim's home, I got my answer.

Jim had a 2-watt, hand-held fm transceiver he wanted to use in his car with a separate microphone. He envisioned a package that would produce either 10 or 80 watts output for 2 watts of drive, depending on his distance from the station he wanted to talk to. In addition, T/R switching had to be accomplished without a separate control line between the transceiver and the power amplifier. As Jim described what he was thinking of, I had the feeling he was going to ask me to tackle the job, which he did. Since he had been more than generous with his time and resources when I needed them, I was glad to help.

Putting together his requirements, I came up with the block diagram shown in **fig. 1**. The power amplifier had to have enough gain to produce roughly 80 watts output from 2-watts input; this amounts to 16 dB gain. A little more gain wouldn't hurt, because there are sure to be losses between the input to the entire circuit and the input to the power amp. The attenuator is switched into the circuit when low-power operation is desired. As it turned out, the attenuator wasn't that simple.

The requirement that transmit-receive switching

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be done without control lines necessitated the use of some type of rf-actuated T/R switch to put the power amp in the circuit during transmit operation. During receive the entire amplifier is simply bypassed with a coaxial line.

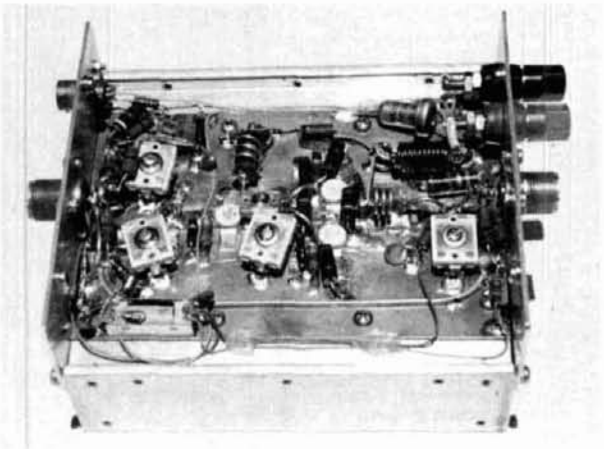
circuit description

I started the project by building the basic power amplifier. The HEP S3041 is equivalent to a 2N6084. A pair of 2N6084s at 150 MHz can be driven to 80 watts output with 25 watts input. This amounts to 5 dB gain; a driver capable of at least 11 dB gain is necessary to ensure a full 80-watt output from 2 watts of input. Since it suited my needs so well, I used the two-stage amplifier circuit and layout described in the Motorola Application Note AN-585, with as few differences as possible.[1] A 2N6083 was used for the driver in the amplifier described in the application note. Lacking a 2N6083 I used a 2N6136, a 25-watt uhf device useful to over 500 MHz and readily available to me.

The initial attempt at the amplifier was a copy of the layout described in the application note (see **fig. 2**). It was unstable for all but low drive levels. Any output greater than a few watts was accompanied by a healthy spur approximately 5 MHz from the desired frequency and several other spurs of lesser amplitude, including one at 5 MHz. I had used a uhf device for the driver, and so this was not entirely unexpected.

The choke/bead combinations in the transistor base circuits are used to suppress spurious oscillations as well as to provide dc return paths for the bases. Ferrite beads are low-Q inductors at vhf and, as such, make excellent broadband chokes. At lower frequencies, such as the high-frequency bands, the resistance of the beads decreases while the inductance increases. The result is an increased Q. Uhf transistors exhibit tremendous gain in the high frequency region, and high Q circuits are an invitation for them to take off. The solution consists of adding 10-ohm resistors in shunt with the choke/bead combinations to guarantee a low Q at lower frequencies.

Amplifier with the dust cover removed. The rf input is on the left side, the 10/80-watt control connector is in the upper left, and the 12-volt connections are in the upper right. The T/R circuitry is in the lower left portion of the chassis.



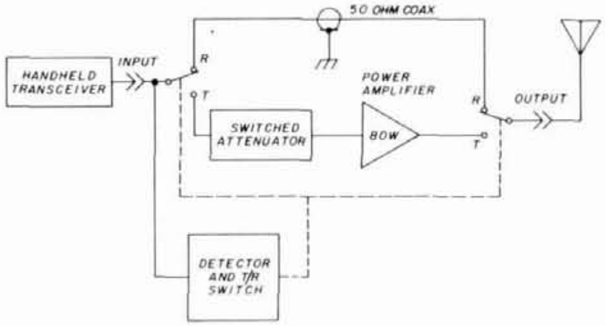


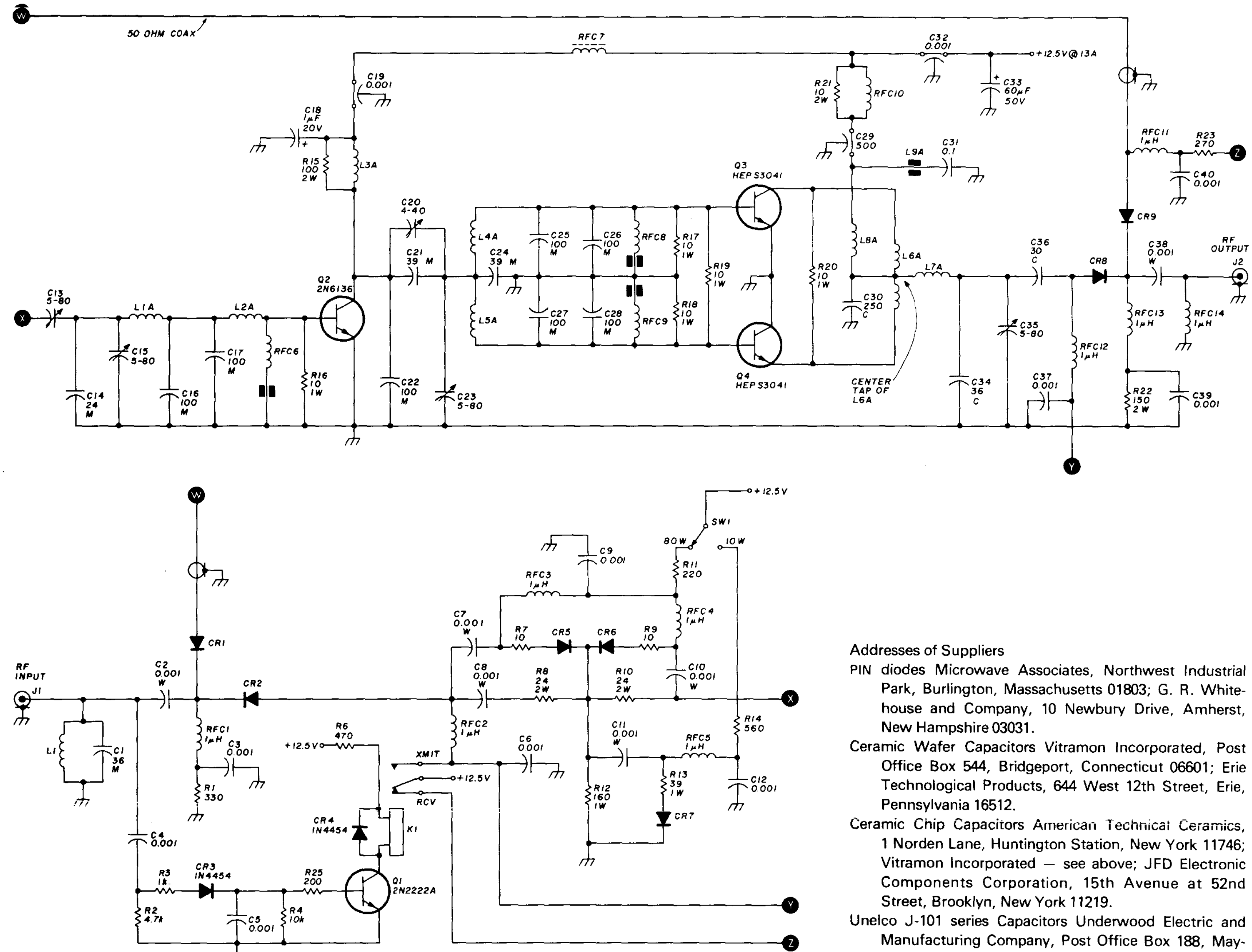
fig. 1. Block diagram of the 10/80-watt amplifier for 2 meters. The switched attenuator is used to vary the drive level into the amplifier to switch between 10- and 80-watt power levels.

Another addition to the circuit in the Motorola Application Note is the parallel tank circuit (L1 and C1) at the input to the amplifier. It helps reduce any second- and third-harmonic energy coming from the hand-held transceiver.

After tuning up the completed amplifier, I had 80 watts output for 1 watt of drive. This was more gain than necessary, since 2 watts were available. Rather than detune the amplifier, I decided to put some attenuation in the circuit for high-level operation. Using an input of 2 watts, 2.6 dB of attenuation results in an output of approximately 80 watts, while 10 dB of attenuation provides an output of approximately 10 watts.

The switched attenuator shown in **fig. 1** is drawn schematically in **fig. 3(A)**. It is a tee configuration in which PIN diodes are used to add or remove resistors that are in parallel with the elements of the attenuator. PIN stands for "P-intrinsic-N" and describes regions of P and N semiconductor material that have a piece of undoped or "intrinsic" silicon sandwiched between them. This construction gives the PIN diode its unique properties. For low-frequency signals, the device behaves like a conventional PN-junction diode. At uhf, the diode is a current-controlled resistor. For high-forward current, the series resistance is low. For low-forward current, the series resistance is high. Reverse biasing the PIN diode further raises its series resistance. An ideal PIN diode should not rectify rf, but in the real world these diodes do rectify and this must be dealt with in rf-switching applications.

For 80 watts of output, the series PIN diodes, CR5



Addresses of Suppliers

PIN diodes Microwave Associates, Northwest Industrial Park, Burlington, Massachusetts 01803; G. R. Whitehouse and Company, 10 Newbury Drive, Amherst, New Hampshire 03031.

Ceramic Wafer Capacitors Vitramon Incorporated, Post Office Box 544, Bridgeport, Connecticut 06601; Erie Technological Products, 644 West 12th Street, Erie, Pennsylvania 16512.

Ceramic Chip Capacitors American Technical Ceramics, 1 Norden Lane, Huntington Station, New York 11746; Vitramon Incorporated — see above; JFD Electronic Components Corporation, 15th Avenue at 52nd Street, Brooklyn, New York 11219.

Unelco J-101 series Capacitors Underwood Electric and Manufacturing Company, Post Office Box 188, Maywood, Illinois 60153.

fig. 2. Complete schematic diagram of the 10/80-watt amplifier including the basic power amplifier (with modifications) described in *Motorola Application Note AN-585*. Inductors with the suffix A are described in the Application Note. L1 is four turns of no. 36 (0.13-mm) AWG enameled wire wound on a 100k-ohm, 1/4-watt carbon resistor. CR1, CR2, CR5, CR6, CR7, and CR9 are Microwave Associates MA47047 PIN diodes, while CR8 is an MA47080. All resistors are 1/2-watt, 5 per cent tolerance carbon composition resistors, unless otherwise specified. For the capacitors M = mica, W = ceramic wafer, and C = ceramic chip. RFC6, RFC8, and RFC9 are 0.15-μH molded rf chokes with a Ferroxcube 56 590 65/3B bead on the ground lead. RFC7 is a Ferroxcube VK200 19/4B wideband ferrite choke, and RFC10 is ten turns of no. 14 (1.6-mm) AWG enameled wire wound on R21. K1 is a 6-Vdc, 500-ohm coil dry reed-type relay.

and CR6, are forward biased, and shunt PIN diode, CR7, is left off. For 10 watts of output the series diodes are left off while the shunt diode is forward biased. The calculated equivalents of the two conditions are shown in **figs. 3(B)** and **3(C)**. The resistor values for the attenuator were chosen so the impedance looking in either direction would be close to 50 ohms. *It is important that a good grade of carbon-composition resistor be used for the elements in the attenuator that carry rf current, and also that leads are kept as short as possible.* Because of stray elements, this type of attenuator becomes more difficult to construct as frequency is increased. Two meters is probably getting near the practical limit.

The last block in **fig. 1** is the rf detector and T/R switch. The detector, CR3, is simply a rectifier with a low enough reverse recovery time to be effective at 150 MHz. The rectified rf signal is used to forward bias the relay driver, Q1, which in turn energizes K1 during transmit operation. The relay closes when approximately 0.7 watt of power is fed to the input jack of the amplifier. The relay contacts are used to switch the bias to the PIN diodes in the actual T/R switch. These PINs are CR1 and CR2 at the amplifier input, and CR8 and CR9 at the output.

For simplicity, a relay was used instead of solid-state bias switching. W9KHC pointed out in his article that the dc reverse bias on a PIN diode should be equal to the peak rf voltage across the PIN in question.[2] For CR9, which is reverse biased during transmit operation, this would amount to 88 volts, assuming 80 watts is being delivered to a 50-ohm load. A higher load impedance would require a higher bias voltage. If a PIN diode rectifies when the reverse bias is not great enough, it will conduct and be destroyed due to excessive power dissipation. However, if the PIN diode bias line is just left open, any rectified current will tend to reverse bias the PIN and protect it. Open is the key word here. A set of open relay contacts is much better than a reverse biased switching transistor, which has some measurable leakage. Since the relay was employed, it eliminated the need for a high-voltage supply which would require an inverter for 12-volt mobile operation. With the arrangement shown, T/R switching is accomplished using the same dc supply that powers the amplifier.

component selection

Since the construction of this power amplifier involves relatively high power levels and frequencies, a discussion about the components used is in order. The PIN diodes used in the T/R switch and attenuator are general-purpose diodes for uhf work. They should have a low series resistance when forward biased, since they contribute directly to insertion loss. The MA-47047 typically has 1.5 ohms of resistance at 30 mA of forward bias, while the MA-47080 has a resistance of 0.45 ohms at 100 mA. In addition, the minority carrier lifetime should be roughly 10 times longer than the period of the lowest frequency in use. This prevents rectification if leakage occurs. The low-power PIN diodes used in the construction

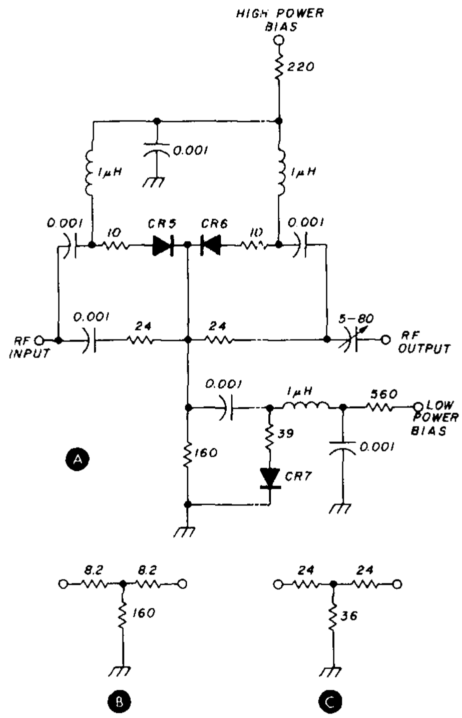
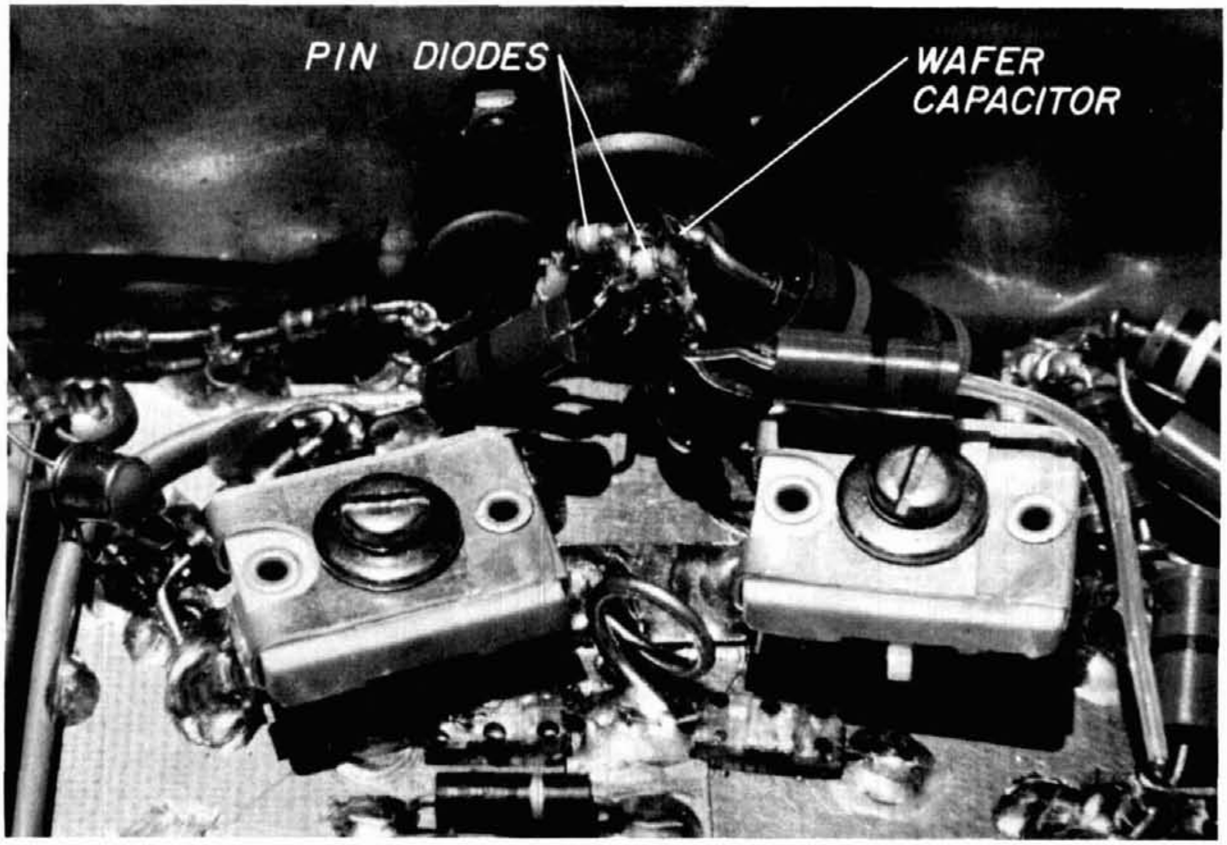


fig. 3. Schematic diagram of the switchable tee attenuator is shown at (A), while (B) and (C) respectively, show the effective attenuators for high power (2.6 dB) and low power (10 dB).

of this project were in a pellet-style case. The MA-47047 comes in a lead-style case similar to that of a signal diode. Either type will work — just keep lead lengths short. The MA-47110 could be substituted for the MA-47047.

PIN diode CR8 must have a high current rating because it carries the output power of the amplifier. It will get fairly warm and should be heat sinked directly to the output connector via C38. Transferring heat through a capacitor may seem unconventional, but it is accomplished with a wafer-style capacitor. Wafer capacitors are a truly leadless capacitor with a ceramic dielectric. They are especially well suited for bypass and coupling applications where tolerance is not critical. The 1000-pF units used in this project are little squares about 4.1 mm (0.16 inch) on a side and 0.76 mm (0.03 inch) thick. The two large surfaces are metallized with a silver compound so they can be sol-



Detail of the circuitry near the rf input connector showing PIN diodes and ceramic wafer capacitors attached to the connector.

dered. It is a good idea to handle the capacitor with fine tweezers so that oil from your hands doesn't make soldering difficult. For best results, the wafers should be soldered using a silver-bearing solder and a very small iron tip. Ordinary tin-lead solder is likely to leach away the silver and make soldering impossible. Alpha Metals is one company that produces a 62 per cent tin, 36 per cent lead, and 2 per cent silver solder.

Another uncommon capacitor used in the amplifier is the ceramic chip. Chip capacitors are made from a material similar to the wafer capacitor dielectric. They are small cubes approximately 2.5 mm (0.10 inch) on a side with a pair of opposite sides metallized to accept solder. These parts are available in tolerances as fine as ± 1 per cent. In addition, they are also leadless and have very low loss. Due to their high Q (low loss), they are capable of carrying several amps of rf current where other types of capacitors would quickly go up in smoke. Since all of this comes at a price, chips were used only at the higher power levels in the amplifier.

A less expensive part is the capacitor made by Unelco which Motorola used in the application note. This is a silvered-mica type and is suitable for use at high power levels. It is physically quite large compared with the ceramic chip, but that is not a drawback for most applications. To reduce the effect of lead inductance of higher value capacitors (which have low reactance), it is often possible to parallel two capacitors, as in the case of C25 and C26.

The coaxial cable used to bypass the amplifier on receive is a solid-jacket type used because it was available. Any good 50-ohm cable can be substituted.

protection circuitry

Many high-power, solid-state rf amplifiers have protection circuitry to reduce rf drive, or completely shut down the amplifier, when a high vswr is sensed. Since this amplifier was designed for a specific installation, the extra circuitry was not included. A load vswr of 2:1 or less should be adequate for the design.

Another circuit commonly associated with high-power amplifers is a thermal protection circuit. These circuits sense high temperature, usually near the output stage devices, and act to reduce drive or shut the amplifier off. Since a substantial heatsink is used and two-way operation is always intermittent, this type of circuit is not justified. If the amplifier is to be operated for more than short periods of time and in a place where air circulation is restricted, some thought should be given to a thermal protection scheme.

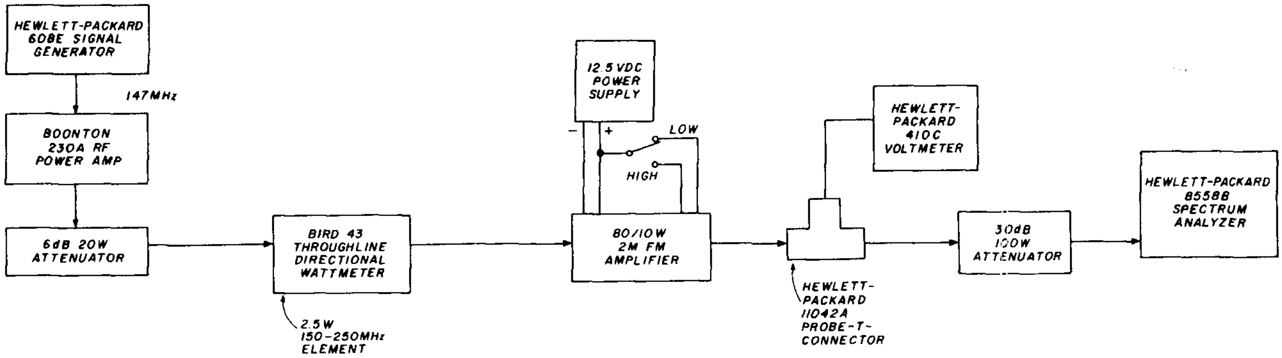


fig. 4. Amplifier test set-up.

Additional comments on circuit protection are given in **ref. 3**.

construction

No attempt was made to miniaturize the amplifier. With the cover in place, the assembly measures 7.6 x 14.0 x 15.9 cm (3 x 5.5 x 6.3 inches). The amplifier was built in "bread-board" fashion on a 10.2 x 15.2 cm (4 x 6 inch) piece of 1.6 mm (0.062 inch) thick epoxy-fiberglass, copper-clad board. Connections, which are isolated from ground, are made on miniature standoff insulators. This construction lends itself well to building power amplifiers, as it nearly eliminates rf ground problems. The heatsink may be a little larger than necessary, at 3.8 x 14.0 x 15.2 cm (1.5 x 5.5 x 6.0 inches), but since size was not important, the extra area is cheap protection. Long pieces of wire and ferrite beads are held in place with RTV-type adhesive.

Parts placement follows the general layout of the amplifier pictured in Motorola Application Note AN-585.

The amplifier is easily installed in the trunk of a car and controlled from the driver's seat. A pair of wires and a switch are used to turn the power on. Use at least no. 14 (1.6-mm) AWG wire for the dc connection to reduce the voltage drop from battery to amplifier. The output power level is controlled by three wires and a single-pole, double throw switch. One of the three wires can be the + 12 volt line that supplies the amplifier. Adding a coaxial cable to carry rf from the hand-held transceiver to the amplifier completes the installation.

Placing the cover on the amplifier slightly detunes it. This is compensated for by slightly mistuning the trimmer capacitors by trial and error until the performance with the cover on is optimized.

purity of emissions

Since the completion of this project, the FCC has issued new requirements on transmitter spurious radiation. For transmitters and amplifiers operating between 30 and 235 MHz and having more than 25 watts of output power, all spurious emissions, including harmonics, must be at least 60 dB below the mean carrier level. The second harmonic of the carrier is the most important spurious output to be dealt with in this amplifier. Since at 80 watts output the second harmonic is already 50 dB below the fundamental, a simple pi or tee network after the amplifier will provide enough attenuation to meet the 60 dB requirement. A summary of the performance data from the amplifier test circuit (see **fig. 4**) is given in **table 1**.

table 1. Performance data for the 80/10-watt amplifier at 147 Mhz with the cover installed.

receive insertion loss — 1.1 dB
drive level to accomplish T/R switching — 0.7 watt

	low power	high power
drive power	2 watts	2 watts
output power	7.2 watts into 50 ohms	74 watts into 50 ohms
current drain	4.0 amps	12.0 amps
2nd harmonic	40 dB	50 dB
3rd harmonic	44 dB	47 dB
input swr	1.6:1	1.4:1

acknowledgments

I would like to thank Bob Yankowiak for his patient technical assistance during the construction of the amplifier and the writing of this article. I would also like to thank George Johnson, K9ODF, for converting my color negatives to black and white prints for use in this article, and my wife Karen for typing the manuscript.

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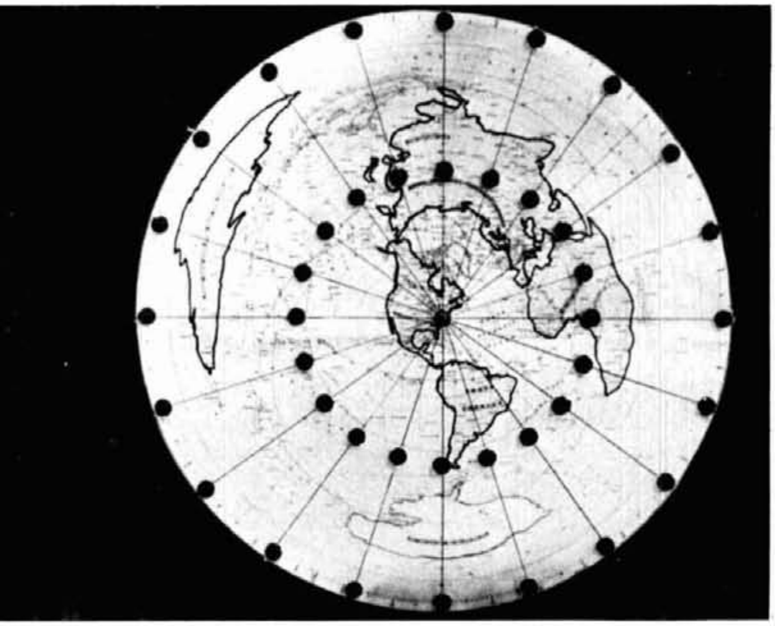
1. "VHF Power Amplifiers Using Parallel Output Transistors," from *Motorola Application Note AN-585*, Motorola Semiconductor Products, Phoenix, Arizona, 1973.
2. James K. Boomer, W9KHC, "PIN Diode Transmit/Receive Switch for 80-10 Meters," *ham radio*, May 1976, page 10.
3. "Design Techniques for an 80-Watt, 175-MHz Transmitter for 12.5-Volt Operation," from *Motorola Application Note AN-577*, Motorola Semiconductor Products, Phoenix, Arizona, 1972.

ham radio



solid-state antenna position display

Another approach for converting an antenna rotator to digitized readout, using discrete LEDs to show bearing segments

The need for an improved indication of antenna heading is most apparent during contesting and DX chasing. Many operators have considerable difficulty relating the compass heading of the beam, as indicated on the typical antenna-rotator control box, to the location of a particular country on the globe.

Normally, a table, slide rule, or a chart is used to find the correlation between the prefix of a call and the beam heading. People have a general sense of where a country or area is, but they are less than adept at translating this sense into an angular heading.

I once saw a global display scheme using a balanced pointer driven by a pair of selsyns, one coupled to the mast on the tower and the other driving the pointer on a wall-mounted map. My goal was then established, to construct an electronic display using the existing analog voltage at the rotor control box.

A completely solid-state design using available low-cost components was a must, and the display had to be suitable for mounting at my operating desk. I considered several options for the construction and presentation of the display. Personal taste and the builder's skill are involved in making the choice; a simple and effective version is described which is suitable for any home craftsman with moderate skills.

principles of operation

The Ham II provides a 13-volt signal swing which drives a 1-mA meter, calibrated in degrees, to indicate antenna position. This voltage provides an ideal analog signal source for digitizing with a high-impedance CMOS A/D converter.

The Motorola MC14433 DVM chip was judged as most suitable for this display. It will accept a 0 to ±1.999 volt analog input voltage swing, and provides a binary-coded-decimal (BCD)/TTL compatible output. In addition, it has a self-contained clock and provides timing pulses to drive the TTL control logic and decoder drivers that make up the remainder of the display circuitry. Low-cost BCD-to-decimal drivers were chosen because of the display format.

The display background itself is a polar great-circle map of the globe divided into twenty 18-degree sectors. With a 0 to 1.99 volt input signal swing and a maximum 360-degree rotation, each sector corresponds to an increment of 100 mV. Each hemisphere corresponds to an increment of 1 volt of analog input signal. For example, as seen in **fig. 1**, sector 1 is displayed when the input potential to the MC14433 has any value from 0.0 to 0.1 volt. This corresponds on the map to the sector between 180 and 162 degrees in the S-SE quadrant of the Ham II meter scale. Likewise, sector II would be displayed when the input voltage is between 1.0 and 1.1 volts,

By W. K. Springfield, AE4A, 2607 Deerdell Lane, Reston, Virginia 22091

which corresponds on the map to the sector between 360 and 342 degrees in the N-NW quadrant of the scale. My rationale for choosing twenty 18-degree sectors was

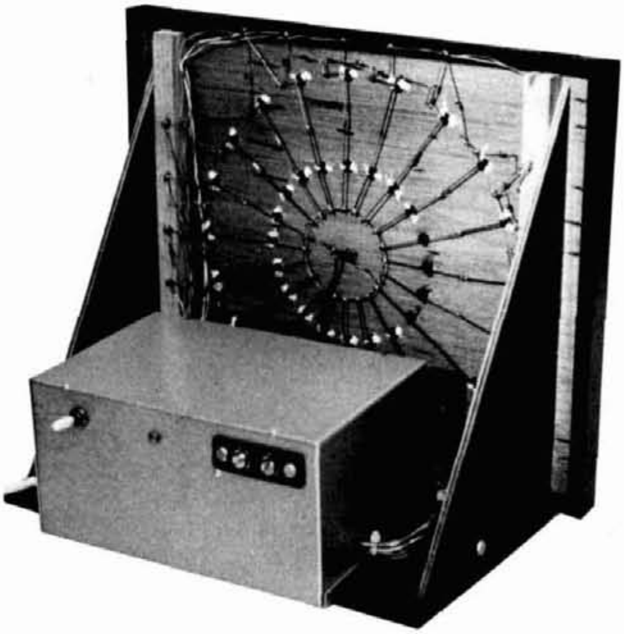
1. the system fits the decimal system;

2. 18 degrees is roughly the beam width of many rotary antennas;

3. 18-degree sectors provide enough room for a beam to coast to a stop once the motor drive is stopped.

Fig. 1, the block diagram, and **fig. 2**, the schematic, show the signal and logic flow from the input of the A/D converter to the sector indicator. The analog signal between 0 and 2 volts is applied to pin 3, the analog input, from the voltage divider, R5, which accepts the 13-volt signal from the rotator. C3 and R7 are used to provide RFI immunity.

The encoded TTL data from U1 is available in a multiplexed form at the Q_0 through Q_3 outputs. The A/D converter used in this indicator normally drives a four-digit multiplexed display, with the outputs Q_0 through Q_3 acting as the data lines, while DS1 through DS4 are the corresponding digit-select lines. For example, when the right-most digit (LSB) is to be displayed, the DS4 line is high and the data appears in a BCD format on the Q lines. As the display is scanned, the appropriate digit-select line goes high with the correct BCD code for that digit appearing on the data lines.



Rear view of the display showing the LEDs, diode-AND gates, and electronics enclosure.

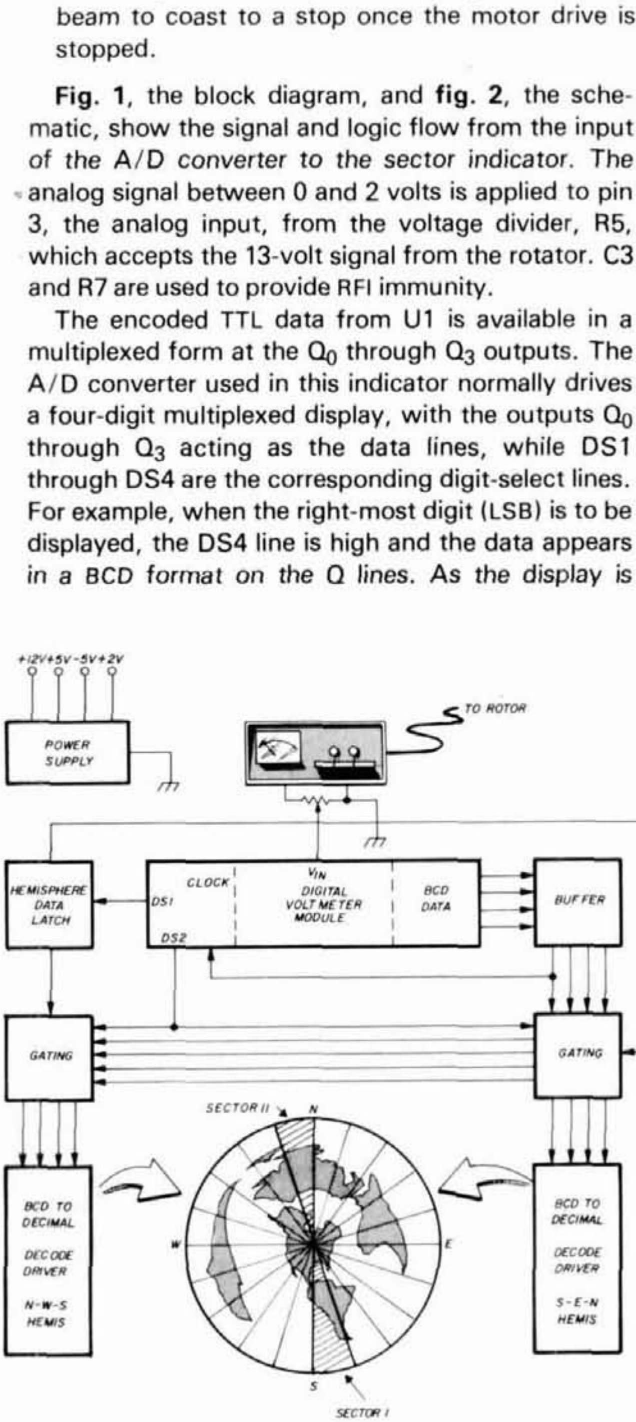


fig. 1. Functional block diagram of the solid-state antenna position display. The entire system is based on a 3-1/2-digit DVM IC manufactured by Motorola.

The analog input to this IC is in the range of 0 to 1.999 volts. Therefore, the left-most digit will change between only 0 and 1. Or, it can be thought of as breaking the input voltage range into two segments, 0 to 0.999 and 1.000 to 1.999 volts.

In my display, the left-most digit, or MSB, provides the hemisphere data. That is, pin 6 of U2C (or Q_3) is low for an input voltage to U1 with any value between 0.0 and +0.99 volt. This corresponds to an antenna heading anywhere in the S-E-N hemisphere. Conversely, pin 6 of U2C is high when the input voltage to U1 is any value between 1.0 volt and 1.99 volts. This corresponds to an antenna heading anywhere in the N-W-S hemisphere.

The hemisphere data from U6 is stored in latch U8B by clocking the data in during DS-1 time, and holding it through the complete scan cycle of the A/D converter. The latch outputs, pins 5 and 6 of U8B, gate the appropriate hemisphere decoder/driver, U6 or U7, through NAND gates U3, U4, U5B, and U5C. U8B is reset after the end of each A/D converter scan cycle by an "EOC" pulse from pin 14 of U1.

To derive the appropriate 18-degree sector within the hemisphere, the next most significant digit of the A/D converter is used. As previously mentioned, each sector has an incremental voltage width of approximately 100 mV, thus allowing ten sectors in a 1-volt increment. The next most significant digit represents the correct 100-mV segment with the

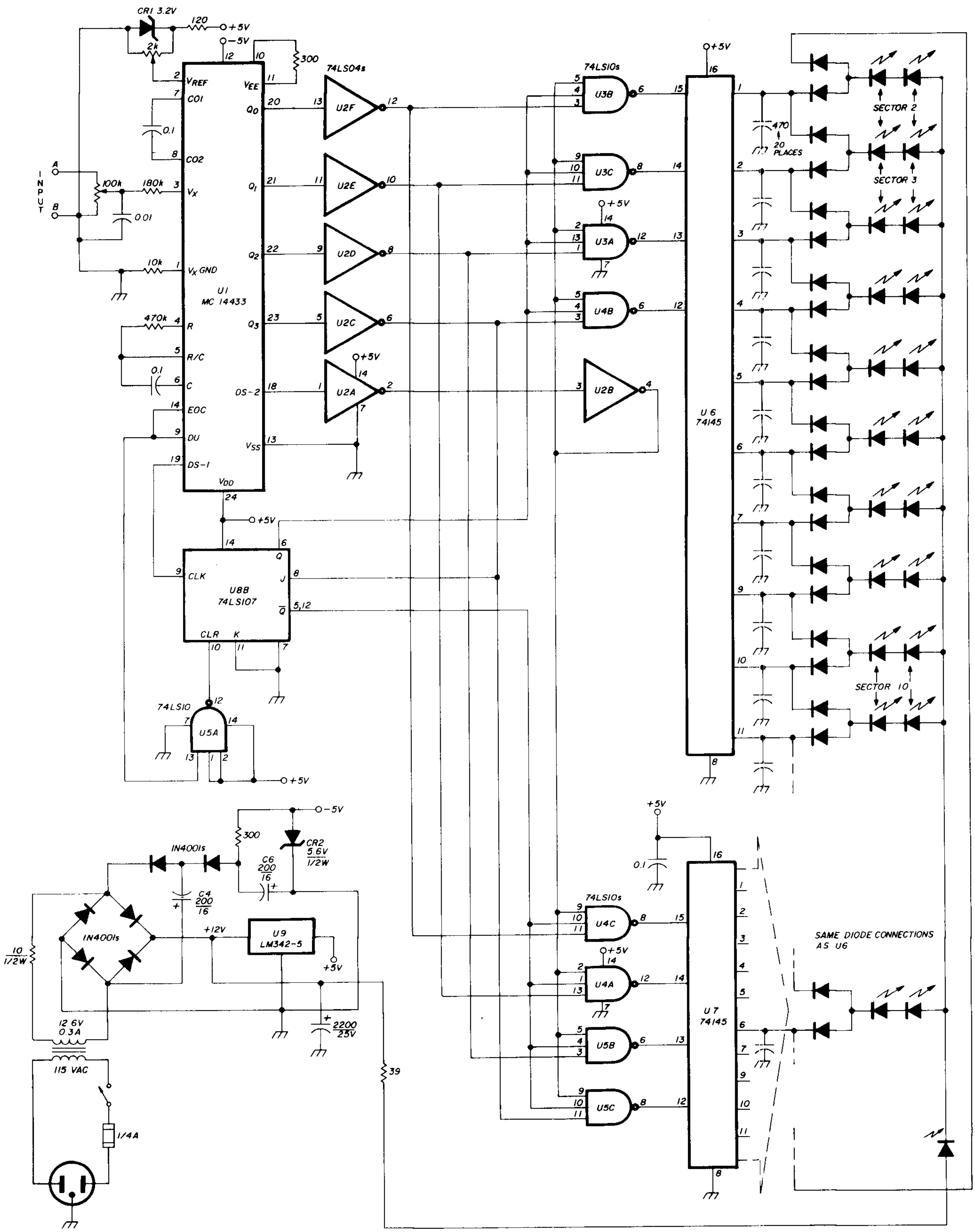


fig. 2. Schematic diagram of the solid-state antenna position indicator. Each output from the decoder is bypassed with a small disk ceramic capacitor to help eliminate noise problems (see text). C1 and C2 are mylar capacitors.

BCD value appearing during DS2 "time" at the output terminals.

During DS2 pulse, pin 18 of U1 is high. U2A and U2B act as a noninverting buffer which drives gates U3, U4, U5B, and U5C. These gates in turn drive U6 or U7 to display the appropriate sector during DS2 time. Valid input data to U6 or U7 is in a low-level state. This is provided when all three inputs to the NAND gate driver are high. The outputs of U1 require the buffering because of the limited source/sink current capability of the CMOS circuitry.

With this scheme, any one of twenty sectors can be displayed. The DVM chip can provide A-to-D conversion on the hundredths and thousandths decimal value of input voltage and are presented to the multiplexed outputs during DS3 and DS4 time respectively. However, this capability is not used in the display application.

The frequency of the internal clock in the DVM module is determined by R1 and C1. With the values shown in **fig. 2**, the clock runs at approximately 66 kHz, giving a conversion time of 250 ms. In simple terms, the conversion time is the interval required for the DVM circuitry to measure the analog input,

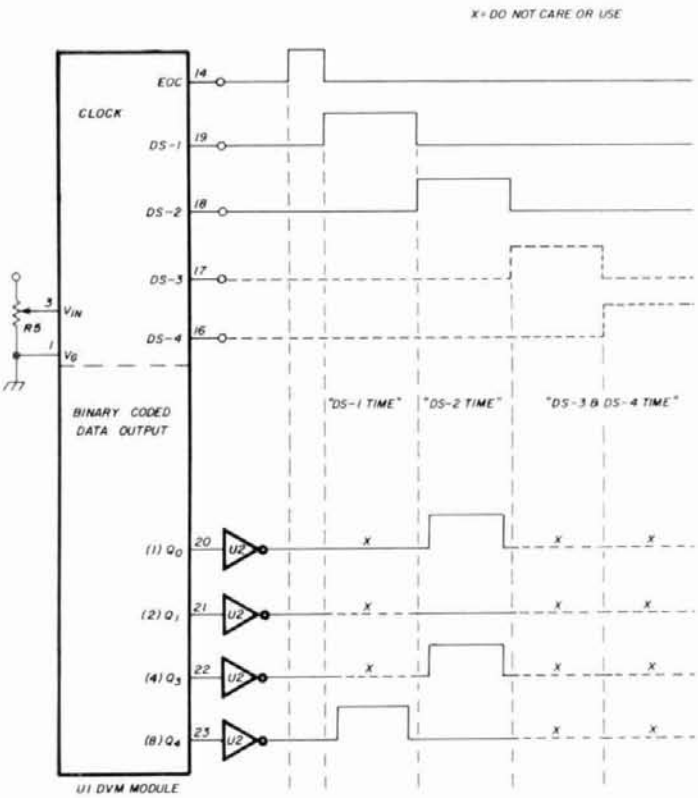
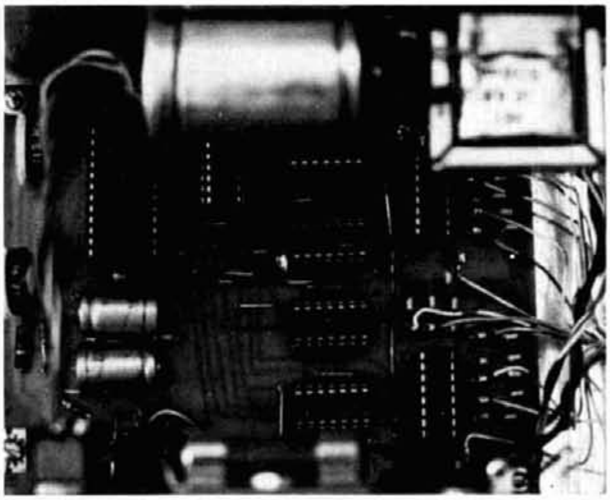


fig. 3. Timing diagram of the digital voltmeter IC. The MSB data output appears on Q_3 during DS1 time. In this example, Q_3 is a one for any input voltage between 0.000 and 0.999 volt, indicating that the N-E-S hemisphere is selected. The next most significant digit appears during DS2 time. As indicated, DS2 shows data for a value of 0.50 to 0.59 volt, indicating that the antenna is within sector 5.



Printed circuit board within the electronics enclosure.

compare references, compensate, integrate, encode data, and develop the output signals for the "3-1/2" digits. In this display application only "1-1/2" digits are used. An individual sector is displayed with five LEDs. The LED in series with R8 is positioned at the center of the display, Kansas City, and is continuously illuminated. The remaining four LEDs are selected from the display electronics and illuminate the perimeter of the sector.

LEDs DA1/DB1, DA2/DB2, etc., are positioned on radial lines drawn on the polar map, which is the background of the display panel. These radial vectors start at the center of the map and are displaced by 18 degrees.

The LEDs designated DA are mounted at the midpoint of the radial, while the LEDs designated DB are positioned at the ends of the radials. Diodes designated DC are low-cost silicon switching or low-PIV rectifier diodes (1N4001s) grouped in pairs to form a negative-OR type gating circuit.

To illustrate, refer to **fig. 4**, where sector A is to be displayed. Either U6 or U7 has been gated on during DS2 time, depending on which hemisphere has been selected, to display one of the possible twenty sectors. Only one of the twenty output pins from U6 and U7 will be in a low-voltage or current-sink condition, as represented by the closed switch at output terminal B in the driver. Current flows through R8, the LED in the center of the map, the two parallel branches formed by DA1, DB1, DC2, and DA2, DB2, DC3, and the low-impedance path (closed switch) in the display driver. With all other driver outputs in a high-impedance state, no current will flow in any of the other radial branch circuits.

Since the LEDs are illuminated only during DS2 time (approximately 60 ms out of each 250-ms conversion period), the LED supply voltage had to be raised to a level of approximately 12 volts. The value

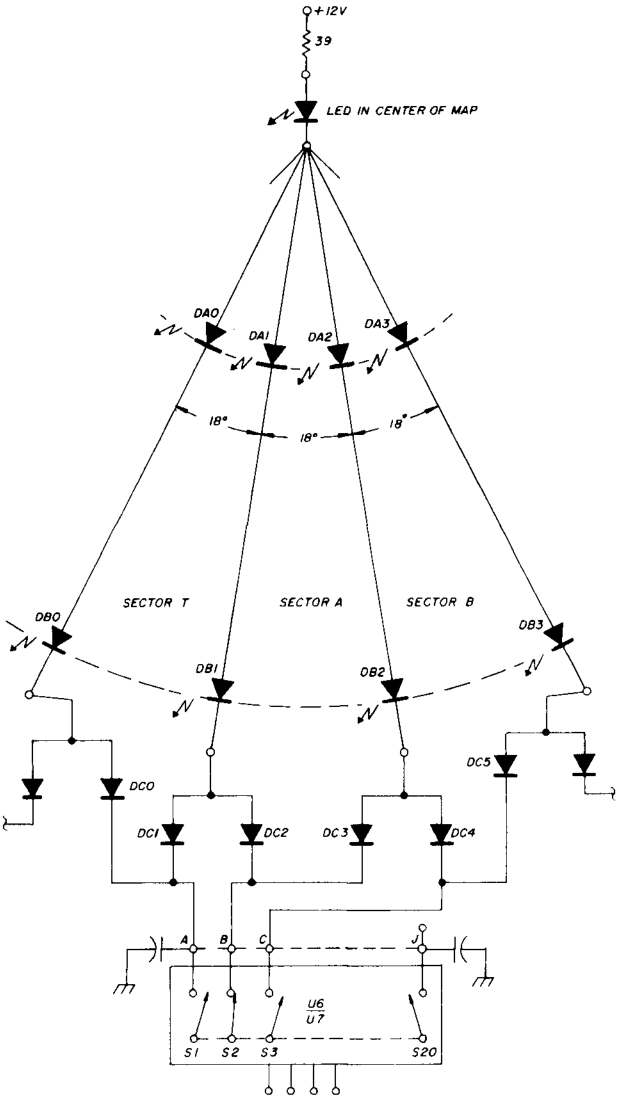


fig. 4. Diagram illustrating the current flow when sector 1 is selected. In this case, the appropriate output of the decimal driver is low, to sink current. The OR gates then determine which LEDS are conducting.

for R8 was chosen to give an average LED current of 25 to 30 mA, providing adequate illumination. Diode matching, to achieve uniform brilliance, has not been a problem. However, I suggest you order more display LEDs than needed and select the best ones.

As seen in **fig. 2**, bypass capacitors are shunted across each driver output. Any disk ceramic from 470 pF to 0.02 μF will work. These capacitors reduce the slope of the driver output signal during the switching transient. The cable between the display electronics and display panel acts as an antenna, and hiss was detected in my SB303 on 10 meters before the output lines were bypassed.

Except for the 12-volt transformer, C5, and R9, all components are mounted on the printed-circuit board. The transformer should be able to supply 300 mA. U1 requires a negative 5-volt supply, which is provided by C4, C6, R10, CR2, and the two-additional 1N4001s. The LM 342-5 provides a regulated, 5-volt supply for logic circuitry. The DVM module requires an external reference voltage of 2.0 volts which is provided by CR1, R3, and R4.

display panel construction

Many options are open in the construction and layout of the display panel, depending upon the creativity, skill, taste, and resources of the builder. A rather straightforward approach yielded the display shown in **fig. 5**. The printed-circuit board is mounted inside a 15 × 7.5 × 10 cm (6 × 3 × 4 inch) box-type interlocking chassis. The box should be mounted on the rear of the base to provide stability for the display panel and frame.

adjustment

Before inserting any of the ICs (especially U1), the +12, +5, −5, and the 3.2-volt reference should be checked. Then adjust R3 and R5 to the extreme counterclockwise position. Use a high-impedance voltmeter, 1 megohm or greater input (probe) impedance, when checking any terminal voltages on U1. Lower impedance voltmeters present a significant load to the circuitry resulting in erroneous readings. With U1 in the circuit, adjust R3 to provide +2.00 volts at pin 2 of U1. Temporarily jump a wire from a +5 volt supply point to the A side of R5, the non-grounded rotator input terminal. Only sector 1 should be illuminated. Turn R5 clockwise about half way. The display should step through the 20 sectors as this is done, and only one sector should be illuminated at any one time. If the LEDs light out of sequence, check the cable wiring between the driver outputs and diode gates in the display panel. If a

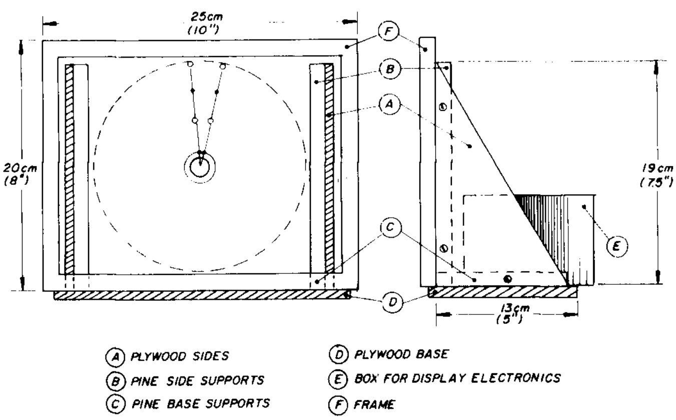


fig. 5. General construction plan for the display panel. The outline dimensions are approximate, depending on the frame used. Exact details of the map and framework have been omitted since they will depend upon the builder's resources and abilities.

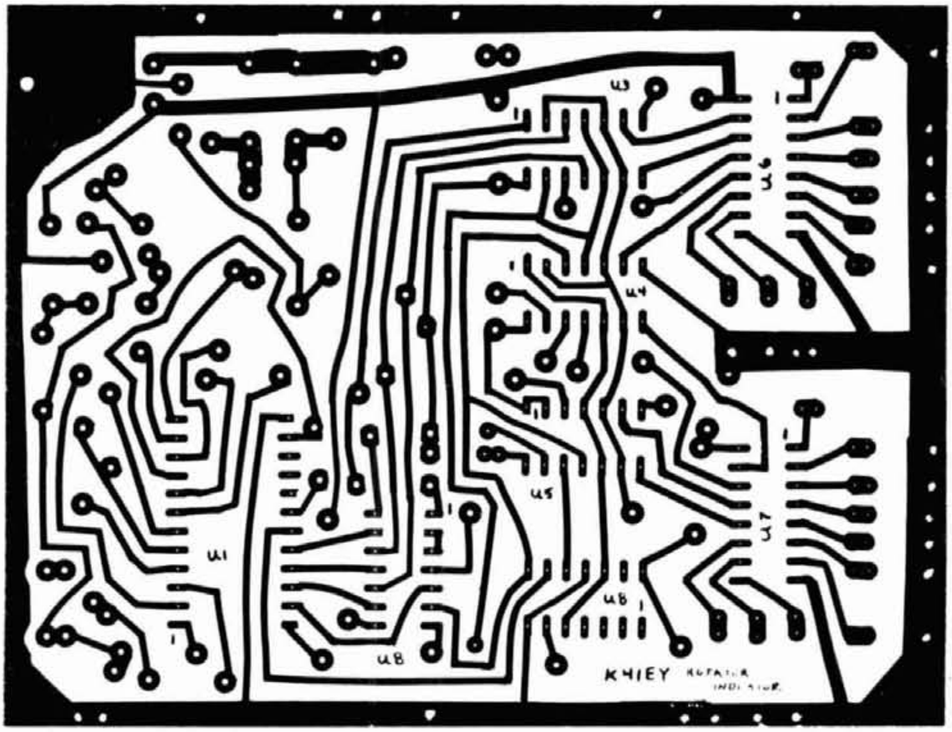
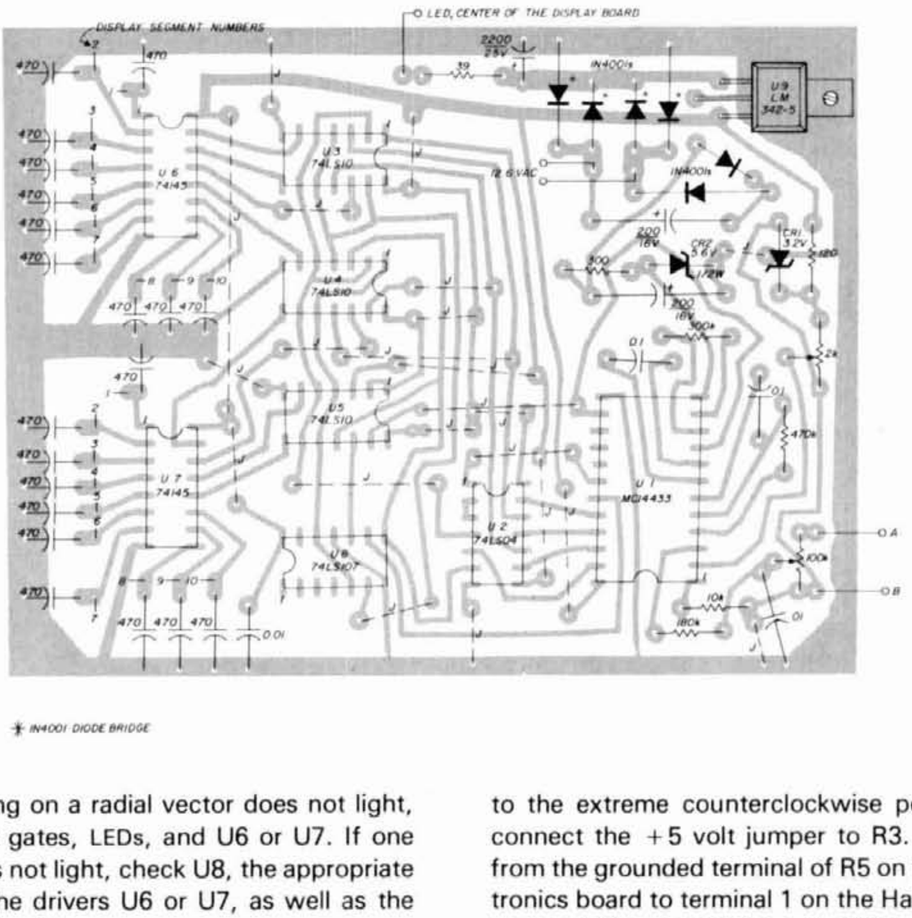


fig. 6. Printed circuit board pattern for the antenna position display (above) and parts placement diagram (below). A printed circuit board can be obtained from the author.



series diode string on a radial vector does not light, check the diode gates, LEDs, and U6 or U7. If one hemisphere does not light, check U8, the appropriate input gates to the drivers U6 or U7, as well as the appropriate display driver.

With the above check-out completed, return R3 to the extreme counterclockwise position and disconnect the +5 volt jumper to R3. Attach a cable from the grounded terminal of R5 on the display electronics board to terminal 1 on the Ham II control box terminal strip, and another lead from the other side of R5 to terminal 3. Rotate the antenna to 198 degrees in the S-SW quadrant. Approximately 13 volts should appear across R5 on the display input terminals. Carefully rotate R5 clockwise until just the 198 to 180 degree sector illuminates. Rotate the antenna back through a clockwise rotation and watch how the sectors on the map light up as the appropriate compass headings on the control box are passed. With minor adjustments of R3 and R5, tracking to within 10 degrees can be maintained throughout the entire antenna rotation. Linearity of the potentiometer in the rotator can cause minor variations.

I have not attempted to connect this display to makes of antenna rotors other than the Ham II. However, with the circuit explanation given, it should be easy to adapt this system to other rotators as long as there is more than a 0 to 2 volt analog signal swing available to drive the display electronics.

component procurement

All the ICs, with the exception of U1, the Motorola MC14433, are readily available from most supply houses. U1 was purchased through Circuit Specialists, Box 3047, Scottsdale, Arizona 85257, for under $15.00. With the influx of low-cost DVM kits, there should be lower costs for this item in the future. The total cost of the electronic components and printed circuit board came to less than $45.00. LEDs can be obtained at good prices when ordered in a quantity of 100 rather than on a per-diode basis.

I wish to express my appreciation to Dick Keil, N4JU, Bob Winter , WB4AYW, and Walt Short, N4SW, for their advice, and particularly to N4SW and K4GOK for their assistance in artwork preparation.

ham radio

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great-circle maps

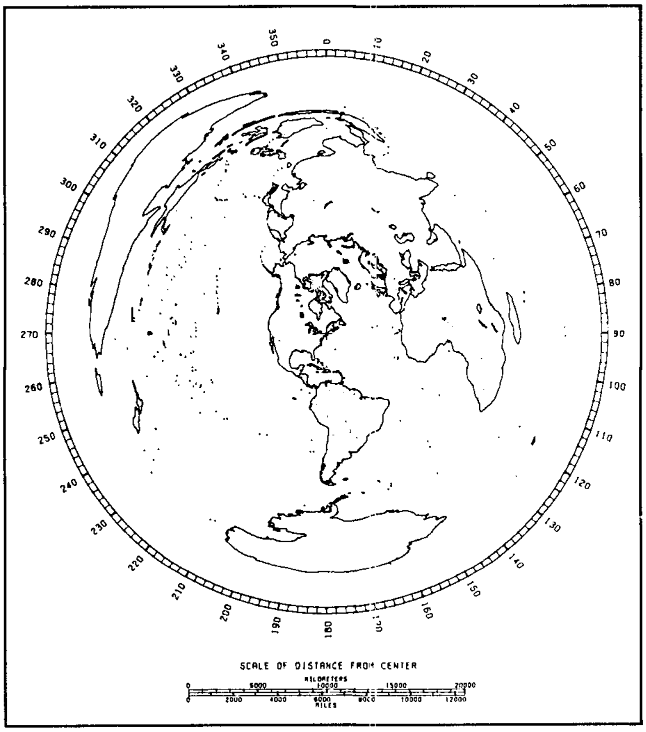
For the past several years I have been offering computer generated great-circle bearing printouts which have been extremely popular with DX operators.* In a recent article in *Ham Radio Horizons* I discussed the use of these charts and also showed several great-circle maps. The great number of readers who inquired about obtaining azimuthal equidistant maps prompted me to complete work on a computer program I started several years ago to draw such maps.

The program itself is straightforward, but the data base associated with it is truly staggering, consisting of almost 20,000 data elements. This is why I put off completing it for so long. The computer time required to process and draw each map is much more than that required for the standard great-circle printout, so the cost is slightly greater. The maps are printed on 11 × 14 inch paper; in addition to geographical data, all major political boundaries are shown, but no attempt has been made to label individual countries because of the enormous programming complexities it would entail, not to mention the additional cost.

I will supply custom-made azimuthal equidistant maps to interested *Ham Radio* readers according to the following price schedule:

- $5.00 postpaid via 3rd class mail, worldwide
- $5.75 postpaid via 1st class mail, USA, Canada, Mexico
- $6.50 postpaid via Air Mail, worldwide.

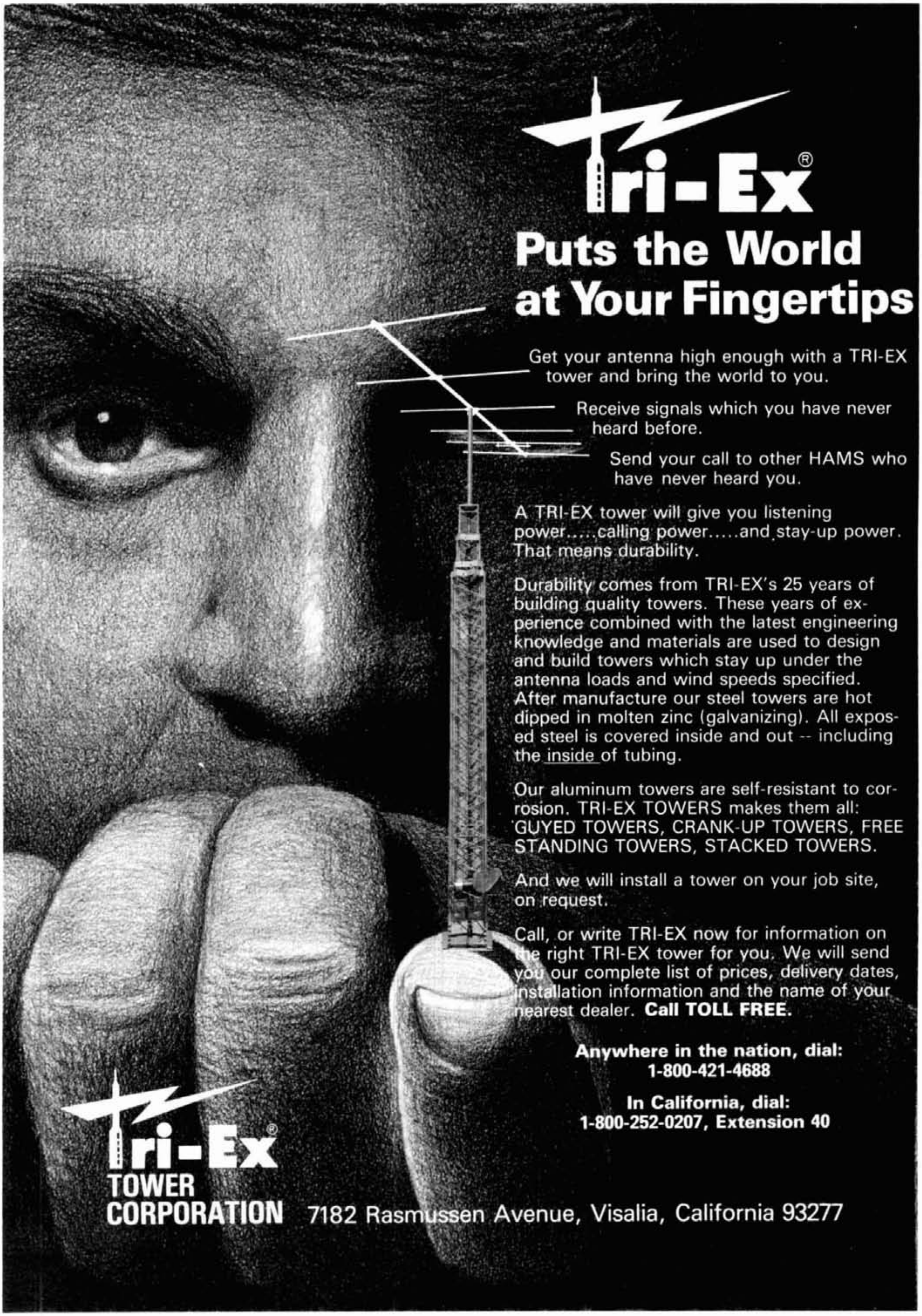
*Computer generated charts for your station location are priced at $1.00 for surface mail or $2.00 for air mail, and list 660 distant locations along with bearings, distances, and return bearings.



Computer-drawn great-circle map centered on Greenville, New Hampshire.

When ordering your map, be sure to include your mailing address and the location for which the chart is to be made. If you live in a rural area or a town of less than 10,000 population, carefully describe your location with respect to other nearby towns so your latitude and longitude can be determined.

Bill Johnston, N5KR
1808 Pomona Drive
Las Cruces, New Mexico 88001



More Details? CHECK — OFF Page 126

phase coherent RTTY modulator

Discussion of the need for using a phase-coherent AFSK system to generate FSK with a single-sideband transmitter

With amateur activity increasing on RTTY there is a growing need for a quality RTTY modulator and demodulator to interface between the station receiver and transmitter. Although some transmitters have provisions for a frequency shift key (FSK) input to the VFO circuitry, most do not. One of the standard means of producing an FSK carrier on the high-frequency bands is to insert pure tones into the audio input of an ssb transmitter. An audio FSK generator (AFSK) will produce an FSK signal when it is applied to a single-sideband, suppressed-carrier transmitter. However, there are limitations to this technique.

FSK problems and approaches

In RTTY circuits, there are basically two frequency shifts used, a narrow shift (170 Hz) and a wide shift (850 Hz). The only FCC requirement is that the shift be less than 900 Hz. The reason for a shift at all, of course, is to distinguish a mark from a space, thus conveying information. Using digital language, the mark and space may be redefined as a logic one or a logic zero. The definition of which frequency will be used as a mark (one) and a space (zero) must be compatible between all communicators, otherwise the shift will be inverted from the one expected. On the 20-meter band, for example, the mark frequency is normally defined as the higher of the two frequencies, while the space is the lower of the two. The mark and space frequencies are offset by 170 Hz.

The narrow-shift mode is almost always used on the high-frequency bands, while some amateurs use the wide shift on the vhf bands. **Fig. 1A** shows the ideal frequency spectrum of a narrow-shift FSK signal. The bandwidth due to the information rate is not shown, but it will be centered on each carrier frequency and will have the effect of widening the FSK signal spectrum. The information bandwidth depends on the speed at which the RTTY is sent, and it will not be considered here. Note in this figure that the mark frequency is exactly 14.097875 MHz, while the space frequency is exactly 14.097705 MHz. This would be a narrow-shift FSK signal, since the difference is 170 Hz.

If the frequency determining unit in the transmitter is appropriately modified, a frequency shift of 170 Hz is easily attained. As an example, an oscillator may be modulated by using a varicap to generate the required FSK signal. An alternate method, however, would be to modulate a single-sideband transmitter with pure sine wave tones. Theoretically, in an ssb, suppressed-carrier transmitter, (SSB-SC), if a single frequency is used to modulate the transmitter, a single frequency appears at the output of the transmitter. When in the upper-sideband mode, the frequency that appears is the sum of the modulating frequency and the suppressed-carrier frequency, or the suppressed-carrier frequency minus the modulating frequency when in the lower-sideband mode. By changing the audio frequency, a signal can be generated which is FSK modulated in step with the audio signal. Although the signal fed into the transmitter is AFSK, the signal generated by the transmitter is a true FSK signal; only one rf frequency exists at a time at the output.

There are certain demands put on the transmitter

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when either FSK or AFSK is used. If FSK is implemented by modifying the VFO, then the short-term and long-term stability of the shift circuits must be considered. Since the frequency-determining section of a transmitter is being modified, the circuits must not cause drift in the oscillator that will make the selected frequency unstable. For instance, if a dc potential is used to vary the capacitance of a varicap diode, the dc may have an ac ripple component. This ripple, from a poorly filtered dc power supply, will cause fm modulation to appear on the shifted-carrier frequency. This is certainly not desirable. Also, any temperature drifts in the dc control circuits or in the varicap diode response will cause a frequency change to occur in the fsk carrier frequencies. Care must be exercised to ensure that any modifications do not influence the stability of the oscillator.

With AFSK modulation applied to a transmitter, another set of problems appear. Fortunately, these problems are not associated with the stability of the VFO, since it is not modified. However, the quality of the ssb generation (carrier suppression, sideband suppression, and sideband filter response) is important. In a SSB-SC system, a double-sideband signal is first generated with the carrier suppressed. Normally, a sharp filter is used to pass only the desired sideband, either the upper or the lower. Because of this method, the unwanted sideband will be present, along with the carrier, although suppressed to a large degree (see **fig. 1B**). When operating an SSB-SC transmitter as an FSK generator, care must be exercised to ensure that the carrier is properly balanced out, and, in addition, the unwanted side-

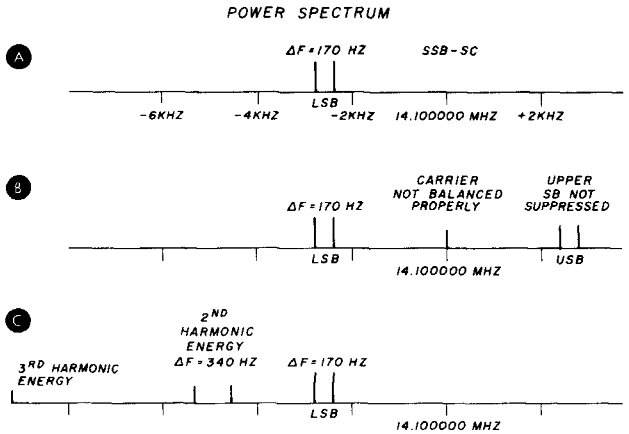


fig. 1. Frequency spectrum of an FSK system on 20 meters. (A) shows the ideal spectrum, with the mark being the higher of the two frequencies. In all cases, the information bandwidth is not considered. In (B), you can see the consequences of improper carrier balance and poor upper-sideband suppression when using an audio tone to generate an FSK signal. The spectrum when the sideband filter does not properly suppress harmonics is seen in (C).

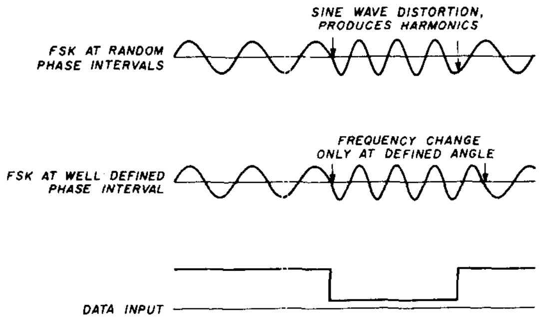


fig. 2. Diagram of coherent and incoherent frequency shifts. The incoherent shift results in sinewave distortion during different parts of the cycle, while a coherent frequency shift has a smooth phase transition, resulting in less distortion.

band is adequately filtered. Carrier null is probably the worst culprit, since balance circuits will change with time. The balance should be periodically checked, otherwise a continuous carrier will be present along with the FSK information.

The sideband filter response is important because of the distortion present in the modulating frequency. Although a sine wave modulating frequency is prescribed, it is extremely difficult to attain. Even if a perfect (no harmonics) sine wave were used, any nonlinearities in the audio stage would introduce some distortion into the modulation. For example, if a frequency of 2125 Hz is used to modulate the transmitter, the sideband filter should suppress the second, third, and other high-order harmonics. **Fig. 1C** shows the effect of insufficient harmonic suppression when a mark or space frequency is being transmitted. As was mentioned earlier, even a pure sine wave injected into the transmitter audio input will end up distorted because of preamplifier nonlinearities. This preamplifier induced distortion will be reduced somewhat due to the response of the sideband filter. If a low audio frequency is used for the mark or space, the second and third harmonic may fall into the passband of the sideband filter. If, on the other hand, the mark and space frequencies are chosen to be high enough in the response band of the filter, the sideband filter will reject the harmonic energy created by signal distortion.

The importance of undistorted wave forms is clear when considering the frequency spectrum of an FSK signal. Too much distortion and the harmonic content is more than the filters can adequately remove. Thus, steps should be taken to ensure that a reasonably clean modulating waveform is applied to the audio input of the transmitter. At the time of the frequency shifts, the phase transitions from one frequency to the other should not contain discontinuities which would contain energy at frequencies other than the space or mark frequencies. This basically means that all space and mark frequency changes should be done smoothly (see **fig. 2**). One way to guarantee this is to always change phases at a zero-crossing of the signal. Using this approach, the audio modulating signal will always change frequency at the same phase point in all information transitions.

and higher-order odd harmonics cause distortion. The lowpass filters are designed to attenuate the third harmonics by 40 dB. The third harmonic of a squarewave is normally 10 dB lower than the fundamental, so this additional 40 dB should put the third harmonic 50 dB below the fundamental. Of course, in practice, this may not be obtained, but the low-

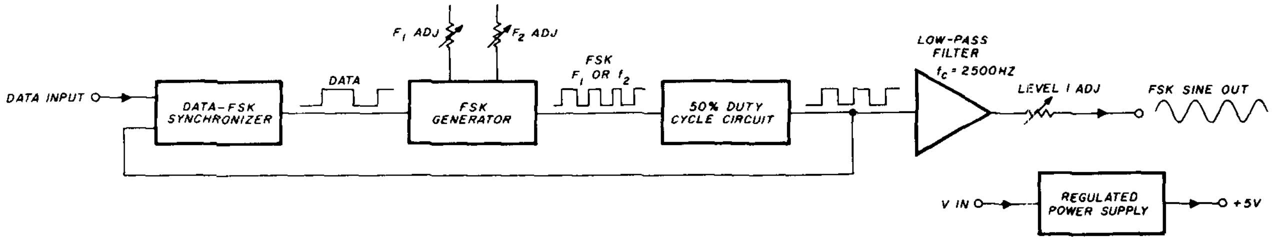


fig. 3. Block diagram of an AFSK generator capable of coherent frequency shifts. Lowpass filters reduce the harmonic content to acceptable levels.

A block diagram of an AFSK generator is shown in **fig. 3**. The generator begins with an oscillator set to twice the needed frequency. When a mark or space is needed, the oscillator's frequency is adjusted for the proper frequency change. A divide-by-two circuit generates the correct frequencies for the AFSK output waveform, with a lowpass filter removing objectional harmonic energy from the square wave — resulting in a near sine wave output.

Because a square wave is used, the even harmonics theoretically do not exist. Thus, only the third

pass filter response coupled with the bandpass response of the sideband filter in the transmitter should give satisfactory suppression of all harmonics. It should be noted that the mark and space frequencies chosen (2125 and 2295 Hz) must be within the response of the sideband filter in the transmitter. Otherwise, these frequencies will be attenuated.

There is nothing magic about these modulating frequencies. Frequencies of 1800 and 1970 (170 Hz shift) could have been used. However, the receiver demodulator must be matched to these same fre-

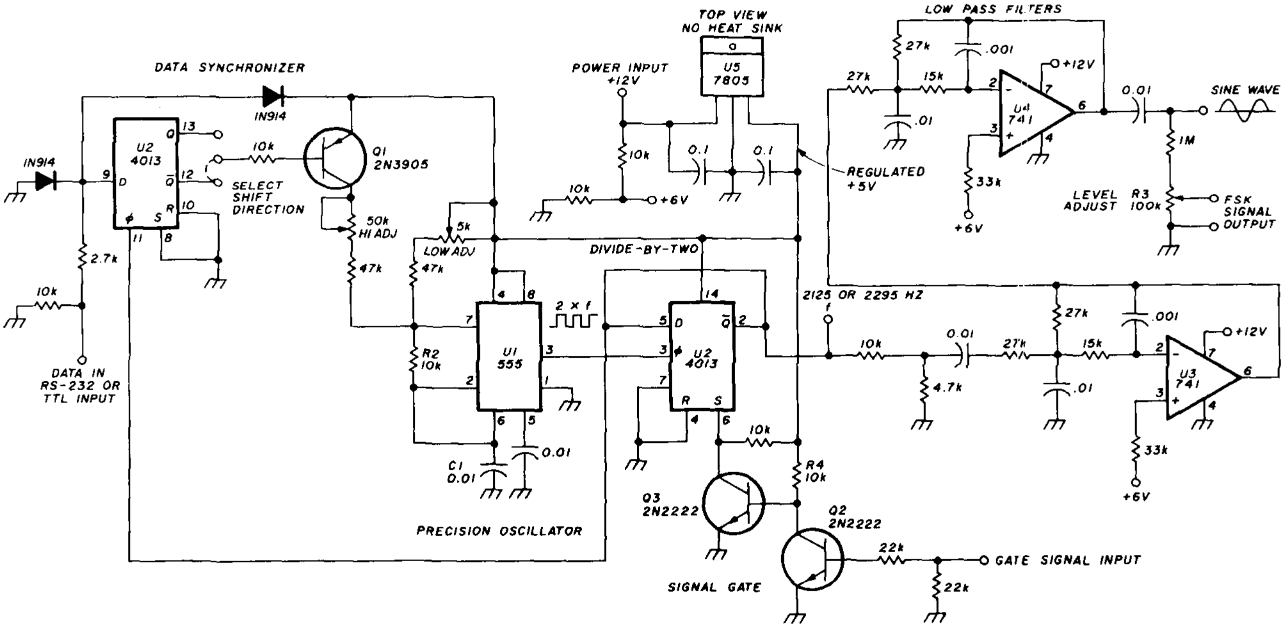


fig. 4. Schematic diagram of the AFSK generator described in the text. The active filter removes harmonic energy while the flip-flop synchronizer only allows phase coherent frequency shifts to occur. All parts associated with the 555 oscillator should have a low temperature coefficient to reduce frequency drift due to temperature change.

quency pairs. This assumes that the receiver-transmitter combination operates on the identical frequency.

circuit description

A circuit which reflects the block diagram just discussed is shown in **fig. 4**. The ubiquitous 555 astable oscillator is used as the frequency generator. Two miniature potentiometers are used to adjust the mark and space frequencies (actually twice the required frequencies). By inputting a logical one or zero, transistor Q1 turns on, changing the RC time constant. For the 555 oscillator, the frequency of oscillation is:

$$F = \frac{1.44}{(R1 + 2R2)\,C1}$$

where R1 is the parallel combination of the resistors used for frequency setting. Normally, the frequency of the 555 is set to twice the mark and space frequencies. U2 is used as a divide-by-two circuit which will guarantee a 50 per cent duty cycle. The other half of this dual-D flip-flop is used as the input data synchronizer. The mark or space input will affect the frequency of the 555 only when the output of the divide-by-two changes state. Thus, all frequency shifts are synchronized by one well-defined phase point in the oscillator's period. The divide-by-two circuit is easily disabled, creating a convenient method of gating the oscillator. This is useful for gating the oscillator off and on to a CW identification. Q2 and Q3 simply buffer the input, which inhibits the divide-by-two. U3 and U4 are used as a dual-stage, active lowpass filter. These filters each have a two-pole Butterworth response. Each lowpass response results in a 40 dB per decade roll-off characteristic. In tandem the responses add, yielding and overall 80 dB per decade response. The lowpass filters use inexpensive 741-type operational amplifiers. The last output of the filter is attenuated by R3. This potentiometer can be adjusted to set the output drive level feeding the audio input of the transmitter.

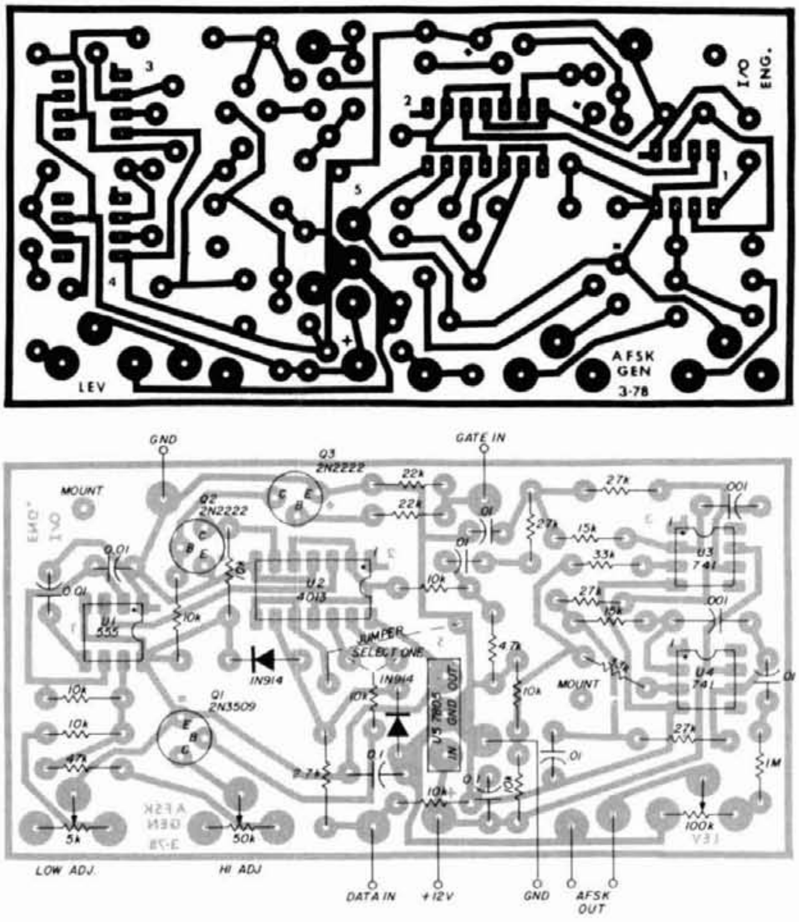


fig. 5. Foil pattern (above) and parts placement diagram (below) for the phase coherent RTTY modulator.

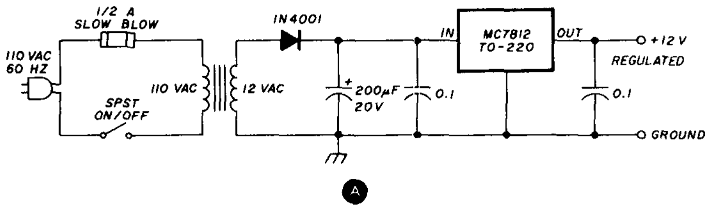
The circuits are powered from a 12- to 15-volt dc source. The voltage reference for the oscillator section, however, is regulated by a three-terminal regulator to ensure that voltage fluctuations will not influence the frequency of oscillation. Even though the 555 oscillator has a specified 0.1 per cent volt tolerance to power supply change, I have found it best to regulate the power source to remove any problems with errors due to this change. If you use a well-regulated power supply, the 7805 three-terminal regulator may be omitted and a jumper wire installed.

Layout of the circuit is not critical since only audio frequencies are generated. A printed-circuit-board foil pattern is shown in **fig. 5**. The power bus should be filtered to remove dc transients which might be propagated from other circuits attached to the same power source. A typical power supply is shown in **fig. 6A**. Also shown is a circuit which may be used to convert a TTY current loop to the proper voltage for driving the modulator input. For those who wish to duplicate this modulator, a circuit board is being made available.*

adjustment

Once the circuit is constructed, checkout is relatively straightforward. A 12-15 volt power supply should be connected between the power input and ground. Since the "shift direction" is jumper programmable, one or the other polarity for the shift direction should be selected. If needed, a single-pole double-throw switch could be used to remotely select the shift direction. Assuming pin 12 is selected for the shift direction and no "Data-in" signal is present, Q1 will be turned off. With Q1 off, the "Low-Adj" potentiometer should be adjusted for an output frequency of 2125 Hz. This frequency can be measured at the FSK signal output port, or monitored at pin 2 of U2. Once the frequency is brought into the proper range, "Data-in" should be connected to a logic one level. Q1 will now turn on, and the "High-Adj" potentiometer should be adjusted for an output frequency of 2295 Hz.

If no signal appears at the output or at pin 2 of U2, measure the level at pin 6 of U2. It should be at ground potential when no "Gate" signal is present. If it is not at ground potential, Q2 and Q3 may be the wrong type or inserted improperly. If the frequency can not be adjusted to the proper range, C1 may be at fault. Since C1 determines the timing, only a high-quality capacitor should be used. Any temperature drifts of this capacitor will create a proportional drift in the oscillation frequency. Although a 0.01-μF capacitor is shown, slight variations of this value are permissible due to the use of potentiometers for determining the frequency. However, extreme variations may not work because of the limited range of the potentiometer adjustment. Therefore, if the frequency will not adjust to the exact frequency needed, try using another capacitor for C1. If all else fails, a smaller-value capacitor could be paralleled with C1 to "fine tune" the frequency.



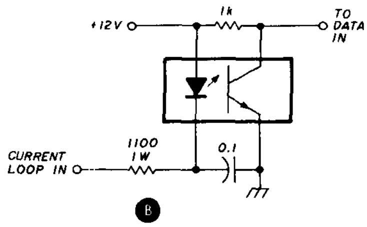


fig. 6. Typical power supply for use with the RTTY modulator is shown in (A). A TTY current loop interface, using an op-isolator, is shown in (B).

After the oscillations are set to within the tolerance wanted, go back and check the frequencies as the "Data-in" line is switched from a logic one to a logic zero. The "Gate" signal input can next be checked by connecting a logic one voltage to the input. The output oscillations should cease. Note, if the opposite logic polarity is needed for the disable gate input (logic zero for disable), Q2 and R4 can be eliminated, but be sure to jumper between the collector and base of Q2. This is why two transistors are used in the Signal Gate circuit, to allow for the option of inverting the gate signal.

I should mention that the 4013 dual flip-flop is a CMOS device. Thus, care should be exercised when handling the unit because static buildup can damage the sensitive MOS input transistors. Also, beware of bargain basement CMOS devices. I have seen some ICs purchased from outlets which by no means met specifications. All outputs should swing from the power supply potential, for a logic one, to practically ground potential, for a logic zero. This assumes no current is being "sourced" or "sinked" by the outputs. If the CMOS device does not meet this simple criterion, send it back; it is defective.

I hope this will clarify the AFSK approach for generating FSK with single-sideband equipment. The pitfalls to avoid should be recognized for compliance with FCC regulations and for reducing interference on the amateur bands.

ham radio

*A predrilled, single-sided printed circuit board is available for $5.00 postpaid from I/O Engineering, 12412 Mossy Bark, Austin, Texas 78750.

More Details? CHECK—OFF Page 126

time-current charging

of nickel-cadmium batteries

Time-current charging: a technique for quickly charging nickel-cadmium batteries

Want to charge sealed, nickel-cadmium batteries quickly and safely? The dump, time-current charging method is not new, but it seems little known by many people using nickel-cadmium batteries in electronic equipment.

Sealed nickel-cadmium cells may be charged and discharged at very high rates (high currents), if certain rules are observed. When discharging, overheating of the cells should be avoided. Not letting them get too hot to handle is a safe rule. (Note: Cells used in portable soldering irons are effectively short circuited by the low-resistance soldering element; for short periods, they may supply hundreds of amperes without damage.)

In the case of charging, the same rule applies — with one major limitation. This limitation is that high currents *must be avoided* when the cell is near or above full charge. When above full charge, the cell will produce gas if it is charged at a rate above 10 per cent of its (one hour) ampere-hour (A-H) rating. In most cases, this is the recommended slow-charge rate and is the rate that can be used for prolonged periods of overcharging without apparent damage.*

*If you are going to trickle charge batteries during idle periods, a rate of 1 per cent of ampere-hour rate would probably be more reasonable.

Open cells are not an altogether different matter. Most of this article applies also to that type of cell. However, the following items are important if you use open cells. Always open the filler vent when charging; do not trust any automatic vent that may be provided. Unless the cell has leaked, add only distilled water to bring the electrolyte back to the proper level. Charge the cell until it freely "outgasses"; that is, until many bubbles start to rise in the electrolyte. The dump, time-current charging method can be used with open cells. However, since the gas pressure problem does not exist and a good "full

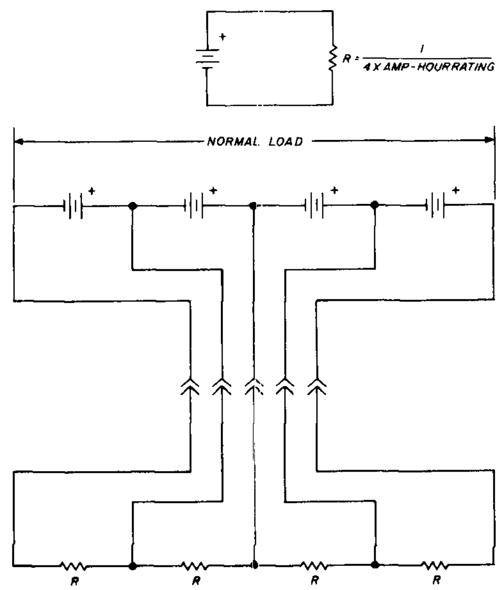


fig. 1. The dump circuits in this diagram can be varied to suit any situation. If the batteries are not soldered into the circuit, a simple battery holder, with dumping resistors soldered across the terminal, can be used. A voltmeter can be used to measure the cell voltages. Discharge each cell to about 0.5 volt; at 0.5 volt, the cell has less than 5 per cent of its full charge.

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charge" indicator does exist, the method is not very useful.

time-current charging

The basics of this system are to first completely discharge the cell and then to recharge it to less than 100 per cent full charge with a known high current for a specific length of time. It may be used at this point, or, if full capacity is required, charging may continue at the normal 10 per cent rate.

To avoid reverse charging, it is very important that fast discharging be done *individually* on each cell. If you are working with a battery of cells, a jig must be made to separately discharge the cells. Even at normal discharge rates, care must be taken not to discharge cells connected in series. In some cases, the discharged cells are reverse-charged and frequently become reverse-polarized. When this occurs, the cell will not recharge in the normal manner; it will retain its reverse polarity. Sometimes the cell can be brought back to normal polarity by giving it a massive charge in the proper direction. Typically, half-ampere cells are charged at rates of several amperes for a few minutes. This "cure" works in many cases, but the reliability of the cell is questionable from that point on.

dumping

Discharging (or dumping) can be safely accomplished at four times the rated one hour A-H current. Typically, a four A-H cell can be safely discharged at 16 amperes. In this case, a full charge will take about fifteen minutes to dissipate. If the cell is less than fully charged, correspondingly less time will be required. Satisfactory values of resistance for several popular nickel-cadmium cell sizes are given in **table 1**. The circuit for discharging single or multiple cells is shown in **fig. 1**.

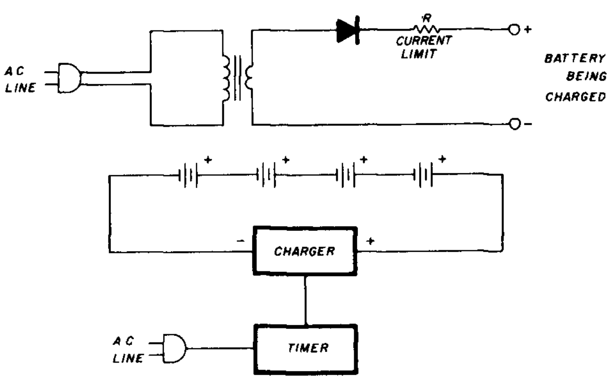

fig. 2. Diagram of a simple charging circuit. Commercial chargers can be used, or a simple transformer/diode circuit can be built.

Table 1. Resistance values needed for a cell discharging system.

cell size	A-H rating	discharge current amperes	resistance	minimum wattage
D	3.5 A-H	12.0	0.1 Ohm	25.0 Watts
C	1.5	6.0	0.2	10.0
AA	0.5	2.4	0.5	5.0
–	0.25	1.0	1.2	1.0

charging

Charging is most effectively done with the cells connected in series. This allows a single charger to put the same charge current through all of the cells simultaneously (see **fig. 2**). Charging can be done at currents as high as 50 times the one-hour ampere-hour rating of the cell; a 150-mA cell can be charged at 7.5 amperes. The charging time is calculated as follows:

$$time = \frac{A\text{-}H\ rating}{charging\ current}$$

$$= \frac{0.150\ A\text{-}H}{7.5\ A}$$

$$= 1.2\ minutes$$

This short a time, however, is an extreme that should be avoided because of the timing accuracy required. Missing by a few seconds could lead to an accident. A misrating on the cell could be equally dangerous. If you choose a rate of five times the A-H rating, the time would be 12 minutes and the time tolerances become reasonable. Plus or minus one minute will result in about 10 per cent of full charge.

At any rate of charge, the 100 per cent charge time may be calculated using the previous formula. Although the prime advantage of the dump, time-current charge method is speed, somewhat slower discharge and charge rates will tend to be safer than high rates, which may ruin a battery if care is not used. I strongly recommend that a timer be used to turn off the charger, rather than trusting the clock-watching method.

acknowledgement

The assistance of Mr. E. L. Williams in the preparation of this article is gratefully acknowledged.

bibliography

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quartz crystals — gems for frequency control

A *misunderstood hero* *of the electronics world* is the quartz crystal. Quietly it awaits your command to put your transmitter on frequency, to reject all but one sideband, or to select one rare CW signal sandwiched between adjacent kilowatt signals. What is the secret of quartz? Can an amateur operator zero-adjust his crystal oscillator, or is he stuck with a bad crystal? How do these pieces of quartz operate in an oscillatory circuit? It's hoped that this article will answer some of your questions and help you in procuring and designing circuits with that celebrated mineral.

the quartz crystal — some background

Quartz technology is based on its piezoelectric property. The application of an electric field causes certain substances to oscillate; conversely, the application of a mechanical force or vibration causes substances to generate an electric field, known as the piezoelectric effect. Quartz is useful as an electrical oscillator operating in a very narrow frequency

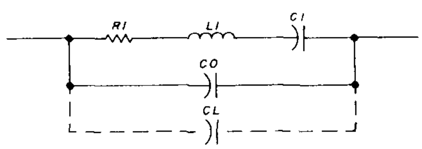

fig. 1. Equivalent circuit of the quartz crystal. L1, C1, R1 are the primary, or motional components, which determine frequency and circuit Q. Capacitance C_0 is the electrode, mounting structure, and holder capacitance. Capacitance C_L is the parallel capacitance across the circuit.

band. The precise frequency, activity, and temperature characteristics are determined by the position and angle of cut on the crystal.

The old concept that the quartz crystal is a *standard of frequency* was born in an age of less-critical applications. Old timers knew that the crystal was much more accurate and repeatable than any LC circuit. Because they didn't have to multiply 18 times and trigger a repeater, it's easy to see how the legend of quartz stability became exaggerated.

The basis for stability in quartz is its high inductance and low capacitance, resulting in extremely high Q. In an 8-MHz crystal unit, for example, the Q might be 150,000 while the Q of a typical LC combination at that frequency is about 300. Yet a crystal's frequency may be pulled; and in time, it will drift.

equivalent circuit

The simplest and most commonly used equivalent circuit of the crystal is shown in **fig. 1**. $L1$, $C1$, and $R1$ are the primary components which determine frequency and Q. These are referred to as the *motional components* and their parameters can't be measured directly. C_0 represents the electrode capacitance, the mounting-structure capacitance, and holder or case capacitance. C_0, the static capacitance, affects the crystal operating frequency, but to a lesser degree than $C1$. C_0 can be measured by a capacitance bridge across the terminals. As you may expect, the capacitance of the circuitry, shown as another parallel capacitance, C_L would also have an effect on the crystal working frequency. The equation for the working frequency is

$$F_W = \frac{1}{2\pi\sqrt{L1\left[\dfrac{C1(C_0+C_L)}{C1+C_0+C_L}\right]}} \qquad (1)$$

With the help of a calculator, you can determine how much the crystal is pulled by the oscillator circuit. By changing the circuit loading, the crystal may be pulled (within limits) for fine tuning or fm applications.

mode of operation

The classic crystal reactance curve, (**fig. 2**), is useful in demonstrating the relationship of different operating frequencies. At two points the reactance is zero; *i.e.*, the crystal looks purely resistive. The lower of these frequencies is the series-resonance frequen-

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cy, (F_S); while the higher frequency is the anti-resonance-frequency, (F_A). The resistance is low at series resonance and high at anti-resonance. The range of frequencies between is known as the natural bandwidth of the crystal. Anti-resonance is a very unstable point, and for amateur purposes, may be forgotten.

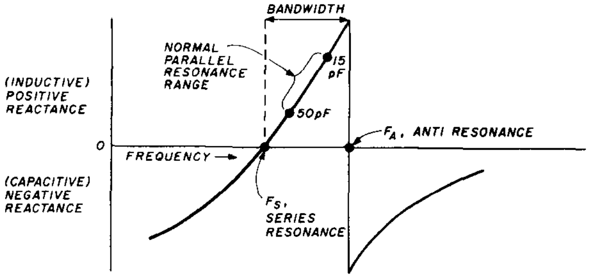


fig. 2. Classic crystal reactance curve, which is useful in demonstrating the relationship of different operating frequencies.

Parallel resonance is commonly recognized as the band of frequencies between F_S and F_A, although classic crystal theorists have another definition. You may think of this band as the range where the crystal will operate if a capacitor is placed in parallel with it. At F_S the capacitance will be infinite; at F_A, the capacitance will be zero. Practical limits are between 15 pF and 50 pF, where poor stability exists at the

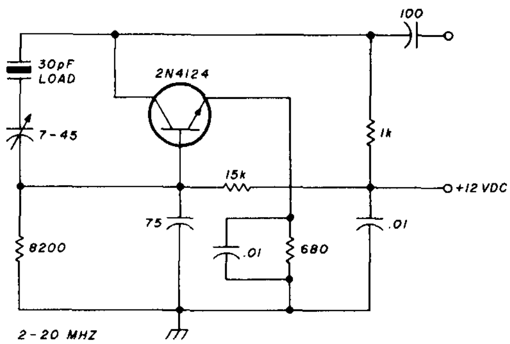


fig. 3. Example of crystal loading. The trimmer capacitor has a negative reactance, so the crystal frequency is shifted into the positive-reactance region of the reactance curve (fig. 2).

low-capacitance end and reduced activity degrades the high end. When ordering a crystal, you must specify *series resonance* or *parallel resonance at a specified load capacitance*.

There's a lot of confusion about crystal loading. Responsibility for this confusion falls directly onto the quartz-crystal industry, whose members have never acted together to educate users. Guidance committees have recommended that we consider the

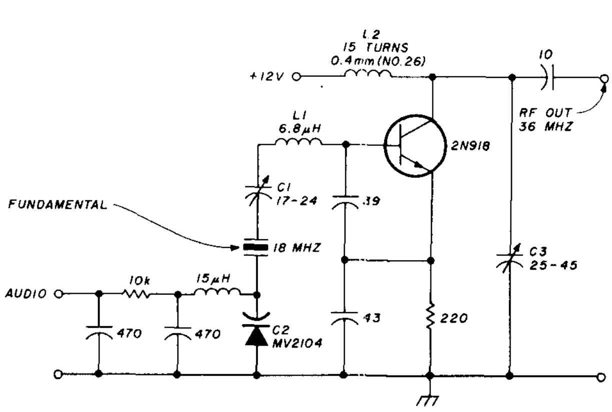


fig. 4. In this example the crystal operates into a complex load at series resonance. L1, C1, and C2 balance the crystal at zero reactance. Capacitor C1 fine tunes center frequency. Tank circuit L2, C3 doubles the output frequency. Circuit operates as an fm oscillator-doubler.

crystal operation in the *positive-reactance* mode when above series resonance and in the *negative-reactance* mode when below series resonance. This recommendation is technically correct and allows us to discuss a useful range of operation (below F_S) of the crystal, which is not usually considered.

In **fig. 3**, the crystal is in a feedback circuit from collector to base. A trimmer capacitor in series shifts the point on the reactance curve where the crystal operates, thus providing a frequency trim. The capacitor has a negative reactance so the crystal is shifted to operate in the positive reactance region of the curve (**fig. 2**).

The series trimmer does not mean the crystal is

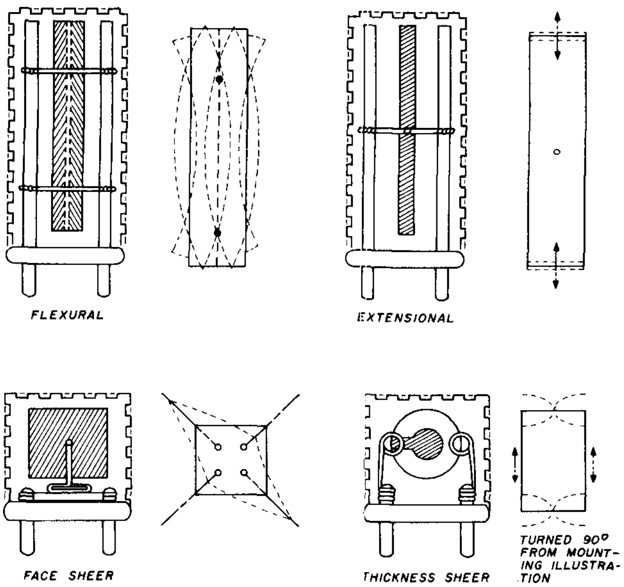


fig. 5. Examples of vibration modes and mounting structures. Note that the quartz supporting structures are fastened at points of least motion (nodes).

operating at series resonance. It is said to be operating in the parallel-resonant mode with a load capacitance approximately equal to the trimmer value. If an equivalent circuit were drawn, you could see that the trimmer would be the parallel load. By placing the capacitor in series, you isolate the crystal from other circuit reactances, enabling the trimmer to tune more effectively than in the parallel connection.

Oscillators using fundamental crystals usually operate in the positive-reactance mode with a trimmer for exact tuning. One reason for operating this way is seen when the trimmer is removed, leaving only the crystal in the feedback circuit. The crystal should operate at series resonance, but circuit reactance will usually pull it slightly off frequency.

In **fig. 4** the crystal operates at series resonance into a complex load. *L1*, *C1*, and *C2* balance the crystal at zero reactance. Capacitor *C1* fine-tunes center frequency. The tank circuit, *L2*, *C3*, doubles the output frequency. As the audio signal varies *C2* capacitance, the crystal will operate alternately in the positive, then the negative-reactance mode.

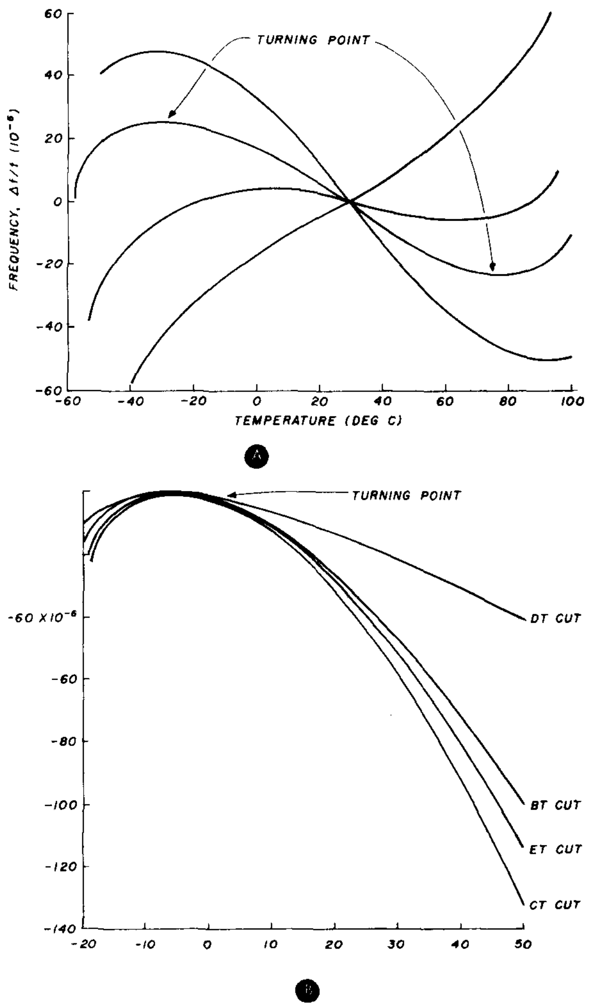


fig. 6. Typical family of AT-cut frequency/temperature variations for change of angle only (thin plates), (A). These are cubic functions centered on 27 degrees C. Sketch (B) shows frequency/temperature curves where the point of zero temperature coefficient can't be controlled. These curves are typical of low-frequency cuts.

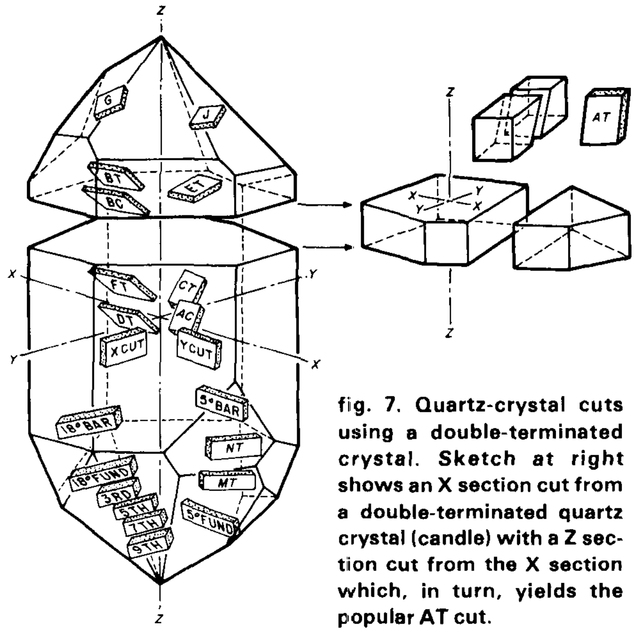


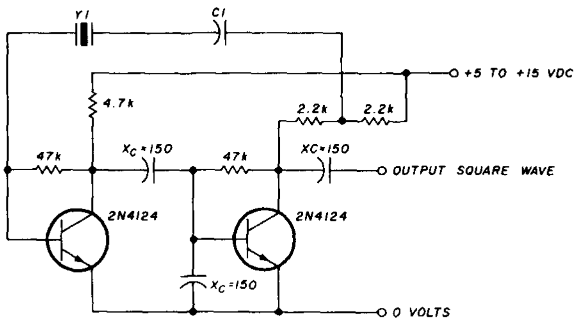
fig. 7. Quartz-crystal cuts using a double-terminated crystal. Sketch at right shows an X section cut from a double-terminated quartz crystal (candle) with a Z section cut from the X section which, in turn, yields the popular AT cut.

Deviation per volt of modulation is greater at series resonance than it would be into a capacitive load (positive-reactance operation). It would be even more desirable to operate this circuit completely in the negative-reactance mode because of more favorable deviation per volt (see **fig. 2**). The average amateur would have a problem designing a circuit for the negative-reactance mode because he must order his crystal at a higher frequency than the design frequency. Crystal manufacturers do not calibrate their crystals to tune into an inductive load. A second, and stickier, problem is that one manufacturer's crystals are more easily pulled than others. Plainly, some crystals won't work in a design acceptable for another crystal.

Using a small inductor in series with the crystal is usually a practical way to lower the frequency slightly. It's the only way to lower a crystal frequency operating at series resonance. A trimmer capacitor, also in series, can be used for fine tuning. Inductor values will depend on the crystal frequency but will be microhenries or fractional microhenries for 1-20 MHz crystals. As suggested before, not all crystals of the same frequency will shift equally. There's a limit to how much each crystal can be pulled and still operate reliably. It's good practice to see if the oscillator will start and maintain oscillation under extremes of temperature and voltage.

practical circuits using fundamental-mode crystals

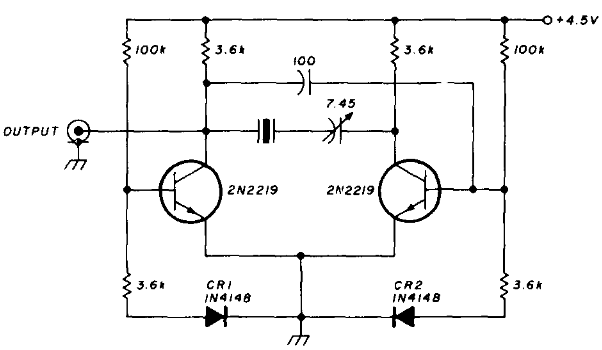
Some practical circuits using fundamental-mode crystals follow. These circuits were chosen to demonstrate a point and should be good for reference. All are believed to be workable although I have not built all of them.



NOTES:

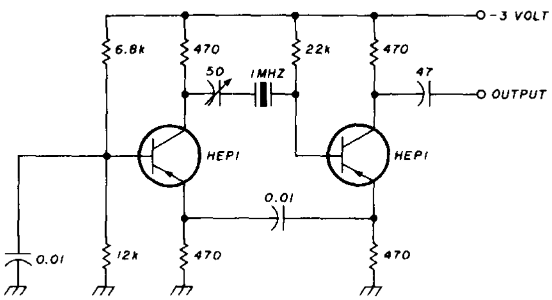
1. Y1 IS H, NT, OR E CUT.
2. C1 IN SERIES WITH THE CRYSTAL MAY BE USED TO ADJUST THE OSCILLATOR OUTPUT FREQUENCY. VALUE MAY RANGE BETWEEN 20pF AND 0.01µF, OR MAY BE A TRIMMER CAPACITOR AND WILL APPROXIMATELY EQUAL THE CRYSTAL LOAD CAPACITANCE.
3. X VALUES ARE APPROXIMATE AND CAN VARY FOR MOST CIRCUITS AND FREQUENCIES; THIS IS ALSO TRUE FOR RESISTANCE VALUES.
4. ADEQUATE POWER SUPPLY DECOUPLING IS REQUIRED; LOCAL DECOUPLING CAPACITORS NEAR THE OSCILLATOR ARE RECOMMENDED.
5. ALL LEADS SHOULD BE EXTREMELY SHORT IN HIGH FREQUENCY CIRCUITS.

Low-frequency oscillator — 10 kHz-150 kHz.



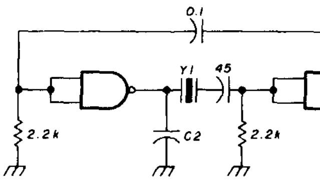
30pF PARALLEL RESONANCE

100-kHz standard oscillator. CR1, CR2 stabilize output.



30pF PARALLEL RESONANCE

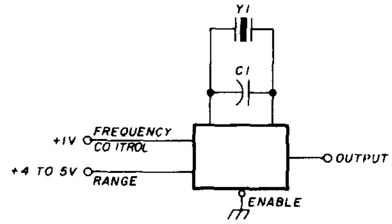
Standard oscillator for 1 MHz.



NOTES:

1. $C2 = \frac{1}{f} \times 10^{-4}$, (f IS IN HZ) PREVENTS SPURIOUS FREQUENCY
2. ICs ARE 7400/7404

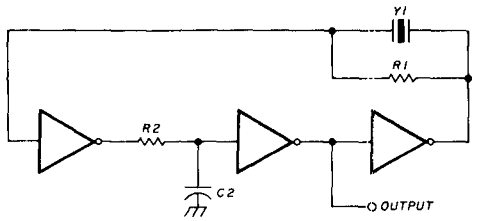
TTL oscillator for 1 MHz-10 MHz.



NOTES:

1. FOR FREQUENCIES ≤ 1 MHZ, C1=5 TO 15pF. FOR FREQUENCIES ≥ 1 MHZ, C1 MAY BE ELIMINATED
2. IC IS SN74S124 FOR f MAX. OF 60 MHZ IC IS SN74LS124 FOR f MAX. OF 35 MHZ

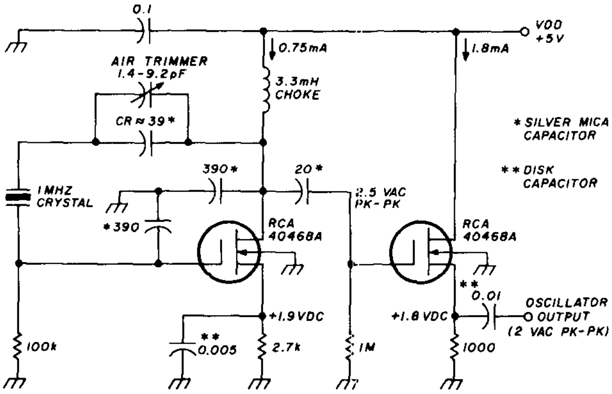
Voltage-controlled oscillator using ICs.



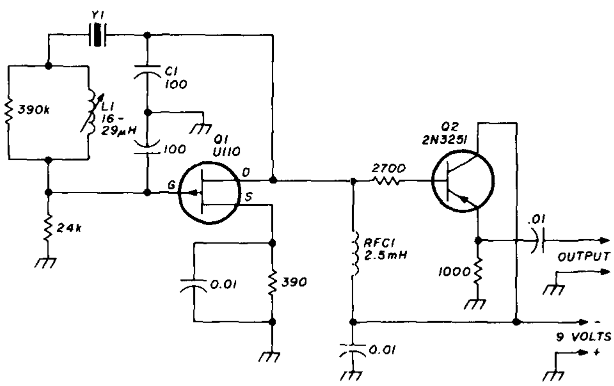
NOTES:

1. 1M < R1 < 5M
2. SELECT R2 AND C2 TO PREVENT SPURIOUS FREQUENCIES
3. ICs ARE 74C04 OR EQUIVALENT

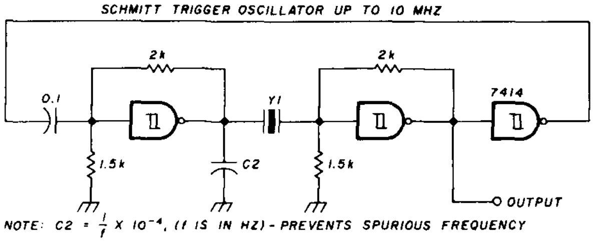
CMOS oscillator — 1 MHz-4 MHz.



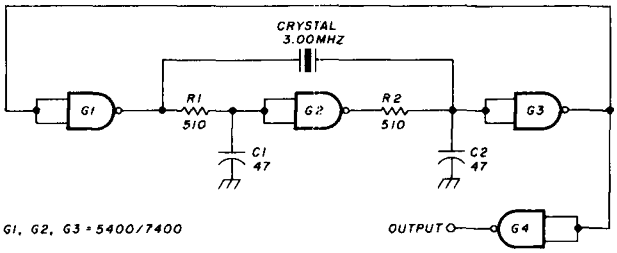
1-MHz fet oscillator and buffer. Circuit exhibits less than 1-Hz frequency change over a V_{DD} range of 3-9 volts. Stability is attributed to mosfets and caps.



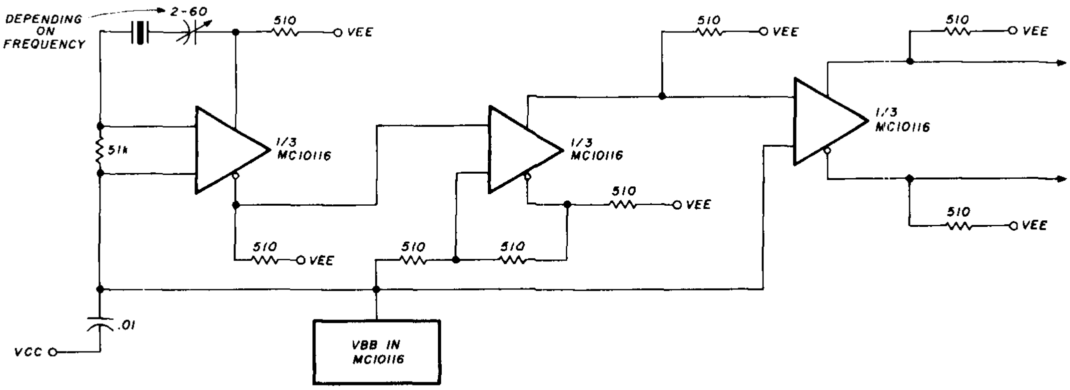
Stable VXO using 6- or 8-MHz crystals uses capacitor and inductor to achieve frequency pulling on either side of series resonance.



Schmitt trigger provides good squaring of output, sometimes eliminating need for an extra output stage.



Crystal-controlled oscillator. This circuit, described in reference 1, oscillates without the crystal. With the crystal in the circuit the frequency will be that of the crystal. Circuit has good starting characteristics even with the poorest crystals.



1-20 MHz oscillator. Circuit operates on fundamental frequency of the crystal selected without a tank circuit. It provides noninverting output. V_{BB} is 1.2 volts, available from the IC; V_{EE} is −5.2 volts. Second section of IC is connected as a Schmitt trigger driving the third section, connected as a buffer, to give good square-wave output suitable for use as a clock driver.

device construction

The natural classification of crystal resonators is according to frequency. The frequency range covered commercially by quartz-crystal units may be taken as a few hundred Hz to over 250 MHz. Use is made of several cuts and patterns of motion (modes). Three common modes of vibration are: flexural, extensional, and shear. **Fig. 5** illustrates these modes and typical mounting techniques.

The designations of certain quartz-crystal vibrators with some of their principal characteristics are summarized in **table 1**. At lower frequencies there are advantages to using one vibrator design over another. Tolerance, activity, and temperature characteristics exemplify the need for choosing. Above 1 MHz most crystals are AT cuts. In general, the choice of cut is that of the manufacturer based on the specification.

table 1. Some quartz crystal vibrators and their principal characteristics.

vibrator designation	usual description	vibration mode	usual frequency range
J	+5° X duplex	flexural	0.2-10 kHz
K	+5° XY bar	flexural	2-16 kHz
H	+5° X plate	flexural	8-100 kHz
N	NT	flexural	8-100 kHz
E	+5° X plate	extensional	40-200 kHz
C	CT	face shear	150-750 kHz
D	DT	face shear	100-500 kHz
G	GT	extensional	90-250 kHz
S	SL	face shear	200-1000 kHz
A	AT	thickness shear (fundamental)	0.8-25 MHz
B	BT	thickness shear (fundamental)	3-40 MHz
A, B	AT or BT	thickness shear (n^{th} overtone; n = 3, 5, 7 etc.)	15-250 MHz

temperature characteristics

Most crystals in amateur service are AT cuts, as our needs are primarily above 1 MHz. A notable exception is the 100-kHz calibrator crystal, which is likely to be an ET cut. Excellent temperature stability and aging are attributed to the AT-cut resonator because of its high Q and cubic temperature curve. In **fig. 6A** a family of AT-cut temperature curves is depicted. The difference between these curves is determined by a change in the angle of cut of the quartz of only a few minutes of arc.

A manufacturer first determines the precise angle that will give the best temperature characteristic commensurate with users' needs. He then determines the crystallographic axes of the quartz and cuts it in the appropriate orientation. Several cuts are illustrated in **fig. 7**.

The temperature characteristics of low-frequency cuts are usually parabolic, as shown in **fig. 6B**. Many of these can be adjusted with respect to the temperature of the turning point; but tolerances are poorer than the AT types. For greatest accuracy in any type of crystal, proportional control ovens, operating at the crystal's turning point, are used.

overtone crystal units

Crystals with frequencies higher than 20 MHz are usually overtone types, although fundamental-type crystals have been made at as high as 35 MHz. Overtone crystals are distinguished from fundamental crystals by their design, which is to operate at an odd harmonic of the crystal basic frequency. It's generally practicable to excite At- and BT-cut plates into third, fifth, seventh, and ninth harmonics of the fundamental frequency; hence a 10-MHz crystal can be vibrated at approximately 30 MHz, 50 MHz, 70 MHz, and 90 MHz. The relationship between overtone and fundamental frequencies is approximately, but never exactly, equal to the integer expressing the harmonic order.

To use the overtone crystal most effectively it's important to know the following characteristics:

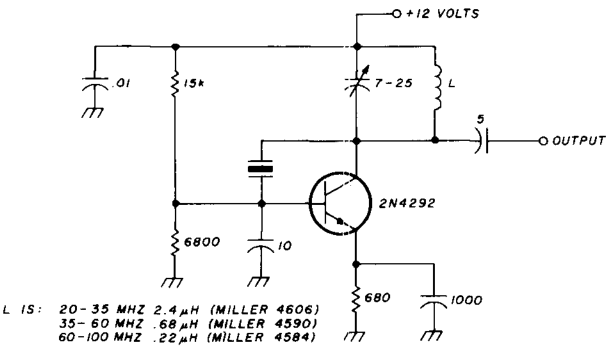
1. A tuned circuit must be used with the crystal to excite it into the desired harmonic mode. If the *Q* of this circuit is too low, improper operation of the crystal will result. The designer must also take care that no other resonances are present that may excite the crystal into another mode.

2. Overtone crystals are designed for operation at series resonance. Because of the narrow bandwidth and low motional capacitance, these crystals are not suitable for fm or for variable-crystal oscillators. Phase-lock operation is practical, however.

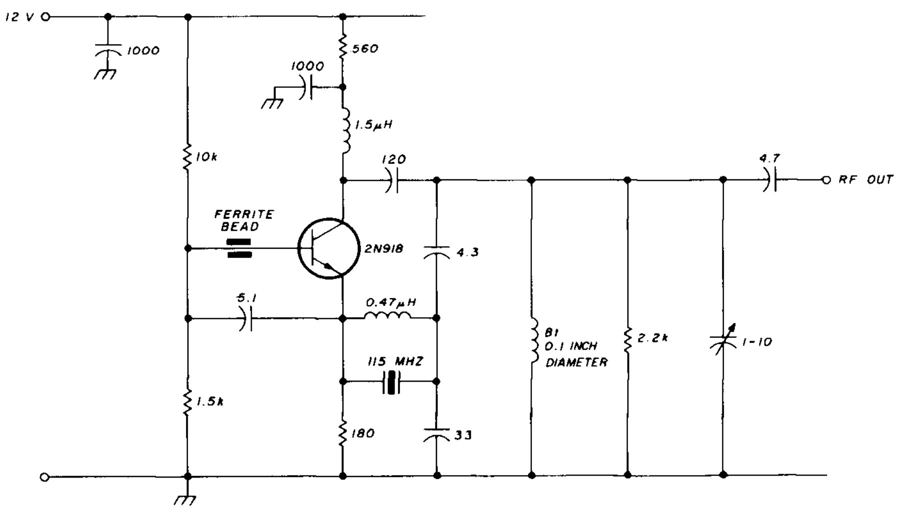
3. The characteristics of a crystal such as temperature coefficient and equivalent resistance apply only to the design frequency. These properties are different for fundamental operation or other harmonic orders.

practical overtone crystal oscillators

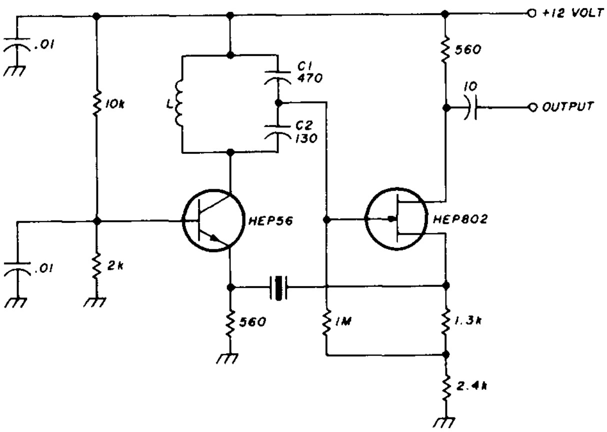
Some useful overtone oscillator designs are shown below, and on the facing page.



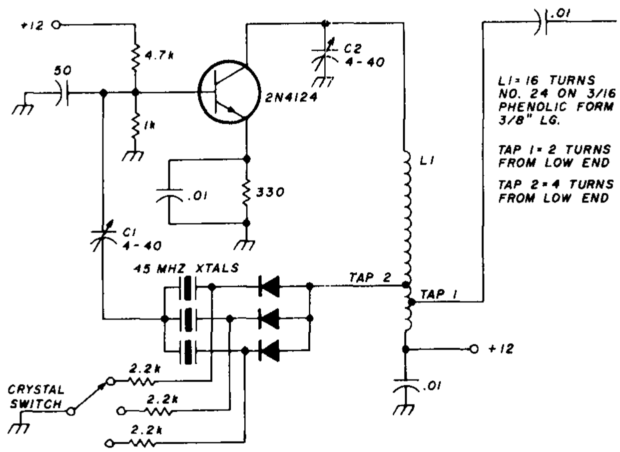
This oscillator is designed for overtone crystals in the 20-100 MHz range operating in the third and fifth mode. Operating frequency is determined by the tuned circuit.



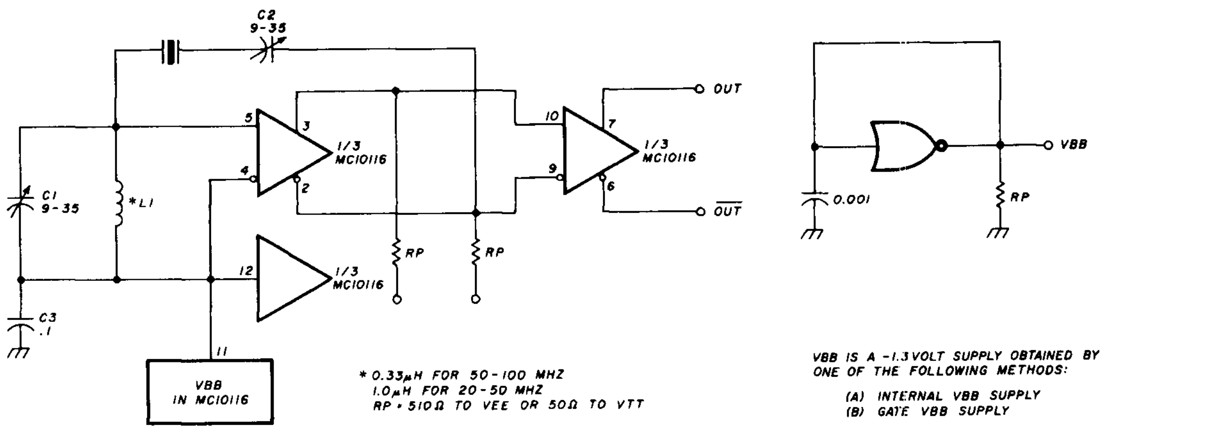
Design for high reliability over wide temperature range using fifth and seventh overtone crystals. Inductor in parallel with crystal causes antiresonance of crystal C_0 to minimize loading. Technique is commonly used with overtone crystals.



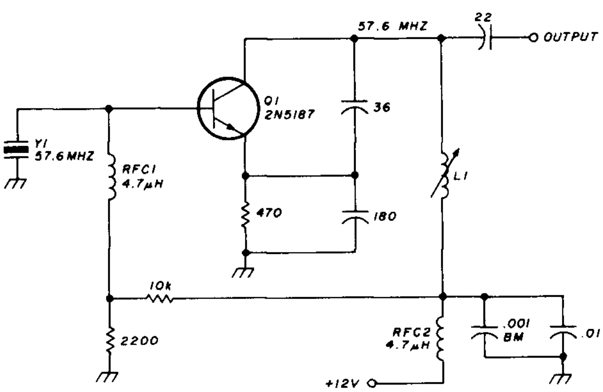
Typical Butler oscillator (20-100 MHz). An fet should be used in the second stage; circuit is not reliable with two bipolars. Sometimes two fets are used. Frequency is determined by LC values.



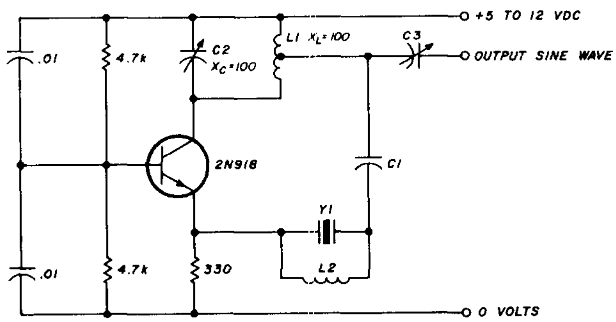
Overtone oscillator with crystal switching. Similar circuits electronically switch the crystals. The large inductive phase shift of L1 is compensated for by C1. Overtone crystals have very narrow bandwidth, therefore the trimmer has a smaller effect than for fundamental-mode operation.



Overtone oscillator using Motorola MECL devices. Frequency range is 20 MHz-100 MHz, depending on crystal frequency and tank-circuit tuning. The tank, C1, L2, is tuned to select the proper overtone mode. C2 compensates phase shift of the IC. More details are given in reference 2.



Fifth-overtone oscillator isolates the crystal from the dc base supply with an rf choke for better starting characteristics.



NOTES:

1. Y1 IS AT CUT OVERTONE CRYSTAL.
2. TUNE L1 AND C2 TO OPERATING FREQUENCY.
3. L2 AND SHUNT CAPACITANCE, CO, OF CRYSTAL (APPROXIMATELY 6pF) SHOULD RESONATE TO OSCILLATOR OUTPUT FREQUENCY (L2 = .5µH AT 90 MHZ). THIS IS NECESSARY TO TUNE OUT EFFECT OF CO.
4. C3 IS VARIED TO MATCH OUTPUT.

50 MHz-150 MHz overtone oscillator uses a 2N918.

effects of drive level

The level of drive imposed on an oscillator crystal is usually specified in terms of the power dissipated in it. Ideally, the crystal oscillator should be regarded as a source of stable frequency, but in practice it must also be considered as a source of power.

Changes in drive level will affect the resonator frequency; therefore, it's important for the manufacturer to know the drive level of the oscillator circuit for calibration of the crystal. Crystals operated at high drive levels will become unstable, sometimes jumping frequency into a spurious mode. Excessive resonator heating may cause a permanent shift in frequency or possibly facture the quartz. The NT resonator is particularly vulnerable to fracturing. A good rule is to operate the crystal at the lowest drive level compatible with good starting characteristics.

The old WWII surplus pressure-type crystals use larger pieces of quartz than their modern counterparts. As might be expected, these can withstand higher drive levels. You may also find that pressure-type crystals can be pulled in frequency more easily than modern units because their motional capacitance, $C1$, is higher.

aging

Like mountain dew, most crystals improve with time. Just after the crystal is manufactured, there are stresses, which when relieved, change the crystal frequency. Most manufacturers age the crystal by temperature cycling or high-temperature aging until the worst changes have occurred. You'll then experience slower drift. In many applications, the drift is negligible but is present. The most stable crystals are those in the 4-5 Mhz range.

Aging can be positive or negative, depending on which factors are present in a particular unit. Migration of small particles within the crystal holder is usually blamed for frequency changes. If these dirt particles land on the crystal, its frequency decreases. These particles are present despite the most rigorous cleaning procedures. Metal-cased, gas-filled crystal units will usually age negatively. Some crystals are evacuated rather than gas filled. These units are cleaner and have better aging characteristics but lower drive level ratings. Evacuated units may age higher in frequency because some of the plating is vaporized. You won't find these crystals on the surplus market; they are mentioned here as a point of interest.

tips on using crystals

The great enemy of quartz is drift. Old pressure types have been known to fail because of particles from the rubber gasket, which may have deteriorated. Careful cleaning with alcohol or similar solvent will bring these crystals back to life. The same procedure will probably increase the frequency of a unit that hasn't failed. Most certainly it will increase crystal activity.

This trick isn't practical with solder-seal holders, but then these units are much more reliable. Don't open the holders on the solder seal units or you'll find that the frequency has changed. This is because of a change in pressure and of gases surrounding the quartz element. Besides the frequency change, reliability is compromised by the introduction of dirt.

Sometimes an oscillator crystal is used in a filter application, but performance will not always be satisfactory. Special designs are used for filter crystals. These crystals have lower activity and are virtually free of spurs (unwanted modes). In oscillator service, the presence of unwanted modes is not as critical as in filter service, where broadband energy will excite all modes.

There are many uninvestigated facets of the quartz crystal. While some people still claim crystal manufacturing is akin to witchcraft, this is just not so. A few years ago, natural quartz, which was mined in Brazil, was used for all U.S. crystals. Synthetic quartz made in the U.S. has been improved to the point where it's now used in all but the most critical applications.

Natural quartz still has higher Q. Synthetic quartz has the advantage of perfect crystalline structure and uniform size. The use of natural quartz incurs much waste — rarely is there a fully perfect crystal, and small structures may not be practical for cutting. Crystallography is certainly a science not fully investigated but one we should study.

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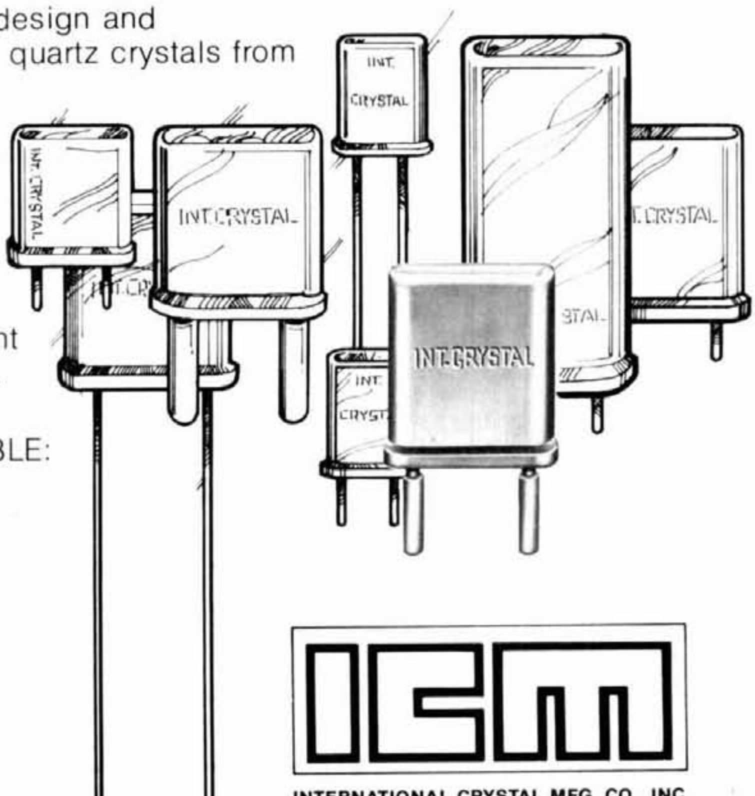

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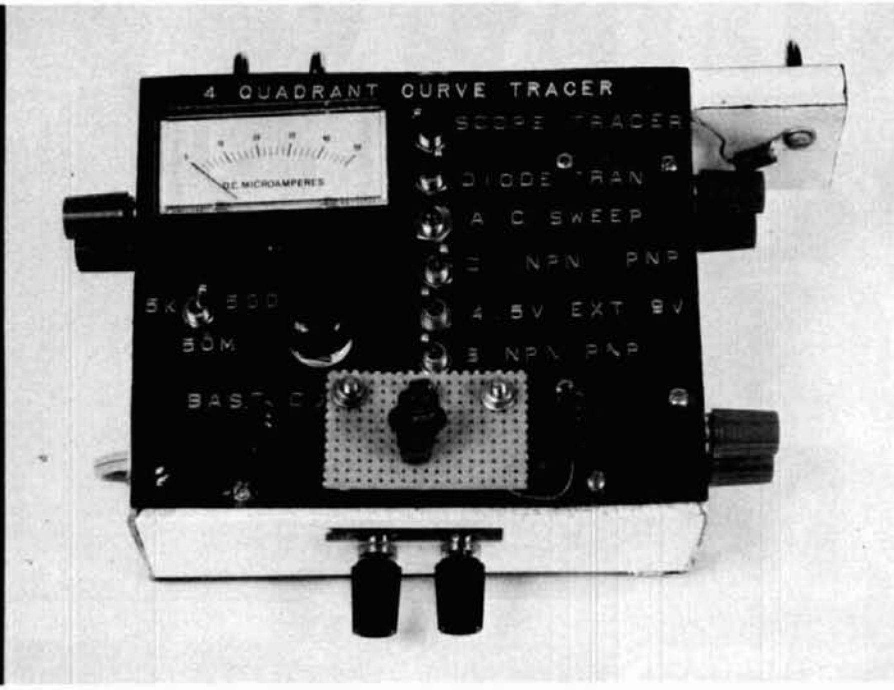


four-quadrant semiconductor curve tracer/analyzer

It's not just a transistor tester — it's a versatile instrument that can be used for checking and designing electronic circuits under static and dynamic conditions

The test equipment used in building and testing electronic circuits is still one of the most interesting parts of Amateur Radio. With the increased use of semiconductors, much more data is needed for their use and replacement than can be obtained from simple transistor testers. This is why I felt it necessary to build the instrument described here.

features

The semiconductor curve tracer/analyzer is as versatile as your imagination yet is economical and simple to build. It can be used for checking as well as designing electronic circuits under both static and dynamic conditions. It can also be used to determine parameters of signal and power transistors, unijunction transistors, field-effect transistors, silicon-controlled rectifiers, and triacs.

Most diodes can be analyzed, including signal and power devices, zeners, protection diodes, bias diodes, point-contact diodes, hot-carrier diodes, and light-emitting diodes.

By Stuart Tuma, W1QXS, 17 Briggs Street, Melrose, Massachusetts 02176

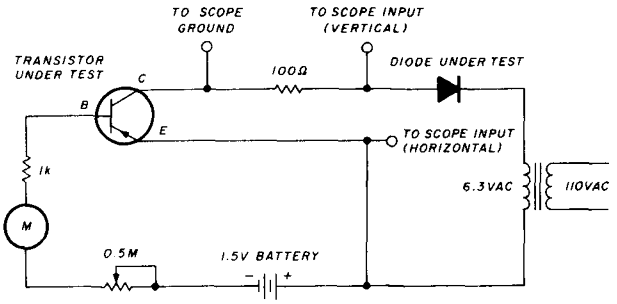


fig. 1. Basic circuit of the curve-tracer/analyzer.

Another feature of the instrument is that of checking photocells made of cadmium sulfide or cadmium selenide. With appropriate adapters, the instrument can also be used to check integrated circuits.

The analyzer is not restricted only to seminconductor devices. It can also be used to check the piezoelectric effect of quartz crystals under various circuit conditions and to check the design of amplifier as well as oscillator circuits. In electrical circuits, the analyzer can be used to check the sensitivity and internal resistance of D'Arsonval meters and galvanometers as well as the sensitivity of relays, including the popular reed relay. With proper adapters, low-power vacuum tubes can also be checked.

theory of operation

As a transistor curve tracer, the unit is designed so that the oscilloscope vertical input measures the voltage across a 100-ohm resistor to ground, which is used to measure collector current. The oscilloscope,

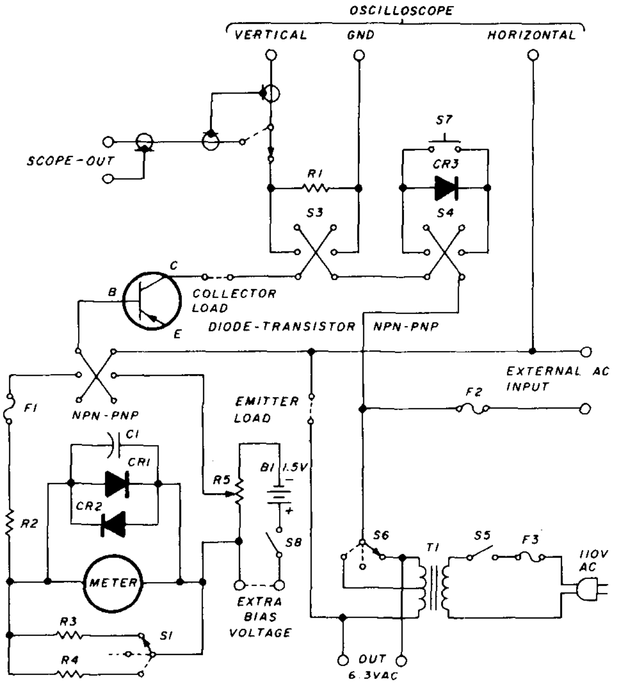


fig. 2. Schematic of the four-quadrant curve-tracer/analyzer.

having a vertical sensitivity of 0.1 volt per division across 100 ohms, gives 1 mA per division. Should you desire to increase the current per division, you can use 1 volt per division, thus giving 10 mA per division, and so on. **Fig. 1** shows the basic circuit.

The oscilloscope horizontal deflection is used to measure collector-emitter voltage. The oscilloscope is calibrated to read 1 volt per division. Since the horizontal amplifier input is not directly calibrated, it will be necessary to use the sweep voltage, which is approximately 9 volts peak pulsating direct current. This gives a value of 1 volt per division horizontal de-

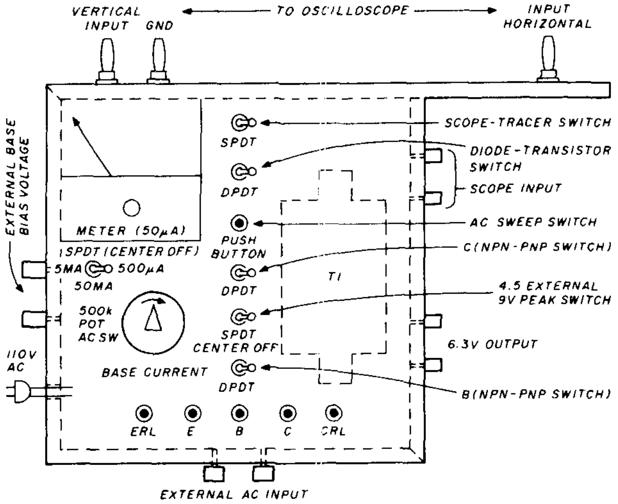


fig. 3. Sketch of front panel showing parts layout.

flection for 9 divisions. Now we have a method of checking both voltage and current from our 60-Hz sweep signal supplied by the 6.3 VAC source.

The emitter-base circuit has a separate supply — a 1.5-volt battery for the base bias. Higher bias voltage can be added to the emitter and collector circuit if desired. The base-current circuit employs a 50-microampere meter movement (which has an internal resistance of 1000 ohms). This circuit will read 50 microamperes (no shunt), 0.55 milliamperes with a 100-ohm shunt, and 5.05 milliamperes with a 10-ohm shunt.

The meter is protected by two silicon diodes (**fig. 2**), which I found to have a forward-bias-voltage drop of 0.4 volt. The voltage drop across a 50-microampere meter, full scale, having a resistance of 1000 ohms, should be 0.05 volt. This gives good protection for the meter.

Most silicon diodes have about 0.6 volt forward bias, so other types of diodes could be used. However, check the diode's forward-bias voltage before you install it. You can do this by connecting the diode in series with a 1000-ohm resistor and a 1.5-volt battery. Measure the voltage drop across the

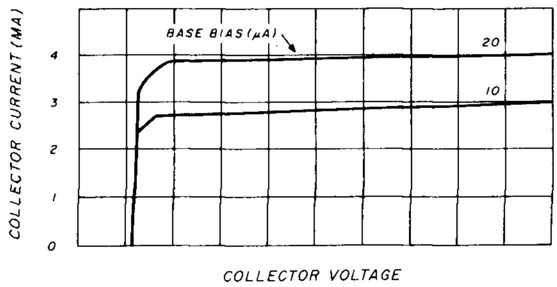


fig. 4. Typical curve for a pnp transistor (type 2N252) showing the relationship between base bias and collector current and voltage.

diode when the diode is conducting. This voltage should be 0.6 volt or less.

A 0.001-μF capacitor is connected across the meter for still more protection from rf coming through the line or from some other external source. The meter circuit is further protected by a 1/4-ampere fuse.

The base-emitter current of the transistor under test is controlled by a 10-k pot (R5, **fig. 2**) and a voltage divider (R3, R4) in series with the 1000-ohm resistor (R2) and the meter. This gives a maximum current of about 1.5 mA.

construction

A schematic of the curve tracer/analyzer is shown in **fig. 2**. A parts list is given in **table 1**. The sketch of **fig. 3** shows parts layout on the front panel.

wiring. Wiring was easy. I used a pencil-type soldering iron, a good grade of solder, and a clean, tinned soldering tip. The circuit was connected with 0.4-mm (no. 26 AWG) hookup wire. The reversing (inverting) switches were wired first. External connecting leads were added for future wiring. All components were mounted on the top panel and wired as shown in **fig. 2**. The extended leads were then soldered to the proper binding posts. I used one strand of 0.08-mm (AWG no. 40) wire for the 1-ampere fuse on the external binding post. I used shielded wire for the oscilloscope tracer switch and scope output.

The five terminal posts on the top front panel were made for plug-in adapters on which you can mount various "TO" sockets for transistors or a module-type socket for testing other devices. Note that different load resistances can be added to the emitter circuit as well as to the collector circuit. This allows you to build prototype circuits before putting them into a breadboard circuit. If desired, a solderless breadboard adapter can be used.

mechanical details. The top and bottom panels were made of two pieces of plastic 152 mm (6 inches) square. The sides were made from 6.4-mm (1/4-inch) plastic channel molding, 51 mm (2 inches) wide, obtained from a local lumberyard dealer. These pieces were cut to form the four sides. I used small metal screws to put it together. Parts were laid out in a convenient order. (Mark or etch parts locations on the top and side panels.)

The top panel required a 38-mm (1-1/2-inch) diameter hole to mount the meter. I found that an old pencil soldering iron was just the thing for this, since the plastic melts at a very low temperature. I used a pipe reamer for the finishing touches. I made a 9.5-mm (3/8-inch) hole for the potentiometer. I used a

table 1. Curve-tracer/analyzer parts list.

component	description	approximate cost	source
B1	1.5 volt battery type AA	$0.20	Lafayette Radio
banana plugs	(screw mounting) M3.5 (6-32) (pkg of 10)	3.40	Lafayette Radio
binding posts	5 way (pkg of 6)	1.69	Lafayette Radio
C1	.001 μF 1000V ceramic	0.15	Lafayette Radio
CR1, CR2, CR3	1A 600 PIV silicon diodes (pkg of 3)	1.19	Lafayette Radio
F1, F3	1/4-A 3AG fuses (pkg of 5)	1.05	Lafayette Radio
F2	see text		
meter	50 μA (99PS1146V)	6.95	Lafayette Radio
R1	100 ohm 1W 10% composition	0.20	Lafayette Radio
R2	1000 ohm 1/2W 10% composition	0.15	Lafayette Radio
R3	100 ohm 1/2W 10% composition	0.15	Lafayette Radio
R4	10 ohm 1/2W 10% composition	0.15	Lafayette Radio
R5, S5, S8	500-k pot with two SPST switches	2.09	Lafayette Radio
S1, S6	SPDT 3A 125V mini toggle switches (center off)	1.39 ea.	Poly Paks
S2, S3, S4	DPDT 3A 125V mini toggle switches	1.95 ea.	Poly Paks
S7	SPST momentary mini switch	0.79	Lafayette Radio
T1	6.3V ct 1A or equivalent	3.75	Lafayette Radio
		$25.25	

smaller reamer for the finishing touches. I used the same reamer for the 6.4-mm (1/4-inch) holes for mounting the switches.

I painted the inside of the top and bottom panels flat black. I used 5-mm (3/16-inch) holes to mount the binding-post terminals. The terminal posts were spaced 19 mm (3/4 inch) apart. I constructed the instrument so it would plug into my Conard oscilloscope. However, with proper external leads, this analyzer should fit into any standard oscilloscope.

There are probably a thousand and one uses for this instrument. I've listed only a few, but enough so you'll become familiar with its use, both as a curve tracer and analyzer. I'm sure you'll find other uses, and I'd like to hear from you in this regard.

testing transistors

Plug the curve tracer into the proper oscilloscope inputs. The vertical output goes to the oscilloscope vertical input, and the horizontal output goes to the oscilloscope horizontal input. Ground the curve tracer to the oscilloscope ground.

pnp transistors. Set up the oscilloscope as follows:

quantity	setting	measurement
vertical gain	0.1 V/division	1 mA/division
horizontal gain	9 divisions (with transistor in circuit)	approximately 1 V/division (peak)
horizontal sweep source	external	
intensity	normal	
focus	normal	

Set up the curve tracer as follows (see **fig. 3**):

switch positions	switch to
scope-tracer	tracer
diode-trans	trans
AC sweep	pushbutton released
C NPN-PNP	PNP
B NPN-PNP	PNP
4-1/2V-EXT-9V	9V (peak)

Switch meter to 50 microamperes or to a convenient current rating. Set **BASE CURRENT** counter clockwise (CCW). Be sure that jumpers are connected between emitter-to-emitter-load and between collector-to-collector load terminal posts (**fig. 2**).

Connect the test transistor to the proper input terminals or into a transistor plug-in adapter. Turn the **BASE CURRENT** control until the current begins to increase. Note the variation of the trace. Increase the trace two or three divisions and note the increase in the base-bias current (record this reading). Increase the collector current until one more division is obtained. Again, record the base-bias reading. From the data taken, the transistor beta and alpha gain can be determined as shown below.

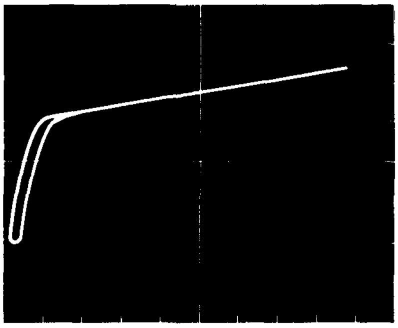
The beta gain is determined by taking the variation of the collector current and dividing it by the variation of the base bias current. Example:

beta gain $=$ *Δcollector current/ Δbase bias current*
beta gain $=$ *Δ1 mA/ Δ10 microamperes* $= 100$

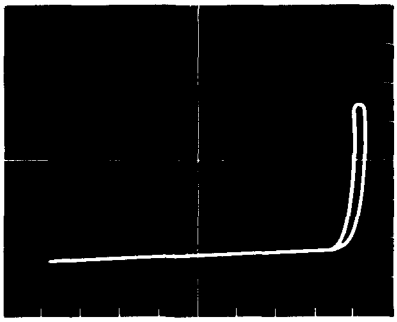
The alpha gain is equal to the beta gain divided by the beta gain plus 1. Example:

alpha gain $=$ *beta gain/beta gain* $+ 1$
alpha gain $= 100/100 + 1 = 0.99$

Fig. 4 shows the relationship of collector current, collector voltage, and base bias for a typical pnp transistor (type 2N252).



NPN transistors. The setup for npn transistors is the same as for pnp transistors, except that the emitter and base **NPN** and **PNP** switches are both set to the **NPN** position. The curve on the oscilloscope may have to be recentered. The beta gain can be determined as with a pnp, except that the curve will be in the opposite direction as shown below.



testing diodes

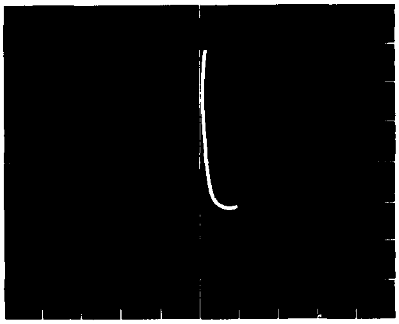
Set up the oscilloscope as follows:

quantity	setting	measurement
vertical gain	0.1 V/division	1 mA/division
horizontal gain	9 divisions (with diode in circuit)	approximately 1 V/division (peak)
horizontal sweep source	external	

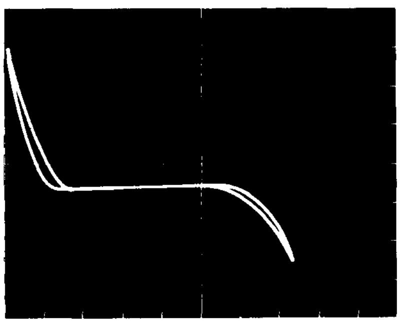
Set up the curve tracer as follows:

switch position	switch to
scope-tracer	tracer
diodetrans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
B NPN-PNP	not in circuit
4-1/2V-EXT-9V	9V (peak)
base-current meter shunt	not in circuit
base-current control	not in circuit

Connect the diode cathode to terminal post **E** and the diode anode to terminal post **C**. Note the L-shaped pattern. To give better diode action press AC SWEEP. This shows when the diode is not conducting.



For checking 5-volt zeners, install a 3900-ohm resistor in series with the emitter load and emitter terminal post. This allows about 10 milliamperes of current flow through the zener. (For checking other types of zeners, a different load resistor must be employed.) Connect the diode cathode to the emitter terminal and the diode anode to the collector terminal post. The oscilloscope and curve analyzer are set up exactly as in the diode setup. For higher-voltage zeners, **EXT** input can be used with a Variac or variable AC voltage source.



testing photocells

Photocells made from cadmium sulfide or cadmium selenide working on the principle of photo conductivity can be checked with the analyzer by using the base-bias circuit. The photocell has only two leads and can be attached to **E** and **B** on the curve analyzer. As the light increases on the photocell, the circuit conductivity also increases. Using the circuit as shown, the conductivity of the photo cell can be obtained.

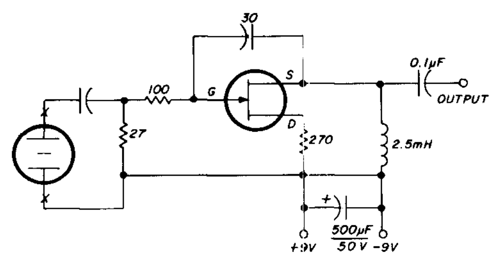
This circuit can also be used as a photographic light meter.

testing D'Arsonval meters

This circuit can measure characteristics of meters with a sensitivity of 10 microamperes to 5 milliamperes full scale. This is done by connecting the meter under test to terminal posts **E** and **B** (**fig. 3**). Increase the current flow with the **BASE CURRENT** control until the meter reads full scale. Note the sensitivity of the unknown meter in amperes. To determine the internal resistance, connect a decade box or a 500-ohm pot across the meter under test. When the shunt resistance decreases the full-scale reading to one-half scale, measure the resistance of the 500-ohm pot to the center arm . This should give the meter's internal resistance.

crystal-oscillator checker

The crystal-oscillator checker uses an N-type junction fet (RS2035) field-effect transistor, which can be obtained for about one dollar. An equivalent type could be used. The module is wired as shown. It is installed into the curve tracer across the **E** and **C** terminals. Using a 50-μF filter across the supply input to the module as shown, the dc input is about 9 volts. The circuit works very well for checking quartz crystals as low as 100 kHz to as high as 15 MHz.



Set the oscilloscope as follows:

quantity	measurement
Vertical gain	1V/division
Horizontal sweep	1 mS/division
Horizontal mode	internal

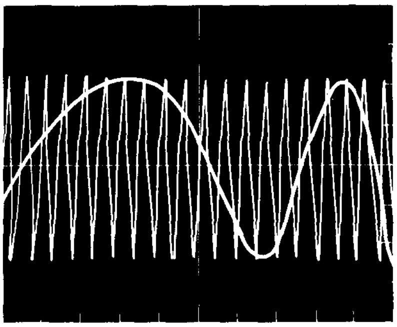
Set all other controls to normal position.

Set the curve tracer as follows:

switch position	switch to
scope-tracer	scope
diode-trans	not in circuit
AC sweep	pushbutton released
C NPN-PNP	not in circuit

switch position	switch to:
B NPN-PNP	not in circuit
4-1/2V-EXT-9V	9V (peak)

Connect the crystal to crystal terminal on the module. The module output should be connected to the



oscilloscope input. The waveform can be observed on the oscilloscope. For a more accurate frequency reading, a frequency counter should be used.

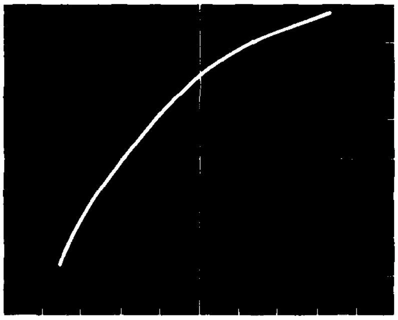
some other uses

The curve tracer/analyzer can be used for checking other devices. Presented below are the results of some tests I've run on junction fets, unijunction transistors, silicon-controlled rectifiers, and triacs. The setup instructions for the scope and analyzer are as in the previous examples.

junction fets

oscilloscope:

quantity	position	measurement
vertical	1V/division	1 mA/division
horizontal	1V/division	9 divisions
horizontal sweep	external	

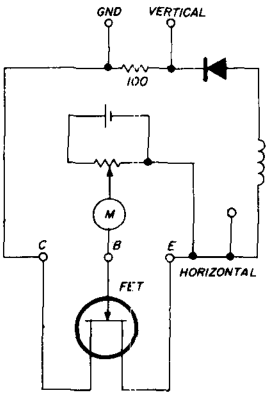


curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
4-1/2V EXT 9V	9V

switch position	switch to:
base-current	5 mA
meter shunt	
base control	adjust to proper gate voltage *vs* source current

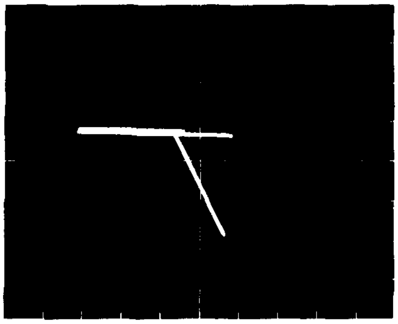
Connect fet as shown below. Use external voltmeter to measure gate voltage *vs* source current.



unijunction transistors

oscilloscope:

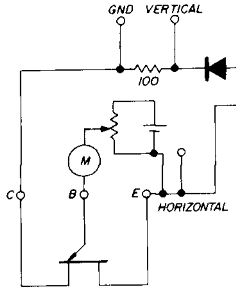
quantity	position	measurement
vertical	0.1V/division	1 mA/division
horizontal	1V/division	4-1/2 divisions
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton released
C NPN-PNP	NPN
4-1/2V EXT 9V	4-1/2V
B NPN-PNP	NPN
base-current	5 mA
meter shunt	
base control	adjust for gate and B1 voltage with external meter

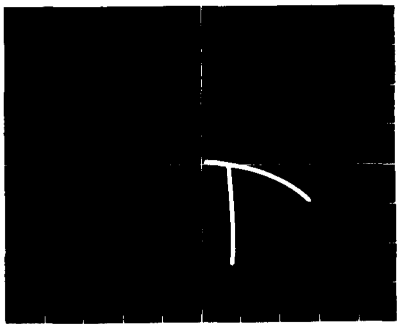
Connect unijunction device to curve tracer as shown using a 0-5-volt dc meter. Measure trigger voltage between gate and B1.



silicon-controlled rectifiers

oscilloscope:

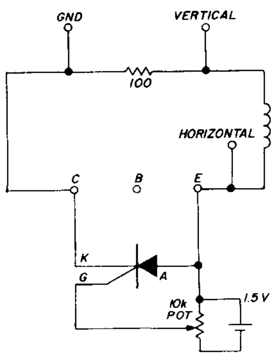
quantity	position	measurement
vertical	1V/division	10 mA/division
horizontal	3V/division	9V
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton (AC)
C NPN-PNP	PNP
B NPN-PNP	not in circuit
4-1/2V EXT 9V	9V

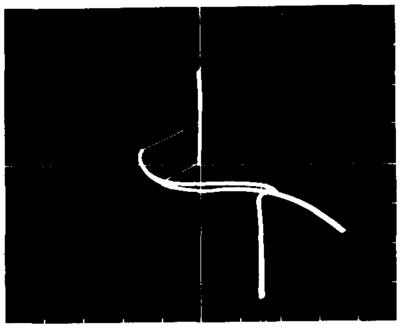
Connect scr as shown in circuit below; anode to (E) and cathode to (C). Use a 1.5-volt battery in series with a 10k-ohm variable resistor as a variable voltage source for gate and anode. With a separate voltmeter measure the gate-anode trigger voltage. Note current flow through the scr.



triac-controlled rectifiers

oscilloscope:

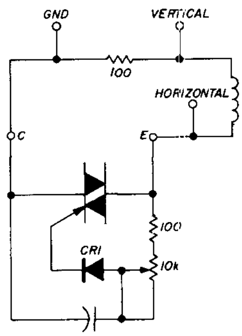
quantity	position	measurement
vertical	1V/division	10 mA/division
horizontal	3V/division	9V
horizontal sweep	external	



curve tracer:

switch position	switch to:
scope-tracer	tracer
diode-trans	diode
AC sweep	pushbutton (AC)
C NPN-PNP	PNP
B NPN-PNP	not in circuit
4-1/2V EXT 9V	9V

Connect as shown in circuit below using external R-C network. Measure gate voltage between emitter terminal and gate. Note ac current flow.



final remarks

I've presented the results of my work in trying to improve the lot of the home builder who likes to work with semiconductors. No doubt you'll come up with other uses for the basic instrument, and I'd like to hear from you. If you have any suggestions or questions, please send them to me in a self-addressed, stamped envelope, and I'll be glad to reply.

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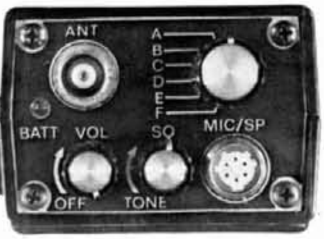

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the ultimate noise blanker

Conventional noise-blanker designs overlook the effectiveness of the blanking switch — here's a new approach to the problem using fm techniques

Noise blankers are commonly incorporated into hf communications receivers but use less than a perfect switch and switch-control timing. This article presents a new concept for a noiseless i-f switch that allows a very effective impulse noise blanker to be constructed. The fundamental concept of noise blankers is also reviewed.

conventional noise-blanker design

All noise blankers operate on the principle that a separate noise receiver listens in a no-signal portion of the spectrum (typically 30-35 MHz) especially for the purpose of receiving noise pulses. The detected pulses are processed and applied to a blanking switch in the communications receiver i-f strip to momentarily interrupt the signal path during the noise pulse period.

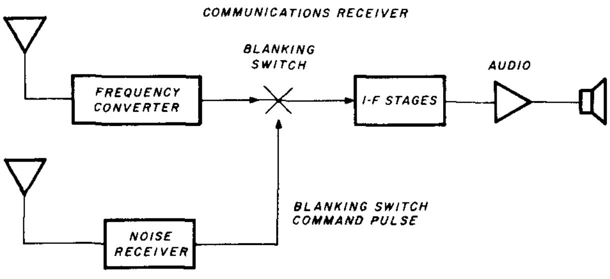

fig. 1. Block diagram of a typical noise blanker used in communications receivers.

design considerations

Typically, the noise pulse has a duration of only a few microseconds. If allowed to pass through the communications receiver, several factors cause the pulse to be stretched, including narrow filters and saturation. If the blanker operates properly, the noise pulse is removed with negligible effect on the communications signal.

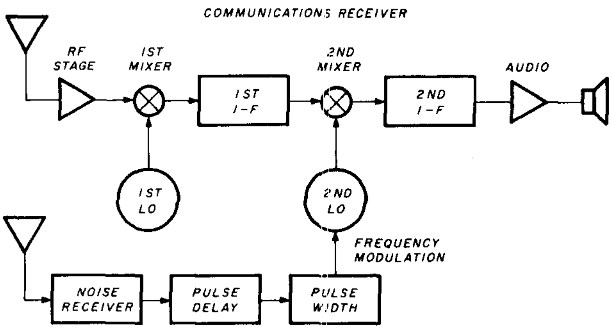

fig. 2. Block diagram of an improved noise-blanking system using a frequency-shift network ahead of the second local oscillator in a dual-conversion receiver.

The major limitation in previous designs has been the effectiveness of the blanking switch itself. Most have a limited on/off ratio and introduce switch-transient noise as well. A typical blanking system block diagram is shown in **fig. 1**.

For proper operation, timing is an important factor. The blanking switch should open the signal path for the pulse period only. If the switch is open too long, unnecessary distortion of the signal occurs. If the switch is not open during the complete pulse period, a portion of the noise pulse will leak through.

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This aspect does not receive proper attention in typical designs.

The design of the noise receiver must include adequate bandwidth to ensure proper processing of narrow pulses to enable accurate control of the blanking switch. Also, the time delay through the noise receiver must be sufficiently shorter than the delay through the communications receiver front end to enable pulse shaping and control before application of the control signal to the blanking switch. Delay through the receiver is a function of the i-f bandwidth. In practice, it has been found that the use of standard 10.7 MHz fm i-f transformers in the noise receiver is the best choice. Adequate delay in the communications receiver (ahead of the switch) is normally realized in the first i-f section.

Optimum design of a communications receiver suggests a multipole narrow bandwidth i-f filter immediately following the first mixer. Conversely, a multipole filter has significant time delay and stretches the noise pulse width ahead of the blanking switch. Although this is undesirable from a purely technical viewpoint, the effect is negligible and can be ignored.

If the blanking is located in the receiver after a substantial amount of gain, saturation of the i-f amplifier may result (especially under weak-signal conditions

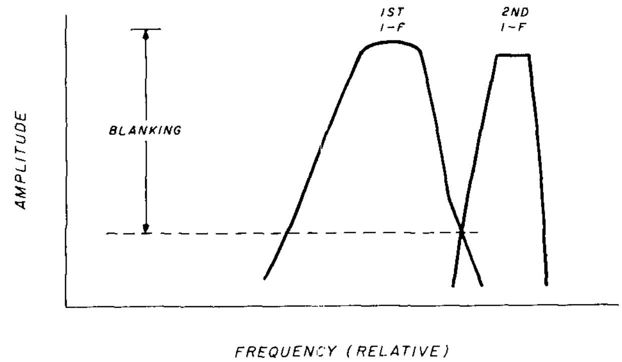

fig. 3. Blanking-switch isolation can be achieved by misaligning the conversion frequency during the blanking period. The amount of frequency shift can be determined by comparing the shape factors of the first and second i-f filters.

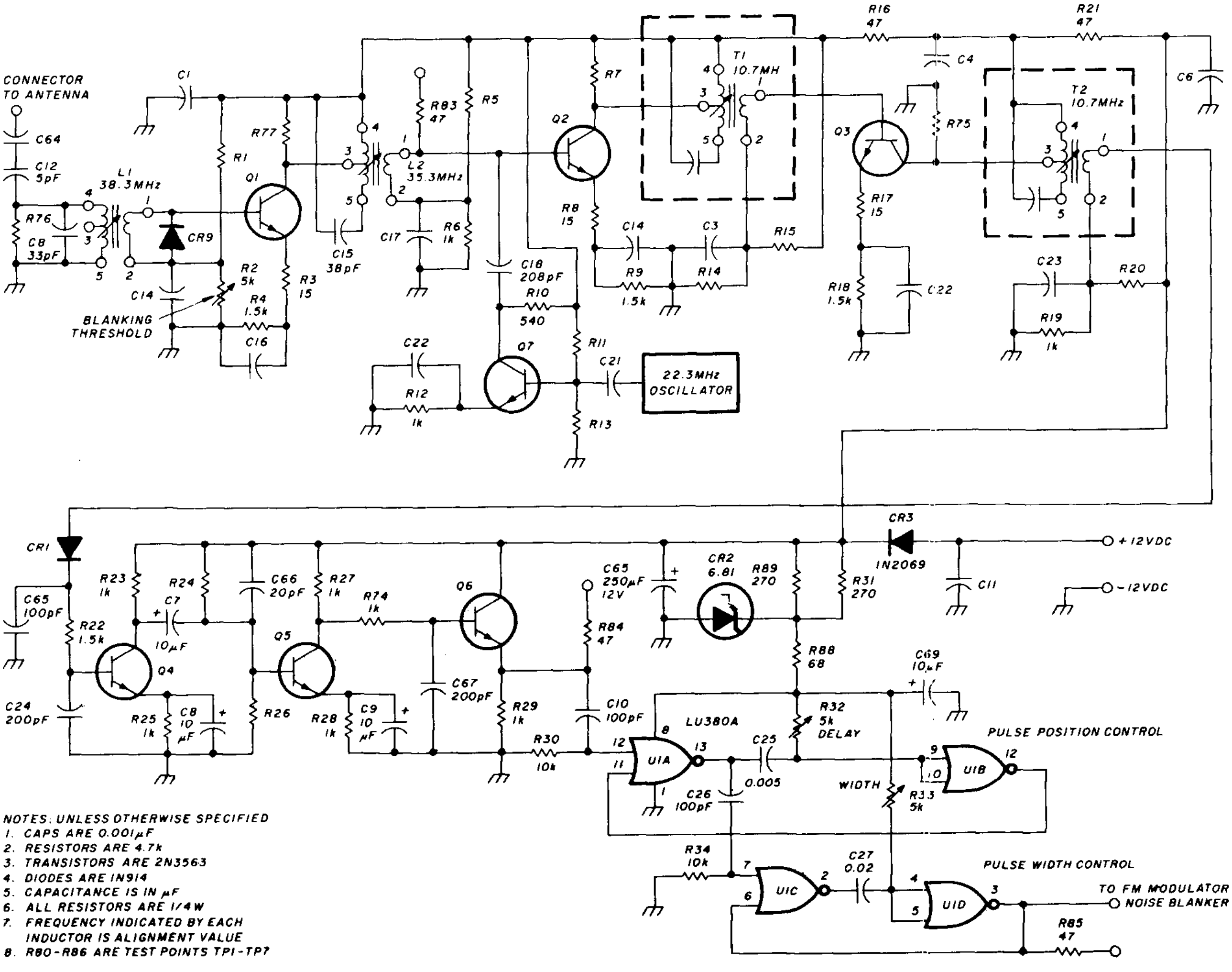

fig. 4. A practical noise-receiver schematic using the technique described for taming switch action.

with no agc). Slow recovery is normally associated with saturation, hence extreme amounts of effective pulse stretching may occur. However, typical i-f blanking switches must operate at relatively high signal levels, since the noise introduced by the switch is proportional to the signal level at which they operate. If the blanking switch is located immediately after the first i-f filter, signal levels will typically be in the microvolt region.

a new switch

Switching transients can be eliminated by using a frequency modulation technique. With reference to **fig. 2**, note that when a command signal from the noise receiver is applied to a frequency-shift network associated with the second local oscillator in a dual-conversion communications receiver, the mixer output will be at a frequency other than that required to transfer the signal from the first i-f to the second i-f. In other words, if the time of the command signal occurs at exactly the time the noise pulse propagates through the first i-f and is applied to the mixer, the noise pulse energy will leave the mixer at a frequency other than that of the second i-f.

Since there is no amplitude noise (switching-transient noise) associated with the frequency shift, no switching noise due to the blanking action will occur when the signal path is momentarily interrupted.

The amount of required frequency shift may be determined by comparing the shape of the first and second i-f filters (see **fig. 3**) and misaligning the conversion frequency during the blanking period to achieve the desired switch isolation. Typically, a few kHz will be adequate. Normally, the second local oscillator can be frequency modulated by tens of kHz with no adverse effects.

In practice, the addition of a small capacitor and switching diode to a receiver second local oscillator implements the desired switch. If the second local oscillator is crystal controlled, it must be replaced with an oscillator that can be frequency modulated.

results

Performance of the fm switch blanking system is astonishing, compared with conventional systems. An hf mobile receiver operating at a busy intersection had no noticeable ignition noise interference with the switch operating. With the switch disabled, communications was impossible.

A complete schematic of a practical noise receiver is included in **fig. 4**. For a particular application, the pulse delay network may require component value changes depending on the delay in the first i-f filter of the communications receiver.

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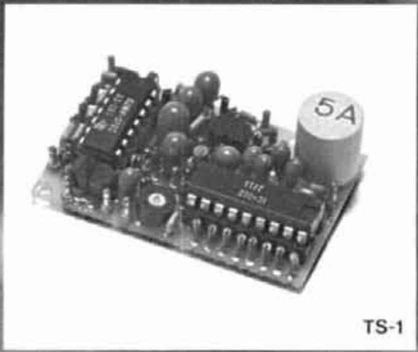
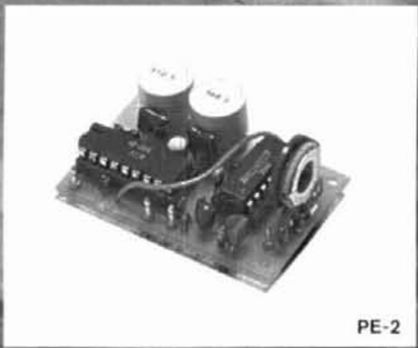

TS-1


TS-1JR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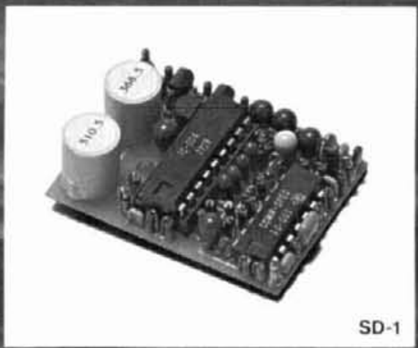
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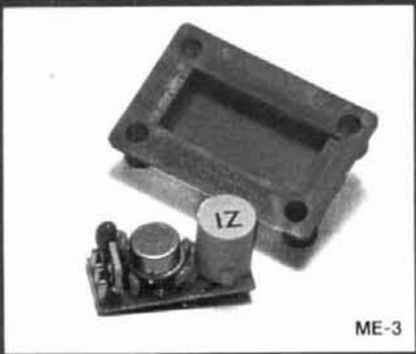
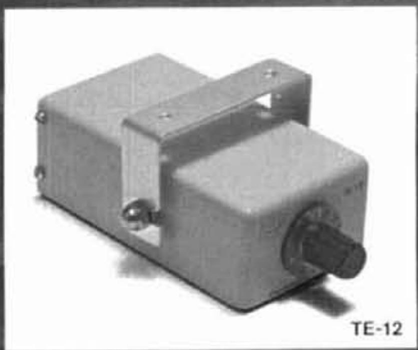
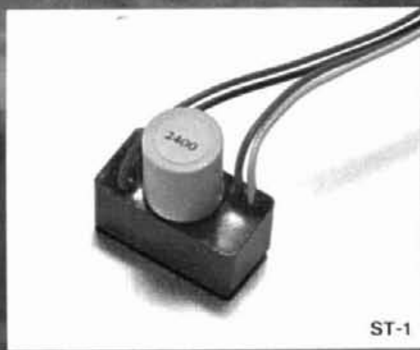


ME-3

TE-8



TE-12



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power-line noise — the cause and cure

Discussion of the steps necessary to locate power-line noise, starting with in-house noise sources and ending with utility-pole-generated noise

Power line noise can be one of the most exasperating forms of irritation experienced by amateurs living in or near metropolitan areas. This problem, which can drive active operators beyond the point of sanity in record time, is characterized by a long-term arcing-type sound similar to that produced by loose antenna or high-voltage connections. Usually, this noise will be apparent (in varying amounts) on the 80- through 10-meter amateur bands. Many cases of line noise arcing or radiation, however, expand this field of interference to include the frequencies associated with television and the a-m broadcast radio. The magnitude of power-line noise interference varies with each case. Sometimes this "hash noise" can be tolerated, but occasionally it approaches an S-9 level and must be eliminated. Since many amateurs find themselves in an awkward position during such times, this article will present an informal guideline which may be used to help eliminate this electrical plague.

clean house first

Many noise interference situations prove to be created by sources other than commercial power lines. Thus, your own house should be in order. Check all antenna connections and transmission lines, being highly critical of any metallic objects that could come in contact with guy wires or antennas. Remember, too, that rain may cause items like wood or cloth to act as conductors. Next, check the plugs and line cords on equipment in your house, being particularly suspicious of appliances used in the kitchen. For example, bad electrical igniters, as found in many gas ranges, can create a surprising amount of interference on the ham bands. While this type in-

By **Dave Ingram, K4TWJ**, Eastwood Village 1201 South, Route 11, Box 499, Birmingham, Alabama 35210

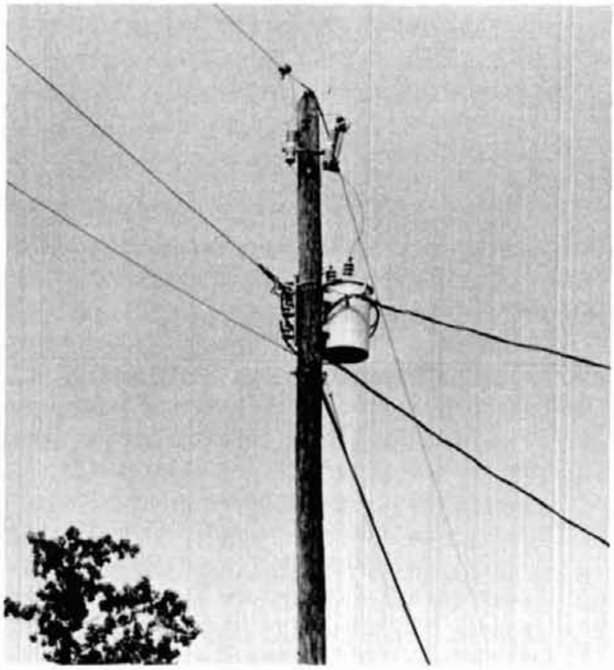
terference can be reduced by installing a bypass on each side of the ac line to ground, that doesn't eliminate the problem. The defective element should be replaced. Likewise, water heater thermostats, electric blankets, and heater tapes (used for wrapping outdoor water lines during winter months) should also be checked by temporarily disconnecting their ac power. Assuming the noise interference still exists, you are now ready for the Sherlock Holmes phase of locating this electrical villain.

locating the villain

If you have a directional antenna, it can be used to determine the approximate direction of your interference. An accurate "fix" on this noise is usually obtained by searching for a null, rather than a peak. Once the approximate direction of the noise has been determined, try mobiling in that area while listening for interference peaks on a portable rig. A two-meter sideband transceiver is particularly convenient for such ventures. Fm handi-talkies should not be used for these tests, since they should be unaffected by a-m (line noise) variations. (While a portable a-m radio could be used for location techniques, its susceptibility to normal radiation from every power line and pole makes its indications very unreliable.) When using this method, you should travel a reasonable distance beyond the point of maximum noise pickup to ensure that you've definitely located the source. While not common, it's quite possible that a remotely generated noise may be propagated along the power lines with a peak occurring right at your door. Using the S meter on your rig for indication, try walking the area of the noise source to pinpoint the interference to a specific pole, transformer, meter box, or home.

Next, study the line noise for several days, and try relating it to various weather and time situations. Noise that is more apparent on warm days than cold days, and disappears during periods of rain, is often caused by loose line clamps or cracked insulators on power poles. Noise that is more apparent during heavy-load evening hours is often caused by leaky transformers or defective heating devices. Thoroughly investigate the area around the apparent source of noise during the day and night, looking for frayed wires swinging in the breeze, small animals that may have become trapped between high tension lines, or a visible arcing near transformers or pole-top mounted insulators. However, don't climb power poles to shake wires, or viciously swing grounding lines to power poles. You could be electrocuted if a loose or corroded connection suddenly broke.

Assuming you have now determined the line-noise producing area and confined it to, say, three or four poles, you're almost ready to notify the local power company. Before doing so, however, reinvestigate the whole area and make notes on which poles have transformers and fuses. These fuses are approximately 51 cm (one foot) long, and swing into mounts near the top of the pole. The local power company



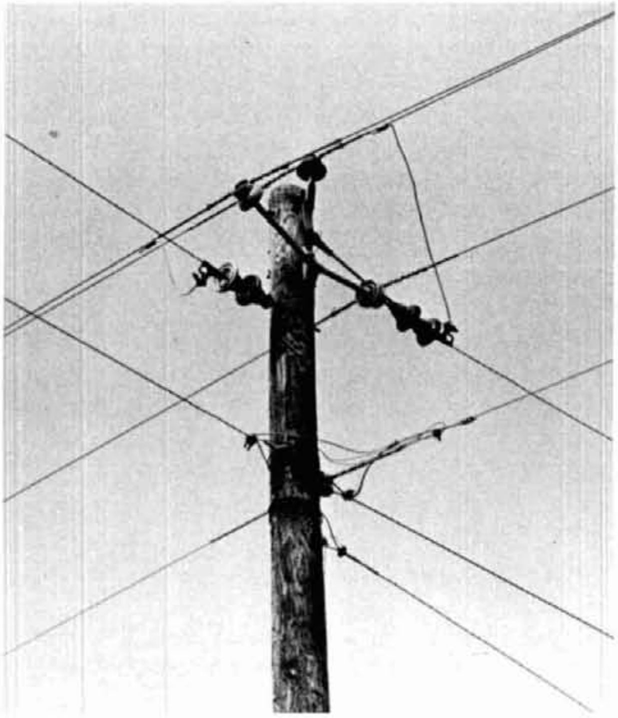
The fuse, which can be used to isolate individual lines, is located at the top of the pole. Note the large clamp on the top wire.

can disconnect these fuses ("drop lines") to locate the discrete points and elements creating interference.

call for help

Your first call to the power company will probably be stopped by the front desk, so, explain your situation, describe the noise-producing area and your detection methods. Leave you name and telephone number so their engineer can contact you before visiting the trouble area. This prescheduled meeting is particularly advantageous for tracking down periodic line-noise problems. There's nothing to be gained by looking for arcing insulators, for example, during or after a mild rain. When the power company's task force arrives, repeat the results of your investigation and monitor, with your mobile or portable gear, as each line is broken or checked. The power company will usually proceed with matters beyond this point, repairing whatever elements are found to be creating interference. Be sure to note the name of the power company employees repairing the line noise, so you can sidestep the front desk should future problems arise.

Many times, amateurs become "trapped" with front-desk executives (who've never heard of Amateur Radio, and know even less about line noise) and can't get line-noise problems resolved. Don't despair. Somewhere in their organization is a communications department — and often one or two radio amateurs. Not only are these people usually sympathetic and understanding, they usually handle special problems like line-noise interference. This group usually has its own array of elaborate noise-locating gear. Often, one or two of these men work exclusively on line-noise problems, because situations that create line noise often also creates problems that lead to power outages. Yes, you may actually be helping your community when reporting line-noise interference to the local power company. Naturally, television reception will improve when "white dots in the picture" disappear, and a-m radio reception is better without its "frying sound." But you also illustrate that neighborhood amateurs can help eliminate, rather than create, interference.



Photograph of clamps and wires which are prime sources of line-noise interference. The aluminum clamps, which hold the jumper wires, can expand and contract with temperature variations. The large insulators may crack, causing "frying" noise interference.

Many noise interference problems, when located, prove to be created by customer-use devices like doorbell transformers and thermostats. Fortunately, most power companies are quite helpful and understanding in locating these troubles. Their step-by-step method for locating these noise sources usually involves sequentially "dropping" various lines and evaluating the effect on the noise. As previously mentioned, a battery-powered, high-frequency (or 2-meter) rig is highly beneficial during these tests.

The danger of experimentation with high-tension power lines by unauthorized personnel (that's you!) cannot be overemphasized. Today's amateurs have learned to respect high voltage in the shack; the number one killer is now outdoor power lines and large antennas. Instant retirement results when these two items are placed together. Don't take chances. Stay clear of problem areas during tests, and don't stand where lines could accidentally break and fall on you.

Here is a synopsis of the more common power-line noises and their causes:

1. Constant noise during the heat of the day, intermittent noise at night, no noise during rain: Cracked insulators or loose clamps on pole. This noise is predominant during summer months.

2. Constant noise during rain, intermittent noise during light mist: Small object or animal caught between high voltage lines.

3. Constant noise in a specific direction: Power transformers or home-user devices.

checklist for line-noise corrections

1. Check your own home area by disconnecting the main circuit breaker while monitoring noise on a battery-powered rig. If noise stops, reconnect the main breaker and switch off power to each area of your home until the noise again disappears. Then pinpoint source.

2. Use directional antennas and portable/mobile gear to isolate the noise-producing area.

3. Relate noise interference to time and weather conditions.

4. Call the power company and arrange to be at home when they disconnect service to various areas.

5. As confirmation of power-line (as opposed to atmospheric) noise, the power company may elect to temporarily cut all power to an area. Monitor the results on your mobile gear.

6. Be patient and persistent. A quiet band is worth the wait and effort.

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DIGITAL TECHNIQUES

gate structure and logic families

The first part of this series explained the basic gate as a component. This part will examine the internal gate structure of TTL and CMOS, similarities and differences, as well as loading. A typical TTL two-input NAND gate is shown in **fig. 1**.

Multiple emitters may seem strange, but they are easily made in integrated circuits. They are the key to NAND-gate operation. When any emitter of Q1 is pulled down to less than +0.4 volts, its emitter-base junction is forward biased via R1. Collector-emitter junction voltage is then low enough to cut off Q2 and Q4. Output voltage goes high, since Q3 is conducting by forward bias from R2. R3 limits output current.

Q1 is nearly cut off when all emitters are at +2.4 volts or higher. Full cutoff is not achieved, since the internal structure is arranged to forward bias the base-collector junction; this allows Q2 to conduct by the forward bias through R1. Q3 cuts off and Q4 will conduct because of the Darlington connection to Q2. The collector-emitter junction of Q4 will then saturate at less than 0.4 volts.

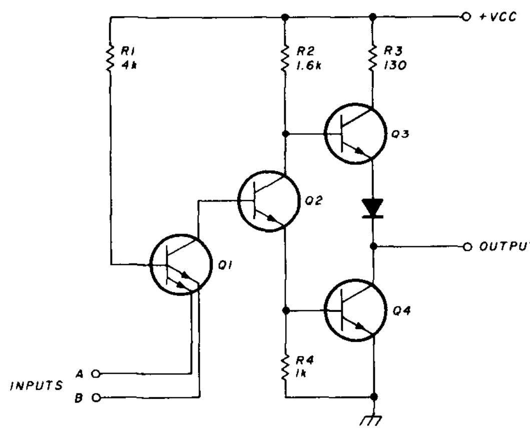

fig. 1. Simplified schematic diagram of the two-input, TTL NAND gate. In the open-collector version, R2, R3, Q3, and the diode are omitted, leaving the collector of Q4 open.

By Leonard H. Anderson, 10048 Lanark Street, Sun Valley, California 91352

TTL outputs will *sink* more current at low output levels (electron flow from ground to output) than they can *source* at high outputs (flow from output to supply). This fits the input requirements for devices connected to the output. Input current at high levels is one-fortieth of the low level current. Output current biasing is set for this 40:1 ratio and *fan-out* is usually set for driving ten inputs with one output.

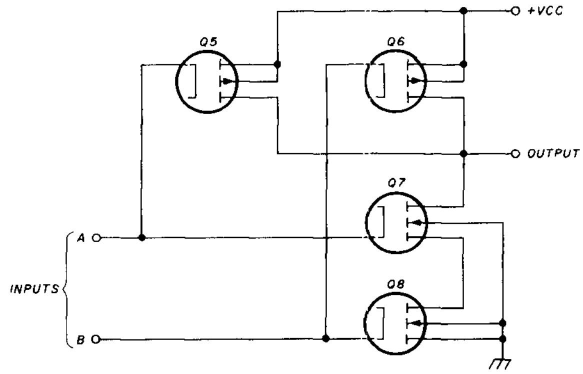

fig. 2. Simplified internal structure of the two-input, CMOS NAND gate.

Fig. 2 shows the equivalent gate function in a CMOS device. The insulated-gate FETs reduce input current to nanoamperes, primarily from leakage. Gate resistors are seldom internally used, allowing all inputs to present a high input impedance.

The N- and P-type FET arrangement in a CMOS gate will determine function. If one input is low, the appropriate parallel-connected P-type (Q5 or Q6) will conduct and make the output high. The low input overrides any high input by cutting off one of the series-connected N-types (Q7 or Q8). *All* inputs must be high for the series path to ground; the NAND RULE is satisfied.

Output sink and source current capability is nearly the same at either level; it varies with supply voltage. A-Series CMOS will work at 3 to 15 volt supplies; B-Series goes to 18 volts. The output FETs do not fully saturate, so exact voltage level is load dependent. Data sheets must be consulted for exact values.

Rearrangement of N- and P-types allows a greater range of functions than with TTL. Inverting the series and parallel connections will form a NOR. Other possibilities exist, and CMOS devices are available with a greater range of functions.

important differences

The high input impedance of the CMOS implies the capability of driving hundreds of inputs from one input. This would be true if it were not for package and circuit capacitance. Capacitance and output FET characteristics limit loading. Capacitance also limits TTL, but to a lesser extent. Input current at dc limits TTL.

Input and output level symmetry allows CMOS to be biased for linear, small-signal operation. This is impossible with TTL. A *Schmitt trigger* must be used with TTL for inputs having slow rise and fall times.

Input threshold voltages are fixed in TTL. This is primarily due to saturated bipolar transistor operation from a fixed supply. CMOS level thresholds are approximately one quarter of supply maximum for a low and three quarters of supply minimum for a high.

CMOS devices are more susceptible to noise pickup due to their high input impedance. Circuit layouts must be carefully done. TTL is more tolerant to noise, but high input levels are still affected; this can be seen from typical TTL characteristics given in **table 1**.

Total circuit current of CMOS is less than TTL. CMOS current demand is influenced by switching speed, *i.e.*, charging and discharge of the circuit capacitance. Supply current increases with increasing logic switching frequency. It is also true for TTL but to a lesser extent. TTL input current masks most of that effect.

Bipolar transistors in TTL allow faster switching speeds. FETs are becoming faster as discrete devices, but CMOS designs retain low power and slower transistors. CMOS is good for portable battery-powered equipment, but note that most such equipment is slow speed.

The NAND gate output of **fig. 1** is called a totem-pole, from the appearance of Q3 on top of Q4. Most outputs are of the totem-pole variety, but a few are *open-collector*. Since, in normal operation, Q3 conducts little current, this transistor, R2, R3, and the diode can be deleted, using an external *pull-up* resistor to the supply line for source current. This is the open-collector version.

open-collector applications

TTL chip transistors have relatively low breakdown voltages. Redesigning just the output transistor into an open-collector chip (Q4 in **fig. 1**) allows driving high-voltage devices such as relays and neon lamps.

Another function is the *wired-OR* depicted in **fig. 3**. This takes advantage of the NAND's active-low input. Several open-collector outputs can be wired to a single input, giving an equivalent OR function. The OR symbol with connecting dot is purely symbolic. No specific open-collector symbol exists.

Why use a wired-OR? One reason is economy, another the number of available packages per board. If you don't have room for a NAND equivalent OR gate, but do have an input, the wired-OR will do the job. Open-collector outputs are slightly slower than totem-poles, so be wary in high-speed circuits.

three-state wired-OR

A modification of the totem pole output was

table 1. Typical TTL gate characteristics.

	conventional			schottky	
	medium speed	high speed (H)	low speed (L)	low (LS)	medium (S)
maximum output source current in microamperes (I_{OH}) when output is high*	400	500	200	400	1000
maximum input sink current in milliamperes (I_{OL}) when output is low*	16	20	3.6	8	20
maximum t_{pLH}, nanoseconds	22	10	60	15	4.5
maximum t_{pHL}, nanoseconds	15	10	60	15	5
maximum input source current in microamperes when input high*	40	50	10	20	50
maximum input sink current in milliamperes when input low*	1.6	2	0.18	0.4	2
nominal input pull-up resistor, kilohms	3.9	2.7	39	18	2.7

*High level output voltage minimum is 2.4 for conventional TTL, 2.7 for Schottky (V_{OH}). Low level output voltage maximum (V_{OL}) is 0.4 for conventional, 0.5 for Schottky. High-level minimums and low-level maximums apply to inputs also.

*Actual high-level circuit voltage may be any value between minimum and the supply voltage, depending on output loading.

designed by National Semiconductor, called TRI-STATE by them and *three-state* by others. The design has the good points of both totem pole speed and the ability to wire-OR.

A control line is added to enable the output or outputs. An enabled output behaves like the totem pole. A disabled output appears as a high impedance. Three-state outputs can be wired in parallel and are

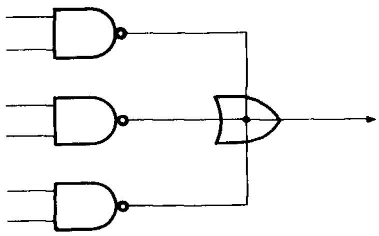

fig. 3. Symbolism used to denote a wired-OR configuration using open-collector NAND gates.

fast, but need the extra control. Such outputs are ideal for computer bus lines. Bus lines are parallel data lines carrying information in either direction.

Both three-state and open-collector output devices require some precaution during circuit design. Data books should be consulted for the fine details.*

unused inputs

All inputs must be connected somewhere. An unused TTL input will automatically assume a high state if left open (leakage in Q1 of **fig. 1**). It is poor practice to tie it high by an open connection; impedance is higher and noise might sneak in. Tie the unused input to V_{CC} or ground, depending on the desired function. Gate inputs can be paralleled. A parallel connection increases logic-1 current, but logic-0 current is the same.

Unconnected CMOS inputs can assume either state due to the inherently high impedance, leakage, and construction. They *must* be tied to either V_{CC} or ground. Removable circuit boards should have potentially open inputs connected to ground through a 100-270k resistor. This trick will protect the input from possible static surges when disconnected.

CMOS handling precautions

Unmounted CMOS devices may be damaged by static electricity; an unfelt static charge is enough. Unused devices should be kept in anti-static containers or pushed into conductive plastic foam. Such foam is usually black, somewhat hard, and will read a few kilohms on an ohmmeter.

*The Texas Instruments *TTL Data Book* is available from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048, for $4.95 plus $1.00 postage.

Everything on the workbench should be grounded, including an aluminum work-area plate. Use a *grounded-tip, three-wire* soldering iron. Strap together and ground all powered test equipment; this should be done anyway, since it is possible to get a lethal shock from some test gear.

Avoid all plastic-fiber clothing if possible. Ground yourself through a one megohm resistor (two megs with 220-volt mains) and flimsy wire. The resistor is a precaution against lethal shock current. The flimsy wire should break if you slip. All this may seem overcautious, but is standard industrial practice. Ask yourself how many expensive CMOS devices you can afford to lose from a static zap.

TTL variations

There are five different versions of TTL, identifiable by one or two letters after the 54 or 74. No letter means the original and is called medium speed. An L identifies low power and low speed. An H stands for high speed and high power. However, these two versions are being phased out of new designs in favor of Schottky versions.

LS, or low-power Schottky, is as fast as medium speed, with only slightly more power demand than L versions. S is as fast or faster than H and takes no more power than medium speed. Differences are internal and you need only consult data sheets for application. Schottky inputs are to bases of grounded-emitter input stages. This and added clamping diodes make them different from ordinary TTLs.

Schottky versions are recommended for new designs. A lot of ordinary TTLs are available at low prices and should work just well, except in very fast circuits.

I^2L, or integrated-injection-logic is similar to TTL and is circuit compatible. It is appearing in many large-scale integrated circuits.

PMOS and NMOS

This family group started before CMOS. PMOS was the first, using P-type MOS (usually junction FETs). PMOS devices need two or three supply voltages, but have inputs and outputs compatible with CMOS and most TTL devices. They find use as microprocessor memory chips; CMOS handling precautions should be observed.

NMOS uses N-type MOS and most devices use only 5-volt supplies. Again, the inputs and outputs are compatible with both CMOS and TTL. The Motorola 6800 microprocessor devices are almost entirely NMOS. NMOS is less sensitive to static shock, but it won't hurt to be cautious.

The next article in this series will go into the importance of propagation delay and discuss flip-flops.

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the **weekender**

combination field strength meter and volt-ohmmeter

Whether you want a weekend project or would like to get your club involved, here's a project that can be used to teach theory, construction practice, and troubleshooting while providing a basic piece of test equipment. The FSVOM is a combination field strength meter and volt-ohmmeter. It evolved as part of a training program for prospective amateurs. The objective was to make learning fun by providing a balance of theory and practice. My hope was that the finished product would provide a degree of confidence and encourage home construction, rather than being something that would end up on the shelf, like the old "two-transistor super" often used as a first project. In this respect, the FSVOM has been a success.

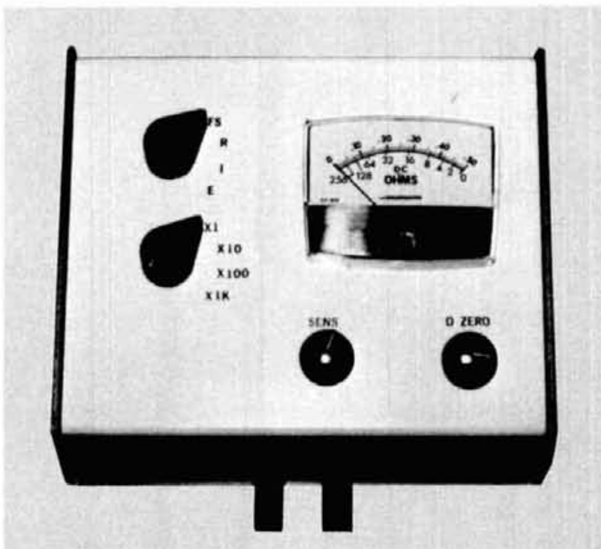
I don't think anyone need have any qualms about building this unit; there are no exotic or hard-to-find parts. Calibration is minimal. If you don't have facilities for etching the printed circuit board, an etched and drilled board (see parts list) is available. It's an easy weekend project for the experienced builder, and an excellent learning experience for the neophyte.

the FSVOM

The unit is a basic VOM, with the added feature of a field strength meter. It has a sensitivity of 20k ohms per volt, so that most readings will agree with standard-notation schematics. Circuit loading is minimal. The controls and meter scales are designed to minimize operator error. Only one voltage/current scale and one resistance scale are on the meter face, and only one set of jacks is used, so there's very little chance of error. I find the FSVOM much easier to use than many conventional units with their many scales and ranges.

The FSVOM contains four functions: field strength, resistance, current, and voltage. A multiplier (range) switch provides four ranges, **X1**, **X10**, **X100**, and **X1k**, for each function with the exception of the field strength function, which has an independent sensitivity control. The multiplier scheme allows the use of a single meter scale. The reading is then multiplied by the factor set into the range switch. A separate meter scale is used for the resistance function, which is nonlinear. Dc voltages are read at full-scale ranges of 0.5, 5, 50, and 500 volts. Dc current is read at full-scale ranges of 0.5, 5, 50, and 500 milliamperes. Resistance center-scale readings of 18, 180, 1800,

By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120



Front panel showing labels and parts layout.

and 18,000 ohms are available. Field strength is read *on a relative scale of 0-50.* The FSVOM is powered by a 1.5-volt cell and is fully portable.

circuit description

The FSVOM schematic looks a bit complex because of the switching functions; but when broken down into the four separate functions, as depicted in **fig. 1**, each is quite simple and each is an exercise in Ohm's law. The fundamental component is the meter movement, and each function is the application of 50 μA to the meter through series, parallel, or combinations of resistance values. You can readily see why this would make a good training project.

Fig. 1A depicts the basic voltage-measuring function. In this configuration we have the meter and voltage-calibration resistor in series to form a total circuit resistance of 10k ohms. Applying 0.5 volts across this series circuit yields a current of 50 μA and full-scale deflection of the meter. Smaller voltages will, of course, yield lower readings because of the reduced current flow in the series circuit. With the range switch in the **X10** position, additional resistors will be added to the series circuit to form a total resistance of 100k ohms.

The application of 5 volts across the 100k series circuit will again yield a current of 50 μA through the meter movement, causing full-scale deflection. This process is repeated again for the **X100** and **X1k** ranges, each time increasing the series resistance of the meter circuit by a factor of ten and increasing the voltage required for full-scale meter deflection by a factor of ten. This very simple circuit forms the voltage measuring portion of the FSVOM.

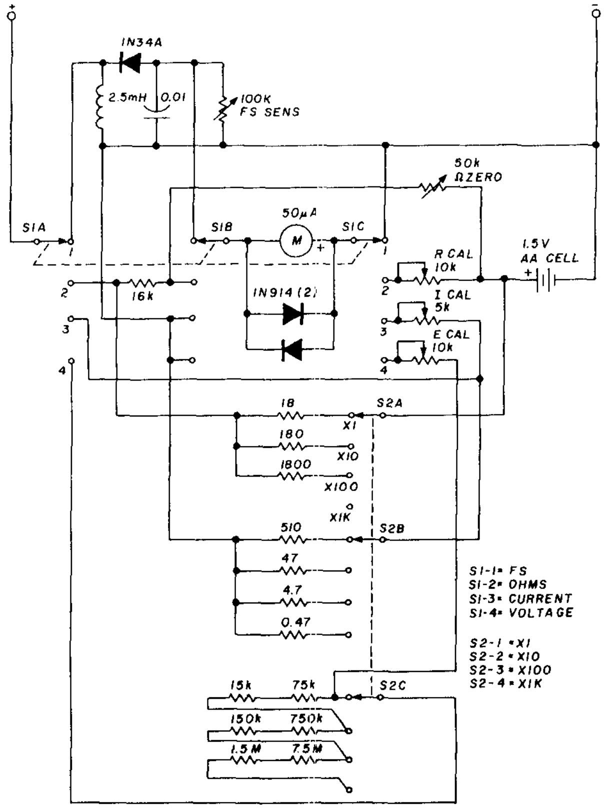


fig. 2. Complete schematic of the FSVOM.

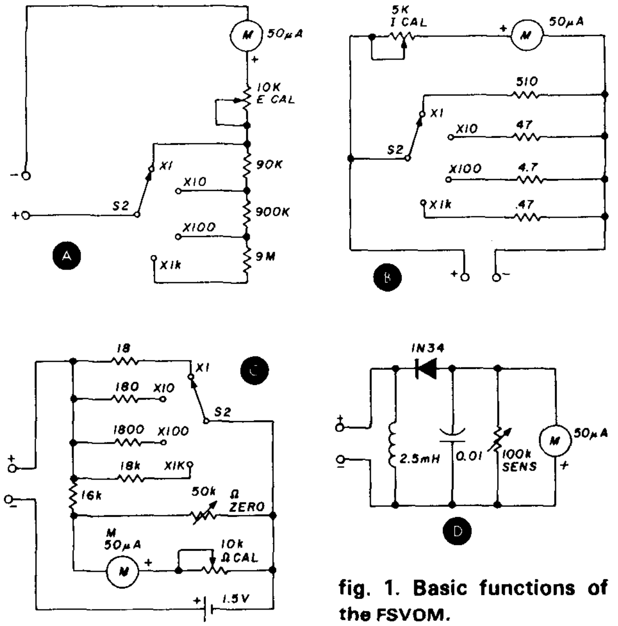


fig. 1. Basic functions of the FSVOM.

Fig. 1B shows the basic current-measuring circuit. In this function we have the meter movement and current-calibration resistor in series to form a total circuit resistance of 5k ohms. In contrast to the voltage-measuring circuit, where we used series resistors to achieve the desired ranges, we now switch resistors in parallel with the metering circuit to change ranges. The meter movement and calibration resistor form a 5k-ohm circuit and would require 0.25 volt applied across this circuit for a full-scale meter deflection. In effect, we're using this 0.25-volt, full-scale voltmeter to read the voltage drop across each shunt resistor as current from the circuit under test passes through the shunt resistor selected by the range switch.

In the voltage-measuring function, we increased the resistance values in proportion to the voltage being measured. In the current-measuring function, we decrease the resistance values as we select high ranges. Once again, our objective is 50 μA full-scale on the meter. The same scale and multiplier factors are used in the current-measuring function as in the voltage function.

Fig. 1C depicts the basic ohmmeter circuit. As with the previous circuits it's a matter of controlling the amount of current through the meter movement, but in this function things get a bit turned around.

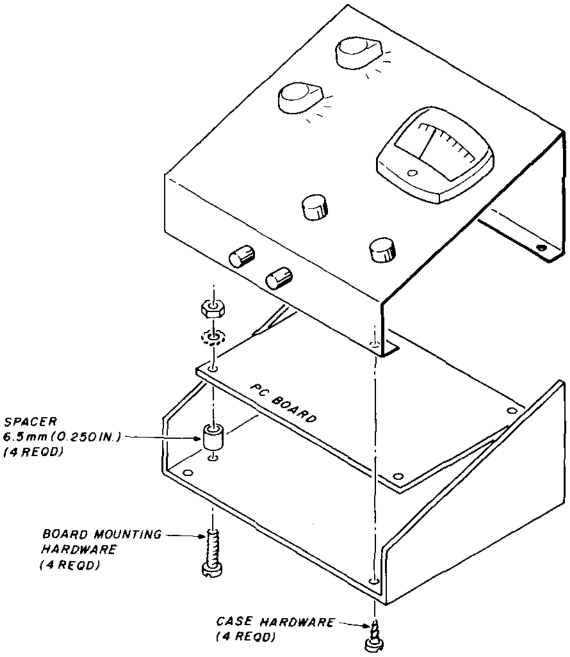


fig. 3. Mechanical details of the FSVOM enclosure.

Zero on the OHMS scale is reflected by full-scale deflection of the meter (50 μA), and for maximum resistance, or infinity, there is no deflection. This fact explains the second scale on the meter face.

In the ohmmeter function we form a voltmeter with the meter movement and the resistance calibration trimmer, just as we did in the current function. The voltage is furnished by the internal battery; it is applied to a voltage divider formed by the resistor selected by the range switch and the external resistance value being measured. When these two values are equal, the meter will read half scale.

We're comparing an internal known value against an external unknown. An OHMS ZERO adjustment control is included on the front panel for setting the meter to zero as the internal battery's output voltage changes with age and use.

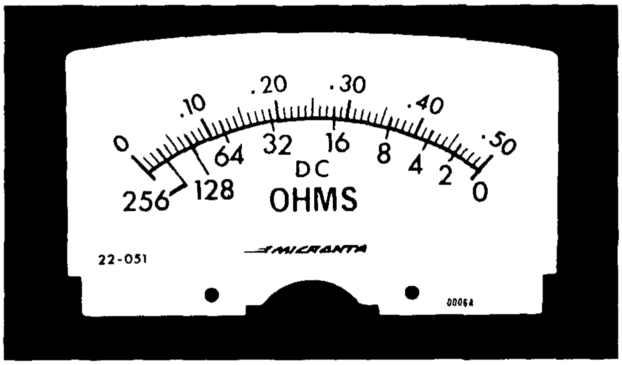


fig. 4. Modification of the meter scale using dry-transfer characters. Resistance scale is in red.

Fig. 1D is the basic circuit of the field-strength-meter function. In this function the positive input jack is the antenna input, and the rf voltage picked up by the antenna is rectified by the germanium diode. The output is applied to the meter movement. A *sensitivity control* shunts the meter to allow adjustment over a wide range of field strengths. Since the field-strength function is a relative reading, no special scale is used.

The overall schematic is shown in **fig. 2**. It's a composite of the four individual function diagrams shown in **fig. 1**. The only components added are the two silicon diodes across the meter to prevent damage from overload. After looking over the individual diagrams, it's easy to understand the composite.

table 1. Parts list for the FSVOM. The parts in this listing are those used for the unit in the photos. Parts of equivalent value may be used without degrading performance.

meter	50 μA, RS 22-051
switch S1	3 pole 4 position, Centralab PA-2007
switch S2	3 pole 4 position shorting, Centralab PA-2006
control, FS Sens	100k linear taper, RS-271-092
control, Ohms Zero	50k linear taper, RS-271-1716
trimmer, I Cal	5k, RS-271-217
trimmers, R Cal & E Cal	10k, RS-271-218
jacks, input	Nylon, RS-274-662
holder, battery	1.5 V AA Cell, RS-270-1432
rf choke	2.5 mH, Miller 6302
case	178 × 152 × 76 mm (7 × 6 × 3 in.), Mod-U-Box 3-7-6
resistors	½W 5%

PC board, etched & drilled: J. Oswald, 1436 Gerhardt Ave., San Jose, California 95125. $4.00 prepaid. (5 or more at $3.00). *Residents add sales tax.*

Resistors are available from local RCA distributors in 2% at no additional cost over 5% prices (series 8300 resistors).

The RCA resistors and cases are available from Quement Electronics, 1000 S. Bascom Ave., San Jose, California 95100.

This should be a boon if trouble with the unit is ever encountered.

construction

The first step is to drill or punch the holes in the case, using care not to mar the finish on the case (see **fig. 3**). Next, apply all the labeling to the case using Datak Letraset #K61 dry-transfer lettering. After labeling, apply two light coats of clear finish such as Datak or Krylon to protect the lettering.

While the case is drying remove the plastic front cover from the meter movement and carefully remove the meter face, which is secured by two Phillips screws. Using the periods from the K61 Letraset, change the meter scale numbers 10 through 50 to read .10 through .50, as shown in **fig. 4**. Now, using the red lines and numerals from Datak Letraset K19, add the resistance scale to the meter face. This two-

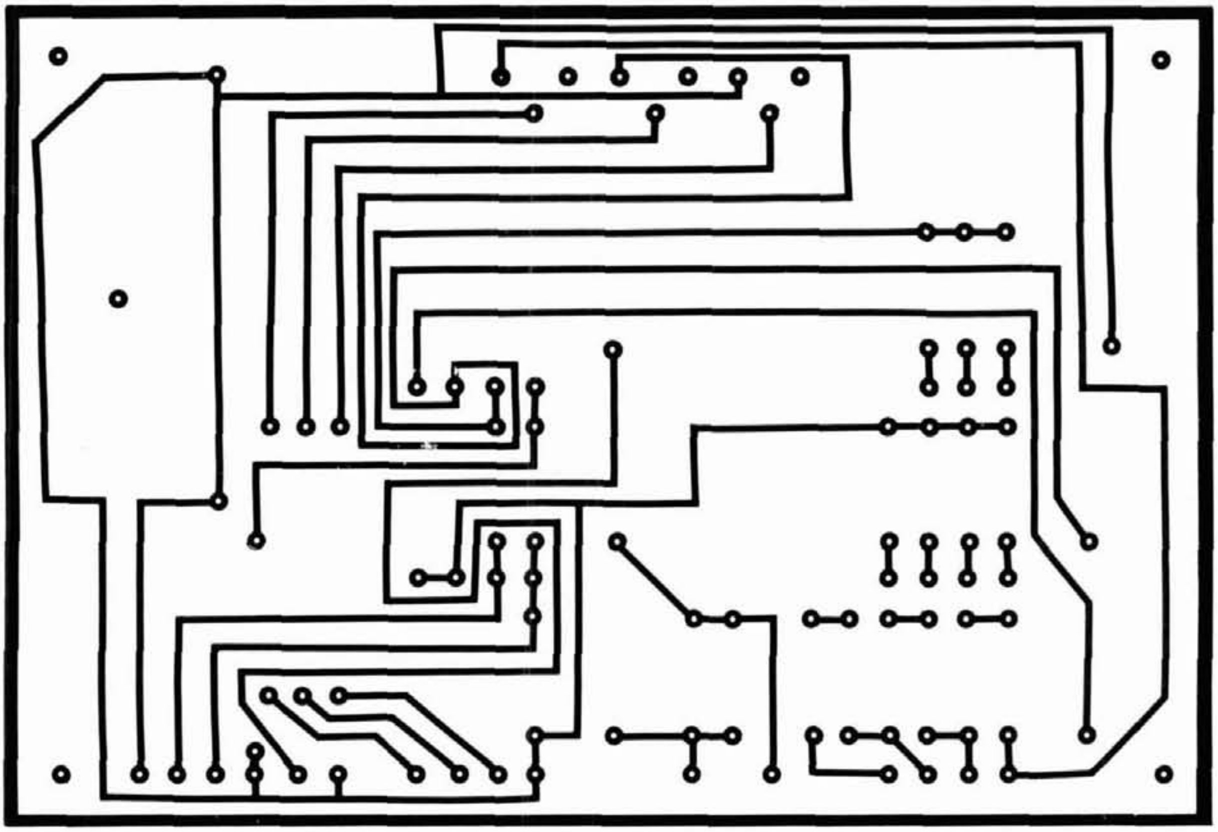


fig. 5. Foil side of the FSVOM PC board.

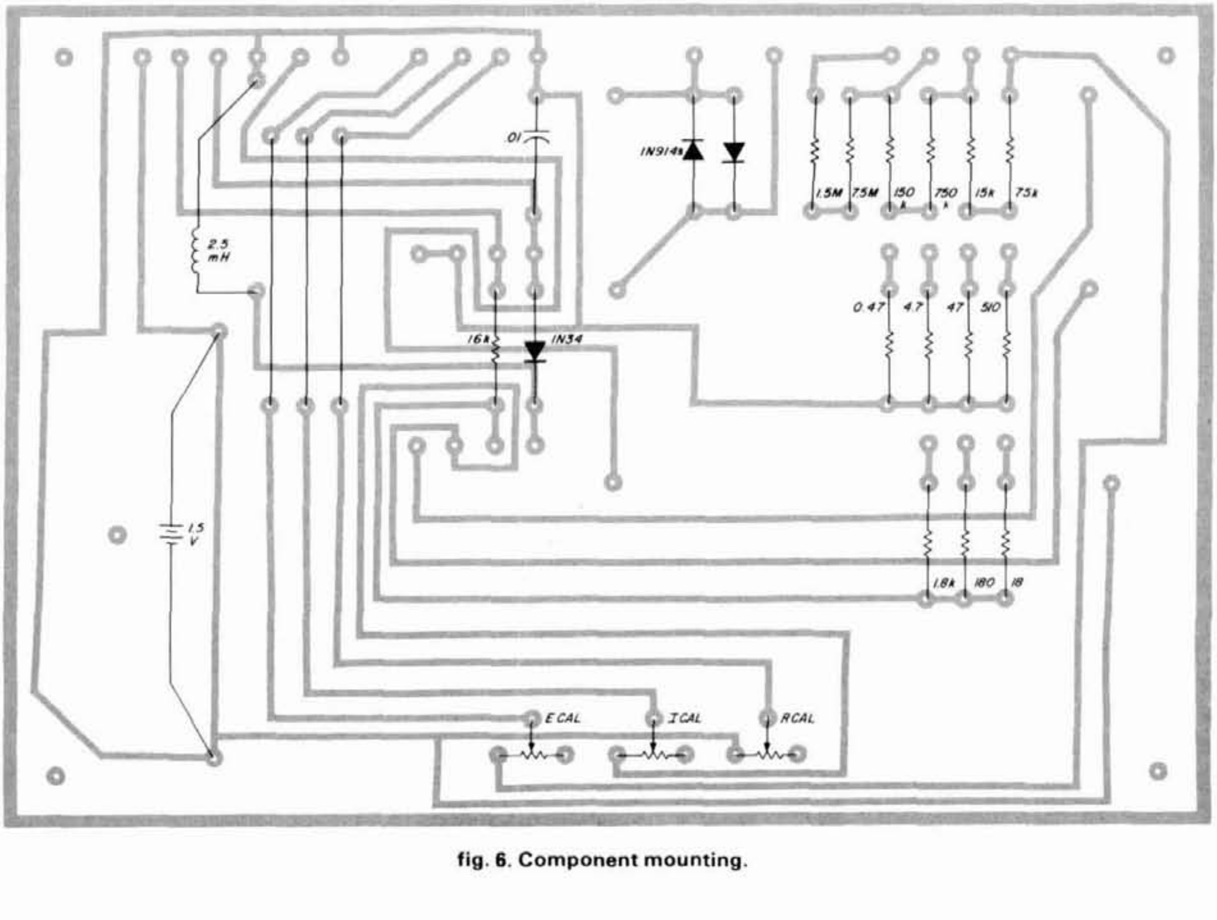
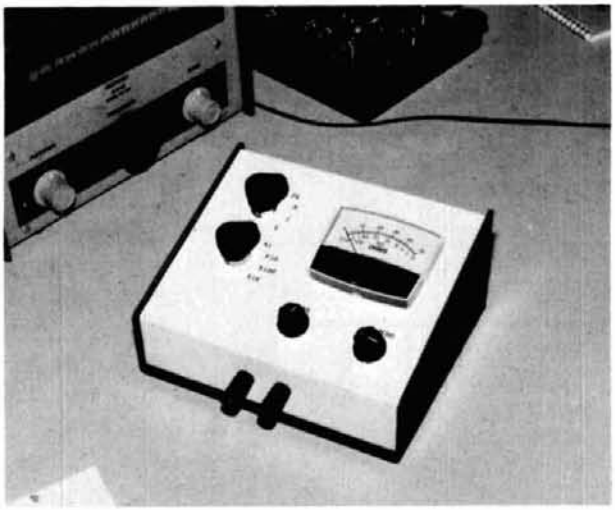


fig. 6. Component mounting.



The complete FSVOM ready for action.

color scheme will make the resistance scale easily discernible. Apply a light coat of protective spray to the meter face. Allow the meter face to dry, reassemble the meter, and mount all the front panel components to the panel half of the case.

A parts list for the FSVOM is shown in **table 1**. The PC board should be etched and drilled as in **fig. 5** and the components mounted as in **fig. 6**. Next, solder the interconnecting wires to the PC board as in **fig. 7**, leaving the wires about 305 mm (12 inches) long and in three groups: S1, S2, and panel components. Mark the far end of the wires with wire-markers or masking tape so they can be identified later.

Mount the completed PC board to the case with the wires extending toward the front. Lay the front panel face down in front of the PC board, and connect the premarked wiring to the panel components. Cut the wires to length after identifying them.

This type of prewiring saves wear and tear on the wires, as the PC doesn't have to be turned over time and again to connect the individual wires. When the wiring is done, the panel should fold back into the case and the wiring should fold neatly into place. The wiring should be laced or spot-tied for neatness. Double check to be certain that the wiring is not pinched in the edges of the case. This completes the construction. Don't button up the case yet.

calibration

The first step in calibration is to zero the meter with its zero-adjusting screw. This is best done with the meter in its normal operating position. Next, install the battery, set the function switch to OHMS

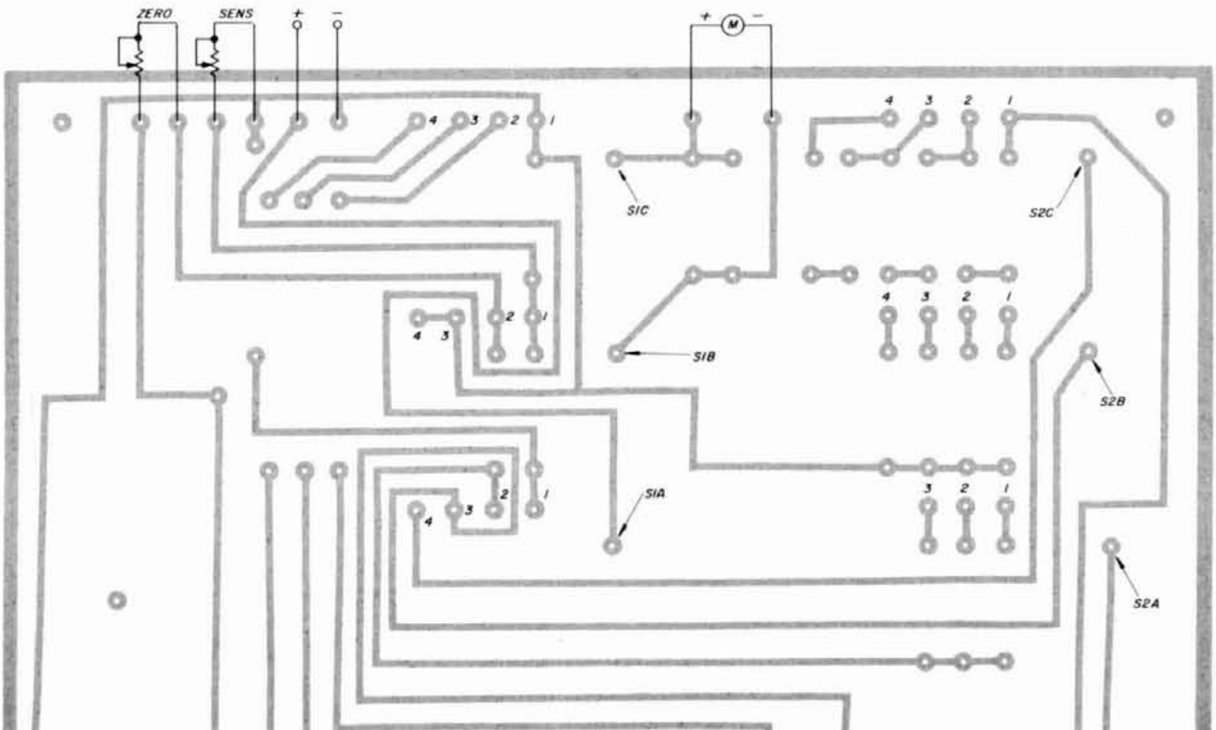
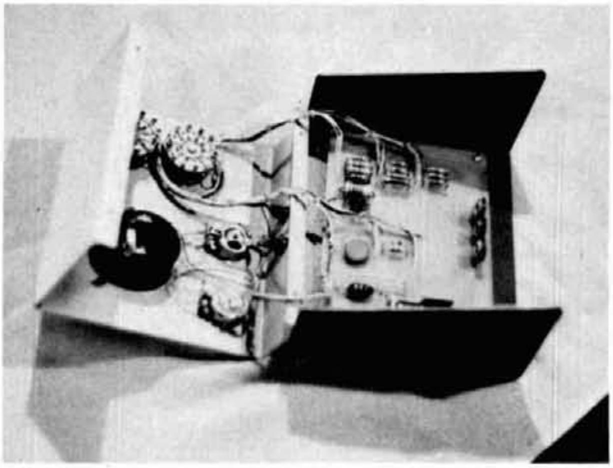


fig. 7. Board-to-panel wiring points.



Wiring details between PC board and panel.

and the range switch to X10. Leave the range switch in the X10 position for the remainder of the calibration procedure. Short the input terminals and adjust the Ω ZERO control to read 0 on the meter's resistance scale. Remove the short from the input terminals and put a 180-ohm resistor across the terminals. Adjust the **R** calibration trimmer for a reading of 0.25 (half scale) on the upper meter scale. This coincides with 180 ohms on the lower, or resistance, scale of the meter.

Switch the function switch to the current (I) position and place 300 ohms (two 150-ohm resistors) in series with a 1.5-volt battery across the input terminals. Adjust the I calibration trimmer for a full-scale reading of 5 mA (0.50 × 10).

Switch the function switch to the VOLTAGE position and place two 1.5-volt batteries in series across the input terminals. Adjust the E calibration trimmer for a reading of 3 volts (0.3 × 10). This completes the calibration procedure. Now you can fire up your rig and check out the field-strength function.

closing remarks

While I can't make any great claims of accuracy for the FSVOM, it's more than adequate for amateur use. As you probably know, most commercial meters use 1 per cent resistors and much more sophisticated circuitry. But the FSVOM is very easy to use and will provide a degree of satisfaction in that you *know* what's going on inside the little meter when you use it.

As a club project, it's fun and gets amateurs involved in areas we generally don't explore. As a useful device, it's hard to beat. When you've finished building this one you will have explored just about all phases of home-brew construction, from PC boards to rotary switches to meter movements. The experience gained on this little project should pay off well on all future building efforts.

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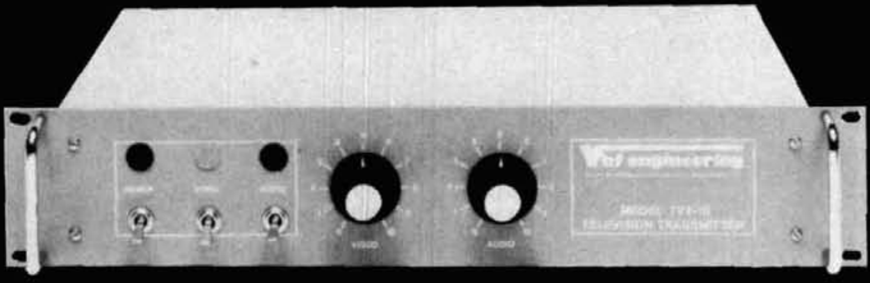

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More Details? CHECK—OFF Page 126

improving the HT-37 ssb transmitter

The Hallicrafters HT-37 is still around and has many excellent features — here are some ideas for upgrading this venerable old rig

Of all the equipment manufactured by the Hallicrafters Company, perhaps the most popular was the HT-37 transmitter-exciter. It uses the phasing system for single-sideband generation, which is a good method if the phase-shift networks are properly adjusted. Once set, however, these phase shifters rarely need adjustment. The rig is as stable as a rock and the ssb audio quality compares well with that from filter-type circuits.

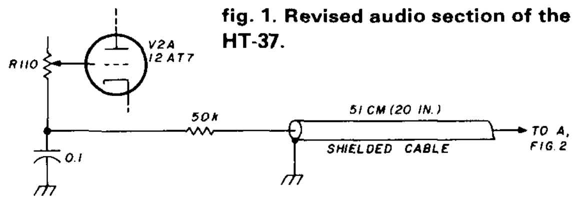


fig. 1. Revised audio section of the HT-37.

This article is presented for HT-37 users who might wish to improve its operation with some minor modifications. For the purist who wants better sideband suppression, an easy modification appears in reference 1, which shows how to install a filter-type sideband generator. If you're happy with the as-built HT-37 sideband generator but wish to make a couple of other simple modifications to improve efficiency and increase power-transformer life, you might be interested in the comments that follow.

Automatic Level Control (ALC) allows you to operate at higher audio levels without overloading the transmitter, which causes interference to nearby stations. With an ALC circuit added to the HT-37, more emphasis is given to lower-frequency speech components, and the higher dynamic range (louder) speech components won't overdrive the final-amplifier stage. Such overdriving creates splatter, "buckshot," and a broader signal.

You'll need the following parts for the ALC-circuit addition:

resistors

quantity	type
1	50k composition, 1/2 watt
1	1 meg composition, 1/2 watt
2	10k composition, 1/2 watt

capacitors

quantity	type
2	0.1 μF paper, 200V working
1	0.25 μF paper, 200V working
1	0.001 μF ceramic 200V working

miscellaneous

quantity	description
2	1N2070 or 1N2071 silicon diodes, 400V PIV
2	solder-lug terminal strips
90 cm (3 ft)	small plastic-covered shielded cable

procedure

1. Remove both top and bottom halves of cabinet.

2. Remove side rail from chassis left side.

3. Remove cover from the audio section to gain access to the audio gain control, R110.

4. Remove the direct connection between R110 and ground. Install a 0.1-μF capacitor between these points.

By Alf Wilson, W6NIF, 1068 Arden Drive, Encinitas, California 92024

5. Solder one end of a 50k resistor to the former grounded terminal of R110, then connect the center wire of a 51-cm (20-inch) length of small shielded cable to the other end of the 50k resistor. Ground the cable shield to a convenient point inside the audio compartment. The revised audio section will then be as shown in **fig. 1**.

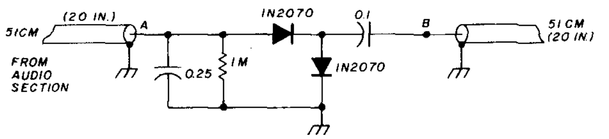


fig. 2. The ALC circuit built from instructions in the article.

6. Feed the free end of the shielded cable through a hole in the audio-section cover shield. Replace the cover shield onto the chassis.

7. Replace the side rail removed from the chassis.

8. Using a solder-lug terminal (4 or 5 lugs), build the ALC circuit shown in **fig. 2**. Be sure to use a terminal strip with a grounded mounting lug at one outer end.

9. Solder the outer end of the shielded cable coming from the audio section to point **A** in **fig. 2**. Attach another piece of shielded cable, about 51-cm (20-inches) long, to point **B** in **fig. 2**.

10. Remove the nut from one of the four machine screws securing the antenna coax fitting to the chassis rear apron. Mount the ALC assembly just constructed on the machine screw, replace the nut, and tighten securely. Of course, a grounding lug should be used to mount this assembly onto the coax connector.

11. Near the bottom of the final-amplifier 6146 tube sockets you'll find a solder-lug strip where R19 (1k) is connected to rf choke L13, which in turn is connected to the 6146 grids.

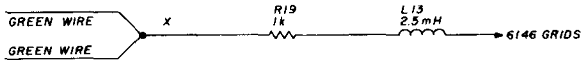


fig. 3. As-built appearance of the HT-37 (see text).

Before performing the next step, note that two green wires are connected to R19 as shown in **fig. 3**. Now proceed as follows:

12. Remove both green wires from point **X** in **fig. 3**. Leave them free for now.

13. Mount two 10k resistors and a 0.001-μF ceramic bypass capacitor on a 4-lug terminal strip, as shown in **fig. 4**.

14. Next mount the assembly just completed under one of the self-tapping screws that hold the lid onto the shield can next to the 6146 sockets. (This is merely a suggestion; use any mounting position that seems convenient.)

15. Connect the two green wires lifted in step **12** to point **C** (**fig. 4**) and run a new wire from point **D** (**fig. 4**) to R19 (point **X** in **fig. 3**).

16. Connect point **E** to the shielded cable coming from point **B** in **fig. 2**. The final-amplifier wiring will now appear as in **fig. 5**.

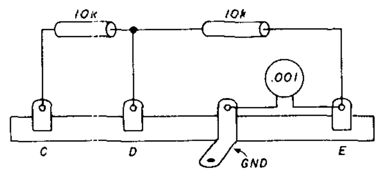
17. Dress the new shielded cables as close as possible to the chassis. Check all wiring for errors. This completes the installation of your new ALC circuit.

A schematic of the complete ALC circuit is shown in **fig. 6**. I suggest you recheck the 6146 bias voltage to ensure it hasn't changed. It should be −49 Vdc.

power-transformer protection

The HT-37 seems to have a history of power-transformer failure. I've talked to several HT-37 owners

fig. 4. More construction details for the ALC addition to the HT-37 final-amplifier section.



who have had to replace the transformer because of a short circuit either in the secondary windings or between primary and secondary windings. It's pretty hard to find an exact replacement for the HT-37 power transformer today, although at least one source of help appears in the amateur ads in which a transformer rebuilding service is offered.

In any event, it's possible to preclude catastrophic failure of the power transformer by simply adding an autotransformer, such as a Variac, in the primary voltage circuit of the HT-37.

HT-37 owners will note that, when the OPERATION switch is turned from OFF to STANDBY, a distinct *"thung"* sound will be heard if the peak of the ac primary voltage occurs at time of switch turn-on. This means that a surge of voltage is presented to the power transformer primary at the instant of switch turn-on.

Why not eliminate this surge by using an autotransformer in the transformer primary? With the OPERATION switch in the OFF position, turn the autotransformer to zero, then gradually advance the autotransformer control until the proper ac input voltage is presented to the power-transformer primary. An ac voltmeter should, of course, be connected across the power-transformer primary.

Another cause of power-transformer failure, according to many HT-37 owners, is sheer carelessness

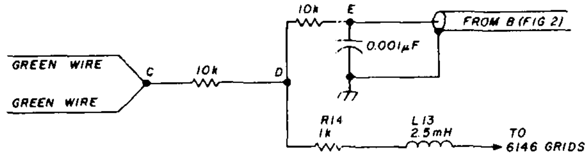
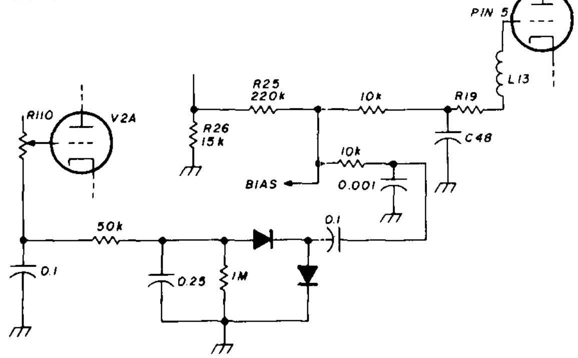


fig. 5. Schematic showing modified wiring in the HT-37 final-amplifier section.

during operation. The HT-37 instruction manual stresses that, when switching to STANDBY from any of the operating modes (MOX, VOX, or CAL), it's important to wait for a few seconds before switching the set OFF. There must be a message here. According to other HT-37 owners, it's a dead cinch that the power transformer will blow if the set is rapidly turned from one of the operating modes to OFF and back on! It's easy to do this with this equipment, especially during the heat of a contest. A Variac won't help in this case, of course, because during operation the Variac will be adjusted for full input primary voltage.

fig. 6. Complete ALC circuit for the HT-37 transmitter.



So if you value your HT-37 power transformer, respect the precautionary advice in the instruction manual for the OPERATION switch. The HT-37 is a great rig, even by today's standards. But if you can't replace a blown power transformer, you may as well try to sell the rig for junk.

acknowledgment

The material on adding ALC to the HT-37 was taken from a paper by WØNCK and KØTYO. This paper was included with the instruction manual for my HT-37, which I purchased second hand. I built the circuit and it is an improvement over the original HT-37 design. Any credit for this improvement should go to WØNCK and KØTYO.

reference

1. Milton L. Pokress, W3CM, "Increased Sideband Suppression for the HT-37," *ham radio*, November 1969, pages 48-51.

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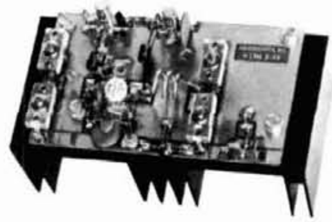

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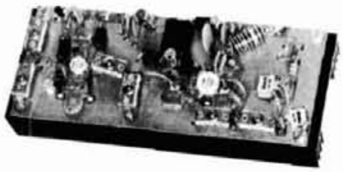

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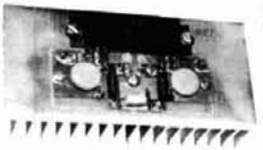

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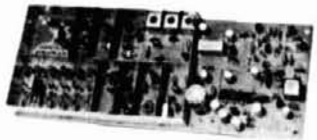

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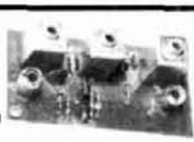

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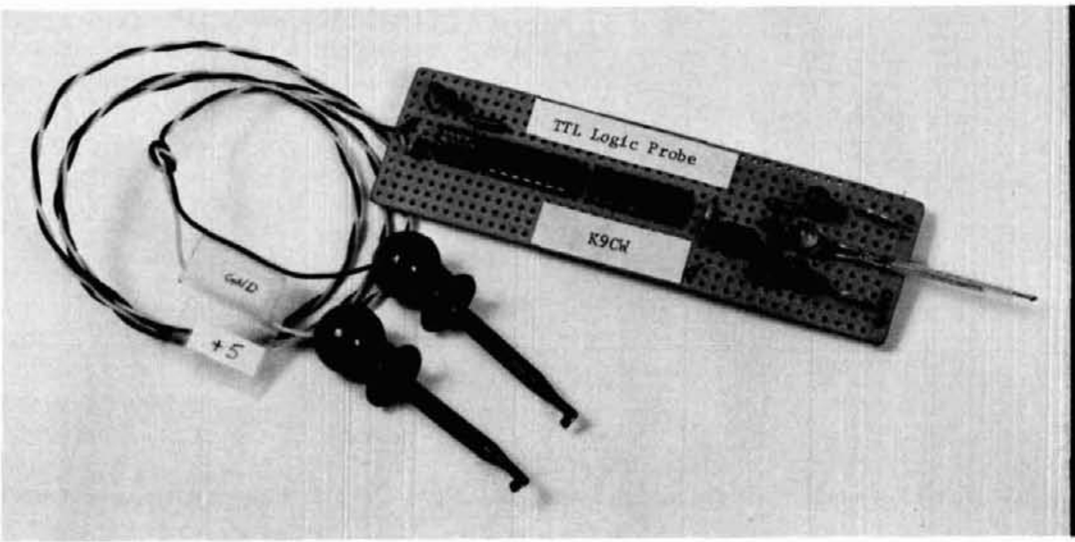

high-performance TTL logic probe

Design of a simple yet high-performance logic probe using three low-cost integrated circuits

Have you constructed a frequency counter or keyer that didn't work when you first turned it on? Has some piece of equipment that uses TTL integrated circuits failed recently? If you answer yes to either of these questions, you've been in a situation in which a logic probe could have been very useful. This article describes a simple TTL logic probe that can be constructed for less than $5. Yet its performance equals that of units costing as much as $45.

operation

The logic probe is to the digital designer what the VOM is to the electrician. With a single touch, the probe displays the logical state or condition of the selected circuit connection. A well-designed probe will indicate an open circuit, a good logic 1 or 0 state, and whether a pulse train is present. This tool enables the digital experimenter to follow signals through the various logic gates and circuit components until the faulty part or wiring error is isolated.

The logic probe has three LEDs of different colors

By Andrew B. White, K9CW, 102 Franklin Street, Urbana, Illinois 61801

mounted near the probe tip. After the probe's +5 volt and ground connections are made to the circuit under test, the tip can be touched to any point carrying a TTL signal. If the point is less than about 0.7 volts, the 0 (green) LED will be the only one energized. If the test point is 2.4 volts or greater, only the 1 (red) LED will come on. If the selected circuit is oscillating between 1 and 0, the red and green LEDs will show the relative duty cycle of the oscillation, and the PULSE (yellow) LED will pulsate or will seem to be continuously on.

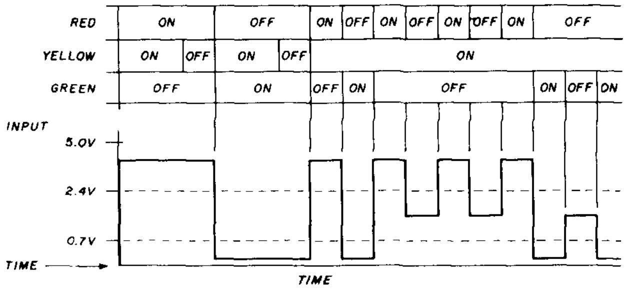

fig. 1. Graphical summary of the indicating LEDs for various input states. When connected to a floating input, none of the LEDs will be on.

The advantage of a special pulse indicator can be seen when considering a normally low or 0 signal with 200 nS high pulses every 200 mS. Since the signal is high for only a small fraction of the total period, the probe's green LED will seem to be always on and the red LED always off. However, each of the pulses is stretched to about 0.5 second and displayed on the yellow LED. Thus, the green and yellow LEDs on together indicate the presence of the narrow, high pulses.

If the line is an open circuit, has a resistance of more than about 4000 ohms to ground, or has a voltage level between 1.0 and 2.4 volts, neither the high nor low LED will be energized. An unterminated TTL input will be displayed in this fashion. A summary of the logic probe's operation, with various input waveforms, is shown in **fig. 1**.

circuit description

The schematic diagram of this logic probe, as illustrated in **fig. 2**, consists of three sections: the high-level detector and red LED driver (Q1 and Q2), the low-level detector and green LED driver (U1A and U1B), and the pulse detector and yellow LED driver (U2 and U3). Q1 functions as a threshold detector, which will turn on with an input voltage of greater than about 2.4 volts. The threshold level is determined by the three silicon diodes connected between the emitter of Q1 and ground. Each diode exhibits a forward-biased voltage drop of about 0.7 volts. When Q1 turns on, Q2 is forced on, connecting the +5 supply voltage to the anode of CR3, turning on the red LED.

The low-level threshold detector uses one section of a 74LS02 and a silicon diode. Without this additional diode voltage drop, the green LED would be energized at about 1.4 volts. With the diode, the input voltage has to be below 0.7 volts before pin 1 of U1A will go low, turning on the green LED. Notice that between the upper and lower thresholds neither CR1 nor CR3 will be energized.

The pulse detection circuit employs a dual D-type flip-flop and a monostable multivibrator to stretch the input pulse length. The detection of a valid 0 or 1 level is signaled, respectively, by U1 pin 4 or the cathode of CR3 going high. Either of these conditions causes one of the outputs of U2 to go high, forcing pin 13 of U1 low, which triggers the monostable and turns on the yellow LED, CR2, for about 0.5 seconds. As long as the output of the monostable is high, both flip-flops remain cleared. The D-type flip-flops are used to insure that U3 is properly triggered. The trigger pulse must be at least 200 nS long and cannot stay low, since that would hold CR2 on. The 74LS74 shown will respond to pulses as narrow as 25 nS.

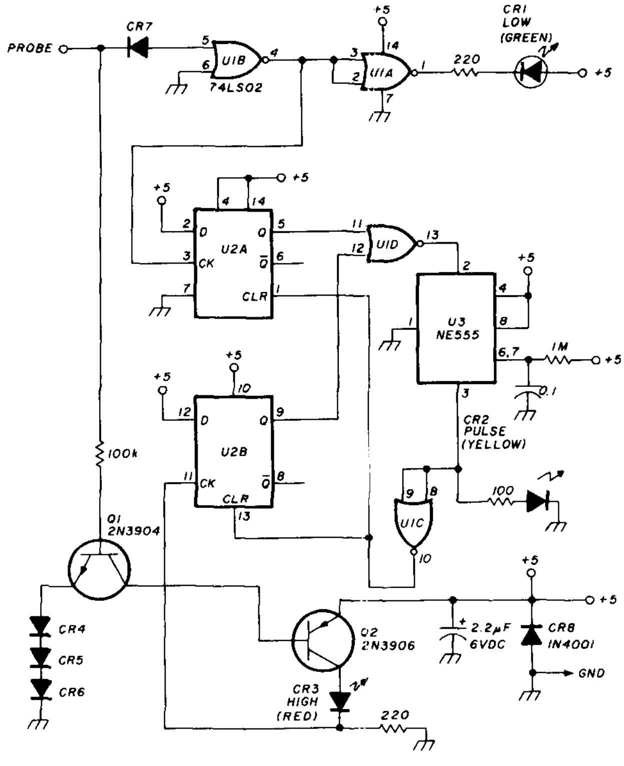

fig. 2. Schematic diagram of the logic probe. All resistors are 1/4 watt; CR4, CR5, CR6, and CR7 are 1N4148 or equivalent small-signal diodes.

A logic probe should not excessively load the circuit under test. This probe will sink only about 25 μA when the input is +5 volts. With a 0-volt input, the circuit input current is about -200 μA. In the worst case, this probe would represent one LSTTL

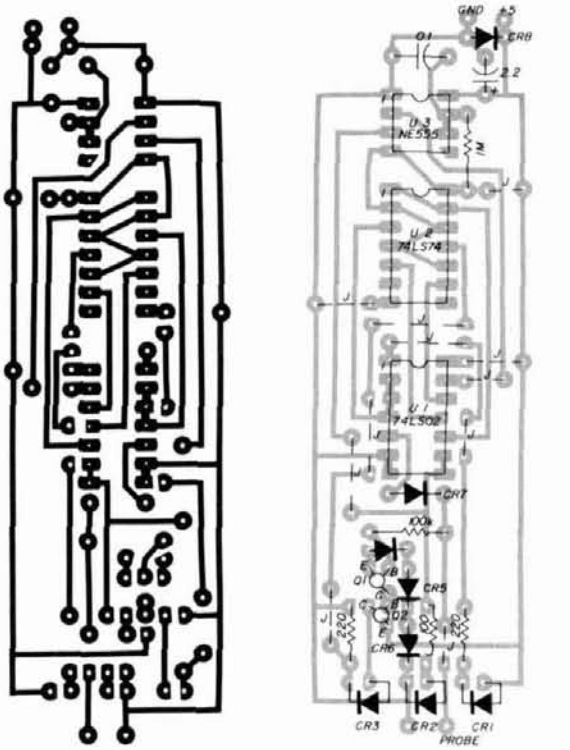

fig. 3. Suggested printed-circuit board layout and parts placement diagram for the logic probe. If desired, prepunched board and simple point-to-point wiring may be used.

load. In addition, the probe will "catch" pulses shorter than 100 nS in duration.

construction

The circuit layout for this device is not critical. **Fig. 3** illustrates one printed circuit board layout design and parts placement for the logic probe. The overall board dimensions are 3.3 × 10.2 mm (1.3 × 4 inches). You could use point-to-point wiring on a circuit board with holes on a standard 2.5 mm (1 inch) grid instead. The probe tip can be made from 16 AWG (1.3 mm) tinned-copper wire or a small brass nail. The +5 and ground inputs should be connected to clip leads so they may be easily connected to the circuit supply voltage source. Standard TTL chips may be substituted for the LS devices shown, but the circuit loading will be increased.

I have found a logic probe to be a valuable aid in debugging TTL projects. After you use one, you may wonder how you were able to get along without it!

ham radio

code speed counter

Do you dislike counting up the words and doing the arithmetic to check your code speed? If so, then you need a code speed counter. All it takes is an ordinary frequency counter and a couple of ten-cent ICs.

In many cases, a frequency counter can be set up to count for one second, performing the latch and reset function during the next 0.2 second. By inserting a divide-by-four counter (7493) in the clock circuit, the counter will then count for four seconds instead of one second.

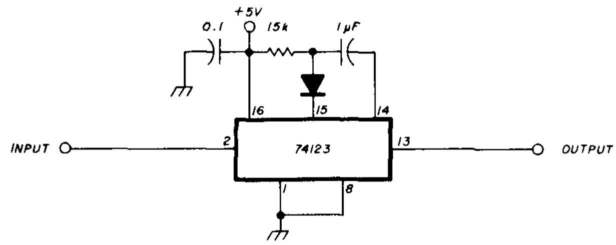

fig. 1. Wiring diagram of the one-shot multivibrator used to convert the sine wave into a pulse for the counter. Pin 2 connects to the output of the wave shaper, while pin 13 goes to the counting circuit.

The goal is to have the counter count the number of key downs during a four-second interval. This will be approximately the code speed in words per minute. A dot or dash from the speaker terminals is, in many cases, a 700 to 1200 Hz sine wave. It is necessary to generate a positive pulse from each character. To do this, insert the retriggerable monostable shown in **fig. 1** before the counter's input.

performance

Using a piece of graph paper, find out exactly what readouts the modified counter will produce. In **fig. 2**, the word *Paris* is represented in code. If you assume the speed is 12 wpm, then each dit is 0.1 second long. The counter reads 0 until the first four seconds have passed. It will then read out 12, since 12 key downs occurred during that time. Those key downs sent during the 0.4 second of the latch and reset cycle are ignored by the counter. The 12 will remain on display until the next four seconds have passed, then read out the number of key downs during that period, and so on.

If you plug the counter into the speaker jack and tune in the W1AW code practice, you'll find that the counter produces readouts which are very close to the announced speeds. On the faster speeds, 20 wpm or more, the counts tend to be low. This is because at faster speeds more words will occur during the four-second count period, and the counter does not make any allowance for the seven-baud spaces between words.

A very useful operating aide is provided by connecting the input of the 7490 counters to one of the decimal points in the readout. This makes it possible to see the dits and dahs which are being counted. By watching the blinking decimal point you can tell when interference is messing up the count. Using the gain control on the receiver you can tune out the interference, if the desired signal is strong enough. Just turn down the volume until only the desired signal is visually evident. You can also tell, by watching the flashes of the decimal point, which four-second intervals are representative of the speed being sent, and ignore counts of intervals containing long pauses or interference.

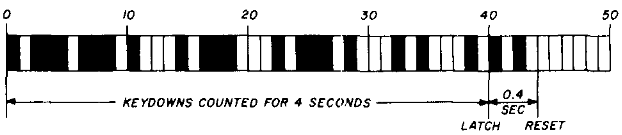

fig. 2. Diagram of the standard 50-baud word PARIS.

If you don't have a counter which you wish to modify, then you can build the code counter from scratch by adding the above modifications to any of the plans which have appeared in the literature.

The code speed counter will add another element of enjoyment to operating CW. It will also give you an easy way to check on your progress if you're trying to build up your code speed.

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By Louis C. Graue, K8TT, 624 Campbell Hill Road, Bowling Green, Ohio 43402

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For outside installations, stainless-steel clamps should be used. The area of the pole that comes in contact with the clamp and wire should be free of dirt and grease. Clean the pole with steel wool or very fine sandpaper. Be sure to wipe off any dust from sanding.

After the clamp is secured to the pole, it should be taped with plastic electrical tape for weather protection. I've employed this method to connect a grounding wire to a ground rod of small diameter, and it proved very successful.

Jim DiSpirito, AB9Q

remote crystal switching

With the rising theft rates of mobile equipment, it is becoming increasingly attractive to mount vhf radios in the trunk. However, such an arrangement can present problems if you normally use a large number of channels. In a conventional remote-switching system, the relays or diodes, which are used to switch crystals, are selected by a simple rotary switch in the control head. A separate conductor, from the control head to the radio, is required for each channel. However, the use of a large multiconductor cable can be avoided by implementing a simple binary encoder/decoder system. By using binary coding, only four conductors are needed to select sixteen channels, or five conductors for thirty-two channels. Such a system has been incorporated in my mobile installation, which was built around a modified Genave chassis.

The control-head circuit for the first ten channels (0 through 9) is shown in **fig. 1**. Of course, this can be expanded to any number of channels. Binary encoding is accomplished through the use of 1N4148 or equivalent steering diodes. If you have trouble finding a rotary switch with enough positions, you might try using a toggle switch to control the most significant bit. In this way, an 8-position rotary switch and a toggle switch can select sixteen channels.

As for decoding circuitry, there are a number of decoder ICs on the market, with a variety of specifications. The user should select one which is best suited to his particular application. The 7445 and 74145 ICs are BCD-to-decimal decoders with open-collector outputs, suitable for driving reed-relay switching.

If diode switching is used, the 7442 BCD-to-decimal decoder or 74154 4-line to 16-line decoders will work well. A 4-line to 16-line decoder may also be built from two 7442s and an inverter as shown in **fig. 2**. This illustration shows the decoder/diode driver currently being used in my mobile installation. This configuration is suitable for use with oscillators in which the crystals are selected by grounding. The 7442s were used instead of a 74154 because they happened to be available. TTL inputs normally float to a high logic level if left unconnected. The four 150-ohm resistors pull down the inputs to approximately ground

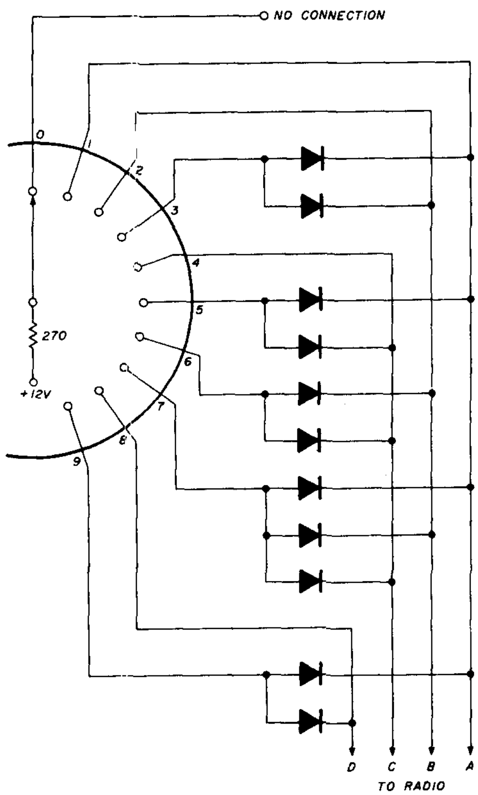

fig. 1. Schematic diagram of binary encoding circuit in the control head. The diodes provide steering such that the four lines to the radio receive the correct binary coding.

potential, unless a positive voltage from the control head is present. The reason for using this method is to provide relatively low-impedance inputs to the decoder to insure immunity to rf pickup and other extraneous signals. To be on the safe side, all four lines should also be bypassed with 0.01- or 0.005-μF capacitors at the decoder.

Inverters of the 7404 type provide a positive voltage and enough current to turn the diodes fully on. Diode current is set by the 82-ohm resistors. Larger-value resistors were initially tried, but they did not allow adequate diode current to flow. As a result, the receiver oscillator output was diminished, which in turn caused a small reduction in receiver sensitivity. A value of 75 or 82 ohms allows adequate current to flow, while remaining well within the current limitations of the diodes.

Notice that no rf chokes or bypass capacitors are required in the vicinity of the crystals or switching diodes, unlike some circuits appearing in the handbooks. A decoder could probably be designed without inverters on the outputs, but rf chokes would very likely be necessary to isolate the crystals from each other. With prices being what they are, it would be considerably cheaper to use hex-inverter ICs for isolation rather than rf chokes.

Binary channel addressing is readily adapted to more sophisticated digital control applications. For example, the radio can be converted into a scanner simply by adding a gated clock oscillator and a 7493 4-bit binary counter. The binary format is also well suited to microprocessor control.

Binary coding circuits have seen continuous service in my mobile installation since late 1974, first with DIP reed-relay crystal switching and more recently with diode switching. They have proven to be rugged and reliable, and have added considerable flexibility to the mobile system design.

J. Lee Blanton, WA8YBT

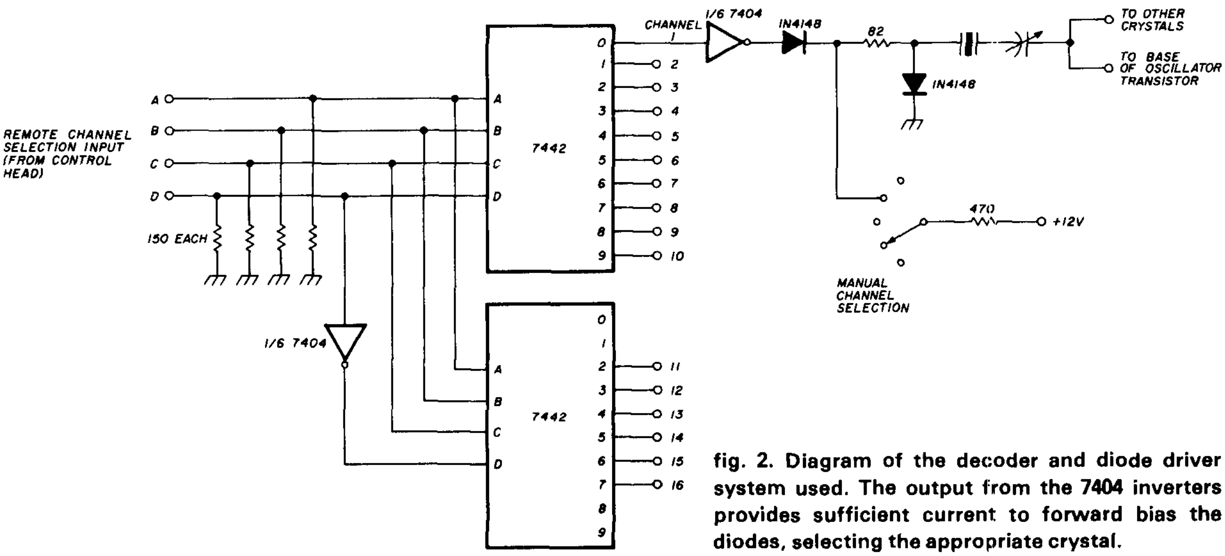

fig. 2. Diagram of the decoder and diode driver system used. The output from the 7404 inverters provides sufficient current to forward bias the diodes, selecting the appropriate crystal.

solid-state rectifiers in the Collins 516-F2

Not long after acquiring my present Collins station, I experienced a problem in the power supply which was directly traced to a cracked high-voltage rectifier tube socket. The socket becomes brittle after some years of use. This is primarily due to the large amount of heat generated by the rectifier tubes. Of prime concern though, is the power transformer which might be damaged due to such an occurrence; they are expensive to replace.

Rather than simply replacing the socket, I decided to eliminate the heat source by substituting plug-in solid-state rectifiers for the tubes. This requires a minimum amount of work since both the 5U4 and 5R4 were replaced by SS-5R4 units purchased from United Page Incorporated. Another procedure, which was suggested by the Collins Service Department, is the use of four Semtech SCH-5000 diodes (CPN 353-0425-010), two diodes replacing each rectifier tube. These are hard-wired to the tube sockets underneath the chassis.

In each instance it would be a good idea to disconnect the rectifier filament leads, insulate and tie them back out of the way, although I ran my unit for some time after installing the plug-in units without doing so. The filament leads are not needed and removing them eliminates one more source of possible trouble. Also, the bias should be re-adjusted to obtain the proper resting plate current since all voltages will have increased about 10 per cent.

Paul Pagel, N1FB

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short circuits

500-watt power supply

The article by WA6PEC, December, 1977, *ham radio*, page 30, contains a drafting error in **fig. 1**. There should not be a connection between the collector and base of Q3.

tone-burst generator

The tone-burst generator by WA5KPG described on page 68 of September, 1977, *ham radio* contains an error in **fig. 1**. The junction of the 1N914 and R2 should not be connected to the PTT line, but only to C3.

active bandpass filters

The article on active bandpass filters (December, 1977, *ham radio*, page 49) contains some errors in the equations. An errata sheet is available by sending a self-addressed, stamped envelope to either the author or *ham radio*.

admittance, impedance, and circuit analysis

In Anderson's article in the August, 1977, issue, there's a glaring typographical error in the second sentence of the right-hand column on page 76. This should read, *"if R = 10 ohms, then G = 0.1 mho."* There should be no minus sign in front of the 0.1 mho.

10-GHz broadband antenna

In the article on the broadband 10-GHz antenna which appeared in the May, 1977, issue of page 40, the chart of optimum feed beamwidth (**fig. 3**) is incorrectly labelled. The ordinate should be labelled as feed angle, *not* half-angle. Therefore, the simple antenna should be used with reflectors with focal length to diameter ratios, *f*, between 0.3 and 0.6. Thanks go to N6TX for spotting the error.

digital frequency counter

The counter article on page 22 in February, 1978, *ham radio* contained some drafting errors. The 9368 IC in **figs. 2** and **3** should have V_{CC} connected to pin 16 instead of pin 11. Also in **fig. 3**, the short across the crystal should be omitted. In all cases, the collector resistor for the 2N5179 is 1000 ohms.

A circuit board layout of **fig. 3** is now available from the author; a paper print is available by sending him a self-addressed, stamped envelope. A film negative is also available from the author for $5.00.

ssb phasing systems

On page 58 of the January, 1978, issue VK2ZTB states that an 88 mH toroid consists of two 44 mH coils. This is incorrect — an 88 mH toroid consists of two 22 mH coils connected in series. Thanks to N3GN for spotting the error.

30-MHz low-noise preamp

Coil winding instructions for the low-noise 30-MHz preamplifier in the October, 1978, issue, which were inadvertently left out of **fig. 1**, are as follows:

L1	(0.77 μH; 17 turns no. 28 (0.3 mm) wound on Micrometals T-25-6 powdered-iron toroid
L2	(1.0 μH) 20 turns no. 28 (0.3 mm) wound on Micrometals T-25-6 powdered-iron toroid
RFC	(10 μH) 20 turns no. 28 (0.3 mm) wound on FT-230-06 ferrite bead

active filters

The letter regarding active filter design (see *Comments, ham radio*, June, 1978, page 102) contained an erroneous equation. The value for R2 is determined from the following equation:

$$R2 = \frac{1}{(2 \cdot Q - (G/Q)) \cdot C \cdot f \cdot 2\pi}$$

spectrum analyzer filters

The two ceramic 10.7 MHz filters used in the spectrum analyzer described in the June, 1977, issue of *ham radio* (FL401 and FL501) are no longer available from Vernitron. Just before the production line was shut down, however, one of the employees was able to obtain 25 matched pairs which he is offering to readers who wish to build the spectrum analyzer; the price is $6.50 per pair, plus postage. Write to William Bowen, 1939 Green Road, Cleveland, Ohio 44121. The only other source for comparable ceramic filters is the Murata Corporation in Marietta, Georgia.

semi-precision voltage calibrator

The digital voltmeter calibrator on page 68 of the July, 1978, issue of *ham radio* contains an incorrect statement. The input resistance of the precision rectifier circuit is actually 10k ohms, instead of the value of between 5 and 65 megohms specified in the article. To correct this problem, **fig . 1** shows a simple voltage follower that can be inserted between S1 and the input to the precision rectifier. This will eliminate the loading effect on the reference voltage source and has negligible effect on the overall accuracy.

general purpose vhf receiver

A capacitor was deleted from the *schematic diagram shown in* **fig. 2** of the general purpose vhf receiver published in the July, 1978, issue of *ham radio* (see page 19). A 22-pF capacitor should be inserted between the gate of the MPF102 and the rotary contact of switch S1F.

R-4C product detector

In the October, 1978, issue of *ham radio*, the schematic diagram of the new product detector for the R-4C shown on page 94 contains an incorrect component value. The 0.01-μF capacitor in series with pin 3 of the TL442 should actually be a 0.1-μF capacitor. In addition, the two inputs for the TL442 are actually taken only across one-half of the transformer's secondary.

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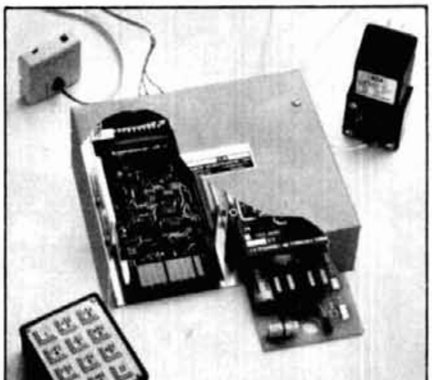

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The Logical Force™ is what Continental Specialties Corporation calls its line of digital test instruments, which includes their new model LP-1 hand-held digital logic probe.

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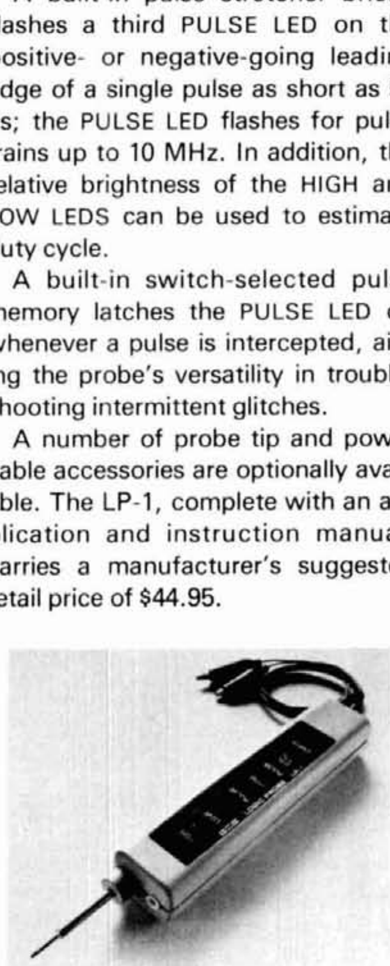
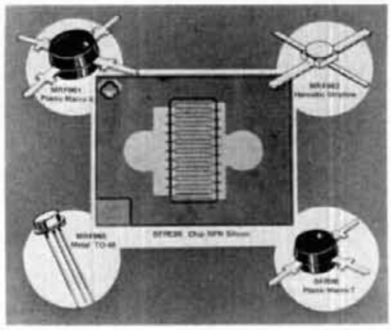


The Continental Specialties Corporation LP-1 Logic Probe is available at leading electronics dealers and distributors worldwide, or direct from factory. For further information, contact Continental Specialties Corporation, 70 Fulton Terrace, New Haven, Connecticut 06509.

Motorola microwave transistors

Motorola has introduced a new series of microwave transistors that combine state-of-the-art performance, a variety of packaging options, and a major breakthrough in pricing for the uhf market. Intended for low-to-medium-power amplification, the new transistor series features high gain (up to 15 dB at 0.5 GHz) and a very low noise figure (typically 2 dB at 0.5 GHz and 10 mA). Prices range from a low of $1.40 (in chip form) to a high of only $7.50 (hundred-up) for high-reliability applications.



The transistor series encompasses five device types, a basic chip and four package options, with optimized specifications and pricing for a wide variety of applications.

Available in unencapsulated form for hybrid application, the basic chips achieve their high-performance characteristics from fine-line geometry, ion-implanted arsenic emitters, and gold top metalization.

Gold top metalization prevents metal migration due to the high-current densities in the fine metal lines required for high-frequency operation. Ion implantation facilitates precise control of dopant densities and gradients. The use of arsenic dopant results in a higher f_T and corresponding improvement in noise figure, compared with the more conventional processing.

The high current, low noise figure, and high-f_T performance of the BFR96 series of transistors makes them eminently suitable for broadband vhf/uhf linear amplifier and oscillator applications.

Low-cost package options offer the basic chip in two types of plastic packages, the three-leaded MACRO-T package (BFR96), and the four-leaded MACRO-X package (MRF961). These are particularly well suited for MATV/CATV applications.

The MACRO-T package has become an industry standard and adapts the BFR96 to existing board layouts and designs. The four-leaded MACRO-X package offers a 2.5 dB higher gain, due to lower parasitics resulting from opposed-emitter lead construction, at no increase in price.

The metal/ceramic, hermetic stripline MRF962 package is intended for use in equipment that is subject to particularly hostile environmental conditions and when high reliability is required. This low parasitic microwave package enables the MRF962 to be specified for operation up to 2 GHz and gives typically a 6.0 dB higher power gain compared with the same die in a MACRO-T package at 500 MHz. The metal/ceramic package allows higher power dissipation than the plastic case.

The higher dissipation rating and hermiticity of the MRF965, in its TO-46 package, allows its use in high gain vhf/uhf Class C amplifier applications up to 400 milliwatts output power. This is in addition to the Class A linear applications discussed for the other package options.

An important feature of this series of devices is a four-part data sheet with common-emitter S-parameters at two levels of V_{CE}, three collector currents, and six frequencies from 100 MHz to 1500 MHz. More information on the BFR96 series may be obtained by writing to Motorola Semiconductors, Post Office Box 20912, Phoenix, Arizona 85036.

appointment clock

Hal-Tronix has announced the availability of *TimeTrac*, a micro-

More Details? CHECK—OFF Page 126

computer-controlled appointment clock. *TimeTrac* is both a clock and a reminder, and can be used in the ham shack, office, or home to fill a variety of needs. For example, it can be used as a simple timer up to 60 minutes 59 seconds, with pause; an alarm clock; an appointment reminder for the next 23 hours 59 minutes, and for future appointments for up to one year! The memory will hold up to 30 appointments, and, even if they are entered out of chronological order, *TimeTrac* will arrange them chronologically. Photographers, attorneys, hams, salesmen, secretaries, cooks, and many other persons will find *TimeTrac* useful — even indispensable. You can plug it into any wall outlet, set the illumination intensity, determine AM/PM by just a glance at the indicator, and even protect the memory against power outage with an on-board lithium cell. The vacuum-fluorescent display is one of the largest available today, and can easily be seen at a distance when needed. *TimeTrac* is complete, ready to go, not a kit, for the special introductory price of $69.95 from Hal-Tronix, Post Office Box 1101, Southgate, Michigan 48195.

free surplus catalog

Hobbyists, educators, technicians, and dealers alike are sure to be interested in the new *Surplus Electronics* catalog just published by Etco Electronics, because almost every item in it doesn't fit the usual concept of "surplus."

A look through the catalog will reveal an amazing variety of items, ranging from parts and components and test equipment to educational, industrial, and consumer equipment acquired from leading manufacturers.

The prices are always very low — in many cases only a small percentage of the normal price — because Etco's items come from surplus inventories, overstocks, and bankruptcies.

Get a copy of this fascinating cata-

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log today; it's free. Write Etco Electronics, Dept. 029, North Country Shopping Center, Route 9N, Plattsburgh, New York 12901.

compact power microphone

With today's generation of home-based communications operators having severe space limitations, Robins Industries has introduced a compact base-station power microphone with features usually found in larger and more expensive units.

This new power microphone is specially designed for Amateur and CB ssb transceivers. It incorporates a built-in, solid-state preamplifier to improve the modulation percentage figure of the transceiver. Low and high ends of the speech-frequency band are adequately attenuated to increase intelligibility. A full-width, press-to-talk bar switch, with positive locking mechanism, facilitates continuous operation. A minimum/maximum sliding gain control is clearly indicated.

The cardioid (uni-directional) polar pattern helps to screen out unwanted background noises when transmitting. Output level is rated at −36 dB at 1000 Hz, with an amplifier gain of 30 dB.

For further information, contact Charles Condike, Robins Industries Corp., 75 Austin Blvd., Commack, New York, 11725.

new bird amateur wattmeters

The Models 4360 and 4362 HAM-MATE directional wattmeters are designed specifically for the Amateur Radio service. The 4360 covers the 1.8-30-MHz range (200 and 2000 watts) and the 4362 is for use in the 140-180-MHz range, which includes the popular 2-meter band. 4362 has 25 and 250-watt scales. The design of the HAM-MATEs is basically that of all Bird THRULINE® rf wattmeters, except that they do not use plug-in elements. Instead, the sensing element is permanently installed in the line section and is rotatable from the front panel to provide the choice of reading either forward or reflecting power.

Both wattmeters are dual-range to allow measurement of both low and high power, and the meter reads directly in watts, with the high range being read on the upper arc and the low range on the lower arc. The down-scale portion of each range is expanded for easier reading. The average power output of CW, a-m, fm, and ssb transmitters can be measured with ease, and the wattmeters can be left in the line to allow continuous monitoring of the power output. The HAM-MATE wattmeters are especially useful during the tune-up of an Amateur transmitter.

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The new HAM-MATE wattmeters have directivity of 20 dB (100:1) minimum, an absolute must for meaningful reflected power (and vswr) measurement.

Bird HAM-MATE directional wattmeters are priced at $94.00 and delivery is from stock. Write Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139.

microprocessor controlled 2-meter FM transceiver

The TR-7600 2-meter fm mobile transceiver with memory and an optional RM-76 microprocessor control unit (which provides six memories and various scanning functions) have been introduced by Trio-Kenwood Communications, Inc., of Compton, California.

This new transceiver provides full 4-MHz coverage (800 channels) on 2 meters and includes a memory channel. It operates on simplex (same transmit and receive frequencies) or repeater (plus or minus 600-kHz transmitter offset) modes. Furthermore, the memory can be used to provide a transmit frequency for accessing a repeater with a nonstandard frequency pair. The TR-7600 also features a digital frequency display with large, bright-orange LEDs. Another LED, called an "unlock" indicator, shows transceiver protection when the frequency selector switches are improperly positioned or the PLL has malfunctioned. Selecting frequencies with the TR-7600 is fast and easy with its dual concentric knobs, 5-kHz offset switch, and MHz-selector switch.

Power output is switchable between 10 watts and 1 watt (adjustable from 1 to 10 watts). The TR-7600 is ideal for mobile use and comes with the MC-30S noise-cancelling dynamic microphone. The optional RM-76 Microprocessor Control unit allows the TR-7600 to perform many interesting new functions. Using the RM-76 keyboard, the Amateur can select any 2-meter frequency (including MARS on 143.95 MHz simplex), store frequencies in six memories, automatically scan up the band in 5-kHz steps, manually scan up or down in 5-kHz steps, set lower and upper scan frequency limits, scan for busy or open channel, reset scan to 144 MHz, stop scan, cancel scan (for transmitting), and select repeater mode. The RM—76's digital display indicates frequency (even while scanning) as well as functions (such as autoscan, lower scan frequency limit, upper scan limit, error, and call channel).

The TR-7600 will have a suggested list price of $375.00. For more information see your authorized dealer or contact Trio-Kenwood Communications, Inc., 1111 West Walnut Street, Compton, California 90220.

Micro Duster cleaning gas

Chemtronics Inc., a major manufacturer of chemical products, recently introduced Micro Duster, a new product that permits compressed-gas dusting of delicate instruments and assemblies. The product contains pure, moisture-free, nonflammable and nontoxic filtered gas, providing controlled removal of dust, lint, oxide particles, and the like without depositing harmful contaminants. Micro Duster has a broad range of applications, including mechanical and electrical miniature assemblies, audio components, computer tapes and heads, clean-room areas, timepieces, business machines, camera lenses and other optics, plus film, negatives, and slides.

A single 425-gram (15-oz.) can of Micro Duster produces over 1800 one-second compressed-gas bursts, or 25 to 30 minutes of continuous dusting. Spraying in short bursts until contaminants are dislodged is recommended for most efficient use. The product comes with an extension tube for pinpoint applications. For precision application in harder-to-reach areas, Micro Duster may be used with Chemtronics' Vibra-Jet attachment which provides extended range with a 30 cm (12-inch) rigid probe on the end of a 66-cm (26-inch) flexible tube.

The product, which contains 100 per cent laboratory controlled, guaranteed pure, inert gas, is available in 425-gram (15-oz.) cans with a suggested retail value of $2.50. It is sold only through Chemtronics distributors. For more information, including the location of local distributors, contact Chemtronics Inc., 45 Hoffman Avenue, Hauppauge, New York 11787.

outlet-strip catalog

Multiple outlet strips are useful when your circuit design calls for a safe method of adding hardware for testing and experimental work. SGL Waber Electric offers a catalog of 85 new UL-listed multiple outlet strips. The catalog also gives detailed descriptions and specs covering 240 multiple outlet strips and 15 wheeled carriers that are widely used in industrial, commercial, and military fields. Also given in the catalog are examples of typical operation, custom designs, and ordering information.

The B5 new UL listed strips use a unique mounting method and provide circuit-breaker protection against overloads and short circuits. All SGL Waber Electric multiple outlet strips are rated at 15 amperes, 125 volts ac, 60 Hz. That's 1875 watts, continuous duty. Write for the free catalog. Address your letter to SGL Waber Electric, 300 Harvard Ave., Westville, New Jersey 08093, or phone (609) 456-5400.

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AED Electronics scanner

Owners of the popular Yaesu FT-227R two-meter transceiver can now use their rigs in the scanning mode, as a result of the recent introduction of a new scanner by AED Electronics in Montreal, Quebec. The AED scanner is available either in kit form or prewired and tested, and makes use of the FT-227R digital readout to indicate frequency. The scanner kit is furnished with schematic, board layout diagram, and complete wiring and assembly instructions that use the well-known, step-by-step method of construction aids. Included with the kit are a double-sided glass epoxy circuit board, silk-screened for component location and predrilled to accommodate the component leads. Thirty-five components, including two ICs (with sockets) and a mini-toggle switch are furnished. The board can be wired in about three hours; it measures 1.3 × 10.2 cm (0.5 × 4 inches). It fits *inside* the case of the FT-227, alongside one of the chassis rails, and is connected into the circuit by eleven lead wires. Because of the CMOS circuitry, the scanner draws very little current. The scanner operates in the manner of a sampler, moving up a predetermined portion of the band in 10-kHz steps until it reaches the upper limit, whereupon it reverses and scans back down the band. When a signal appears, the scanner pauses for three seconds and then moves on, unless the operator toggles the mini-switch on the microphone case to lock the scanner to the busy channel. The operator may select any 1-MHz portion of the band to be scanned, and also may select frequency range and delay. Once the SCAN-OPERATE switch is released, the scanning function continues. A convenient feature is the ability of the scanner to be used with the Heathkit encoding microphone, or with the standard Yaesu mike that comes with the rig. The AED scanner kit sells for $34.95; the preassembled and tested version

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sells for $54. Please include $1.50 for shipping and handling. For more information on this and other products, write AED Electronics Ltd., 750 Lucerne Road, Montreal, Quebec, Canada.

Yaesu 6-meter transceiver

With the six-meter Amateur band heading into an era of world-wide DX similar to that enjoyed by hams in the 1950s, Yaesu Electronics announces the availability of their new all-mode, six-meter transceiver, the FT-625RD.

The transceiver offers USB, LSB, a-m, CW, and fm, with 25 watts ssb PEP, 25 watts fm or CW, and ten watts a-m output.

An rf speech processor is built in and a 600-Hz CW filter is available as an option, as is the memory storage unit, which allows recall of any frequency with just the flick of a switch. The memory unit is ideal for watching a beacon or calling frequency during marginal openings.

Digital readout is accurate to 0.1 kHz. Analog readout is the model FT-625R, at slightly less cost, is better than 1 kHz. VOX, PTT, semi break-in CW with sidetone, and a clarifier for receive or transceive are all included in the circuitry. For fm repeater use, the transceiver features standard ± 1 MHz repeater offset, programmable tone-burst encoder, squelch, and a discriminator zero-center meter. Alternative repeater splits may be accommodated through an optional crystal or the optional memory unit.

A built-in power supply accommodates all line voltages with taps for 100/110/117/200/220 and 234 Vac, 50/60 Hz, or DC voltages from 11.5 to 16 Vdc, negative ground at 5.7 amps transmit and 0.7 amps receive.

For a detailed, four-color brochure, see your nearby authorized Yaesu dealer or write to Yaesu Electronics Corporation, P.O. Box 498, Paramount, California 90723.

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REMOTE CONTROL TRANSMITTER & RECEIVER

- CAN BE USED AS REMOTE CONTROL FOR TV
- USE IT FOR YOUR OWN GARAGE DOOR OPENER
- TRANSMITS BETWEEN 38KHz AND 41.5KHz
- THOUSANDS OF USES FOR REMOTE CONTROL APPLICATIONS
- TRANSMITTER USES A 1.5V BATTERY
- SCHEMATIC INCLUDED

$19.95

Custom Cables & Jumpers

DB 25 Series Cables

Part No.	Cable Length	Connectors	Price
DB25P-4-P	4 Ft.	2-DP25P	$15.95 ea.
DB25P-4-S	4 Ft.	1-DP25P/1-25S	$16.95 ea.
DB25S-4-S	4 ft.	2-DP25S	$17.95 ea.

Dip Jumpers

DJ14-1	1 ft.	1-14 Pin	$1.59 ea.
DJ16-1	1 ft.	1-16 Pin	1.79 ea.
DJ24-1	1 ft.	1-24 Pin	2.79 ea.
DJ14-1-14	1 ft.	2-14 Pin	2.79 ea.
DJ16-1-16	1 ft.	2-16 Pin	3.19 ea.
DJ24-1-24	1 ft.	2-24 Pin	4.95 ea.

For Custom Cables & Jumpers, See JAMECO 1979 Catalog for Pricing

CONNECTORS

25 Pin-D Subminiature

DB25P (as pictured)	PLUG (Meets RS232)	$2.95
DB25S	SOCKET (Meets RS232)	$3.50
DB51226-1	Cable Cover for DB25P or DB25S	$1.75

PRINTED CIRCUIT EDGE-CARD

156 Spacing-Tin-Double Read-Out — Bifurcated Contacts — Fits .054 to .070 P.C. Cards

15/30	PINS (Solder Eyelet)	$1.95
18/36	PINS (Solder Eyelet)	$2.49
22/44	PINS (Solder Eyelet)	$2.95
50/100 (.100 Spacing)	PINS (Wire Wrap)	$6.95
50/100 (.125 Spacing)	PINS (Wire Wrap)	R681-1 $6.95

MINI-BUZZER

- Thousands of applications
- Operates on +6 to +9 VDC
- Output Frequency 800 Hz
- Draws only 15 MA @ 600 ohms
- Size: 1¼" x ⅜"

PART NO. MB-1

1-24 $1.95 ea
25-49 $1.75 ea
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SONALERT. AUDIBLE SIGNAL DEVICE

"Use as a warning Device or Audible Reminder"

- Turns on and off with low power transistor, SCR or IC
- Can be battery operated 6-28 volts
- Solid State - No moving parts
- Panel mounts in 1.25" round hole
- Black Plastic case-includes MTG nut
- **Operating Volts (DC):** 6 (Min) — 28 (Max)
- **Current MA:** 3 (Min) — 14 (Max)
- **dBA:** 68 (Min) — 80 (Max)
- **Frequency:** 2900Hz ± 500

#SC628
$7.50 each

AC Wall Transformer

Ideal for use with clocks, power supplies or any other type of AC application.

Part No.	Input	Output	Price
AC 250	117V/60Hz	12 VAC 250mA	$3.95
AC 500	117V/60Hz	12 VAC 500mA	$4.95

Regulated Power Supply

- Uses LM 309K
- Heat sink provided
- P.C. board construction
- Provides a solid 1 amp @ 5V
- Includes components, hardware and instructions
- Sizes: 3-1/2" x 5" x 2" high

JE200 $14.95

INSTRUMENT/CLOCK CASE

This case is an injection molded unit that is ideal for uses such as DVM, COUNTER, or CLOCK cases. It has dimensions of 4½" in length by 4" in width by 1-9/16" in height. It comes complete with a red bezel.

PART NO: IN-CC $3.49 each

MICROPROCESSOR COMPONENTS

8080A/8080A SUPPORT DEVICES		
8080A	CPU	$9.95
8212	8-Bit Input/Output	3.25
8214	Priority Interrupt Control	5.95
8216	Bi-Directional Bus Driver	3.49
8224	Clock Generator/Driver	3.95
8226	Bus Driver	3.49
8228	System Controller/Bus Driver	5.95
8238	System Controller	5.95
8251	Prog. Comm. I/O (USART)	7.95
8253	Prog. Interval Timer	14.95
8255	Prog. Periph. I/O (PPI)	9.95
8257	Prog. DMA Control	19.95
8259	Prog. Interrupt Control	19.95
6800/6800 SUPPORT DEVICES		
MC6800	MPU	$14.95
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MC6810API	128X8 Static Ram	5.95
MC6821	Periph. Inter. Adapt (MC6820)	7.49
MC6828	Priority Interrupt Controller	12.95
MC6830L8	1024X8 Bit ROM (MC68A30-8)	14.95
MC6850	Asynchronous Comm. Adapter	7.95
MC6852	Synchronous Serial Data Adapt	9.95
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MC6862	2400 bps Modulator	14.95
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Z80(780C)	CPU	$19.95
Z80A(780-1)	CPU	24.95
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2650	MPU	19.95
8035	8-Bit MPU w/clock, RAM, I/O lines	19.95
P8085	CPU	19.95
TMS9900JL	16-Bit MPU w/hardware multiply & divide	49.95
SHIFT REGISTERS		
MM500H	Dual 25 Bit Dynamic	$.50
MM503H	Dual 50 Bit Dynamic	.50
MM504H	Dual 16 Bit Static	.50
MM506H	Dual 100 Bit Static	.50
MM510H	Dual 64 Bit Accumulator	.50
MM5016H	500/512 Bit Dynamic	.89
2504T	1024 Dynamic	3.95
2518	Hex 32 Bit Static	4.95
2522	Dual 132 Bit Static	2.95
2524	512 Static	.99
2525	1024 Dynamic	2.95
2527	Dual 256 Bit Static	2.95
2528	Dual 250 Static	4.00
2529	Dual 240 Bit Static	4.00
2532	Quad 80 Bit Static	2.95
2533	1024 Static	2.95
3341	Fifo	6.95
74LS670	4X4 Register File (TriState)	1.95
UART'S		
A-Y-5-1013	30K BAUD	5.95

MICROPROCESSOR MANUALS			
M-Z80	User Manual		$7.50
M-CDP1802	User Manual		7.50
M-2650	User Manual		5.00
ROM'S			
2513(2140)	Character Generator(upper case)		$9.95
2513(3021)	Character Generator(lower case)		9.95
2516	Character Generator		10.95
MM5230N	2048-Bit Read Only Memory		1.95
RAM'S			
1101	256X1	Static	$1.49
1103	1024X1	Dynamic	.99
2101(8101)	256X4	Static	3.95
2102	1024X1	Static	1.75
21L02	1024X1	Static	1.95
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2112	256X4	Static MOS	4.95
2114	1024X4	Static 450ns	9.95
2114L	1024X4	Static 450ns low power	10.95
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2114L-3	1024X4	Static 300ns low power	11.95
5101	256X4	Static	7.95
5280/2107	4096X1	Dynamic	4.95
7489	16X4	Static	1.75
74S200	256X1	Static Tristate	4.95
93421	256X1	Static	2.95
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TMS4044-45NL	4K	Static	14.95
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TMS2516 (2716)	16K*	EPROM(Intel 2716) *Requires single +5V power supply	49.95
TMS2532	4KX8	EPROM	89.95
2708	8K	EPROM	10.95
2716 T.I.	16K**	EPROM	29.95
		**Requires 3 voltages, -5V, +5V, +12V	
5203	2048	FAMOS	14.95
6301-1(7611)	1024	Tristate Bipolar	3.49
6330-1(7602)	256	Open C Bipolar	2.95
82S23	32X8	Open Collector	3.95
82S115	4096	Bipolar	19.95
82S123	32X8	Tristate	3.95
74186	512	TTL Open Collector	9.95
74188	256	TTL Open Collector	3.95
74S287	1024	Static	2.95

POWERACE 101 -General purpose model for prototyping all types of circuit.
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POWERACE 102 -Complete digit prototyping lab with built-in logic probe
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ALL-CIRCUIT EVALUATORS WITH POWER

- 1680 solderless, plug-in tie points ... will hold up to 18 14-pin DIP's
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B+K PRECISION 3½-Digit Portable DMM

- Overload Protected
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Ranges: DC Voltage: 0-1000V
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Comes with test leads, operating manual and spare fuse.

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CONTINENTAL SPECIALTIES CORPORATION

- 20 Hz-100 MHz Range
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- Size — 1.75" x 7.38" x 5.63"
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ACCESSORIES FOR MAX 100:

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(New) CSC Mini-Max 6 Digit 50MHz Frequency Counter

- Guaranteed frequency range of 100 Hz to 50 MHz
- Full 6 digit display with antiglare window
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- Complete with mini antenna
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MINI-MAX $89.95

Accessories For Mini-Max

Part No.	Description	Price
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The Incredible "Pennywhistle 103"

$139.95 Kit Only

The Pennywhistle 103 is capable of recording data to and from audio tape without critical speed requirements for the recorder and it is able to communicate directly with another modem and terminal for telephone "hamming" and communications. In addition, it is free of critical adjustments and is built with non-precision, readily available parts.

Data Transmission Method	Frequency-Shift Keying, full-duplex (half-duplex selectable)
Maximum Data Rate	300 Baud
Data Format	Asynchronous Serial (return to mark level required between each character)
Receive Channel Frequencies	2025 Hz for space; 2225 Hz for mark
Transmit Channel Frequencies	Switch selectable: Low (normal) – 1070 space, 1270 mark; High – 025 space, 2225 mark
Receive Sensitivity	-46 dbm accoustically coupled
Transmit Level	-15 dbm nominal. Adjustable from -6 dbm to -20 dbm
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Power Requirements	120 VAC, single phase, 10 Watts
Physical	All components mount on a single 5" by 9" printed circuit board. All components included

Requires a VOM, Audio Oscillator, Frequency Counter and/or Oscilloscope to align

TRS-80 16K Conversion Kit

Expand your 4K TRS-80 System to 16K. Kit comes complete with:

* 8 each UPD416 (16K Dynamic Rams)
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TRS-16K $115.00

Special Offer - Order both your TRS-16K and the Sup'R' MOD II Interface kit together (retail value $144.95) **for only $139.95**

COMPUTER CASSETTES

- 6 EACH 15 MINUTE HIGH QUALITY C-15 CASSETTES
- PLASTIC CASE INCLUDED
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- ADDITIONAL CASSETTES AVAILABLE #C-15-$2.50 ea

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SUP 'R' MOD II

UHF Channel 33 TV Interface Unit Kit

Wide Band B/W or Color System

- * Converts TV to Video Display for home computers, CCTV camera, Apple II, works with Cromeco Dazzler, SOL-20, TRS-80, Challenger, etc.
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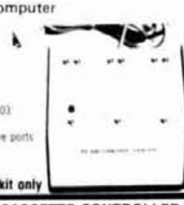
MOD II $29.95 Kit

RS-232 CONTROL CENTER

Plug in your modem, computer prom programmer, terminal, printer, etc. and selectively control data flow.



- Same Contour as "Pennywhistle 103"
- Totally self contained
- Includes 2 master ports and 3 slave ports

PART NO.
RS-232CC $89.95 kit only

CASSETTE CONTROLLER

Ideal for use with the TRS 80 and others.

"Plug/Jack interface to any computer system requiring remote control of cassette functions"

The CC100 controls cassette motor functions, monitors tape location with its internal speaker and requires no power. Eliminates the plugging and unplugging of cables during computer loading operation from cassette.

#CC-100 $29.50

63-Key Unencoded Keyboard

This is a 63-key, terminal keyboard newly manufactured by a large computer manufacturer. It is unencoded with SPST keys, unattached to any kind of PC board. A very solid molded plastic 13 x 4" base suits most application. IN STOCK **$29.95/each**

Hexadecimal Unencoded Keypad

19-key pad includes 1-10 keys, ABCDEF and 2 optional keys and a shift key. **$10.95/each**

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flea market

RATES Non-commercial ads 10¢ per word; commercial ads 60¢ per word both payable in advance. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing). Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (not all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to *corrected ad in next available issue*

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N.H. 03048.

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FOR SALE OR TRADE Excellent Kenwood TS520 transceiver and MC50 Mic. for good Atlas 210X or 215X with Atlas ac supply. George Konnick, Apt. C-5, 1750 West Main Street, Riverhead, New York 11901.

ANNOUNCEMENT: David R. Koberstein, 23 Winona Ct., Appleton, WI 54911, is the new Awards Manager of the Worked All Wisconsin Counties Awards sponsored by the Neenah and Minasha ARC (N-MARC). Cost $1.00 and should be accompanied by list of counties worked, stations, dates, etc. Please send SASE.

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More Details? CHECK—OFF Page 126

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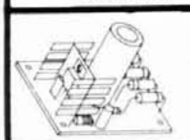

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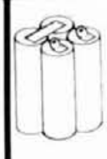

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CW OPERATORS — keyer paddle, model HK-1, brand new: $20.00. Will ship COD. See p. 114, May 1978 HR. Heath DX-60B XMTR $50. GE NBFM xcvr., 43 MHz, $60. M. Gordon, 280 Harvard St., #5C, Cambridge, MA 02139.

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FOR SALE: Measurements 80 Signal Generator, $275. Motorola T1021B Frequency Meter (Measurements 760) $250. Both in mint condition with manuals. W. J. Budzyna, K1BZI, 49 Gill Court, Whitinsville, MA 01588.

THE MEASUREMENT SHOP has used/reconditioned test equipment at sensible prices; catalog. 2 West 22nd St., Baltimore, MD 21218.

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FOR SALE: Gertsch FM-7 frequency meter and DM-3 deviation meter $675; Hewlett Packard 608B signal generator $450; Fluke 801 differential voltmeter $150; Tektronix 535 scope with CA plug-in $425. Lynn Price WB2HQX, 437 Payne Drive, Endicott, New York 13760, phone (607) 754-2194 between 6-8 pm.

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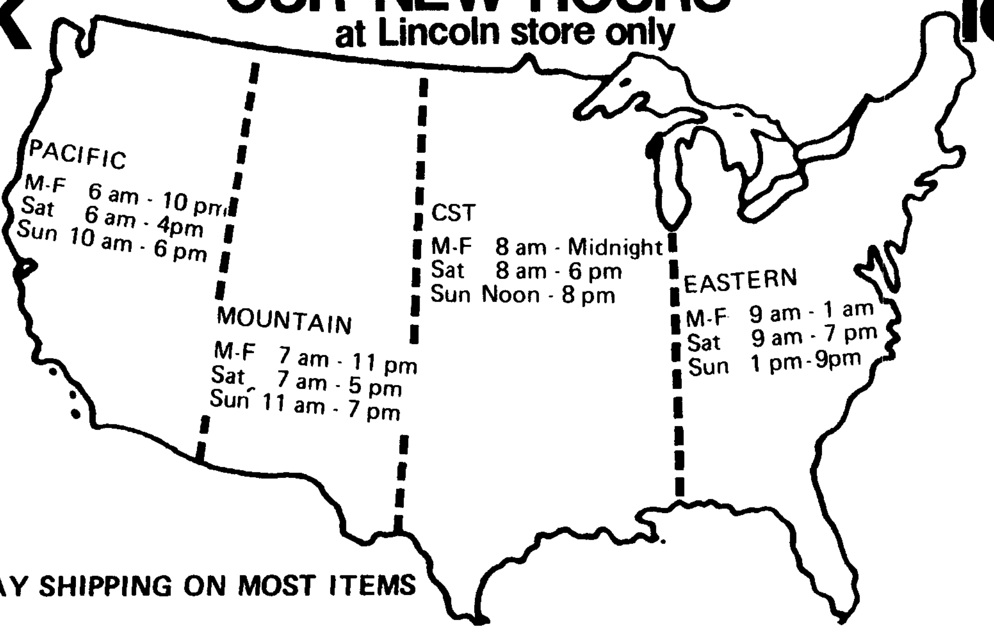

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Coming Events

MICHIGAN: The 9th Annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, February 25, 1979, 8:00 AM to 4:00 PM, (new) Churchill High School, Livonia, Michigan. Tables, door prizes, refreshments and free parking. Talk-in 146.52 Simplex. Reserved table space of 12-foot minimum available. Send SASE to Neil Coffin, WA8GWL, c/o Livonia Amateur Radio Club, P.O. Box 2111, Livonia, Michigan 48151.

RADIO EXPO '79 September 15th and 16th, 1979, Lake County Fair Grounds, Routes 120 and 45, Grays Lake, Illinois. Manufacturers' displays, flea market, seminars, ladies programs. Advance tickets $2.00. Write EXPO, P.O. Box 305, Maywood, IL 60153. Exhibitors inquiries: EXPO Hotline (312) 345-2525.

NEW HAMPSHIRE: QSO Party, sponsored by the Concord Brasspounders, Inc., W1OC, to promote Worked *New Hampshire award. Operating periods 2000Z* February 3rd to 0500Z February 4th, and 1400Z February 4th to 0200Z February 5th. Stations may be worked once per band per mode. New Hampshire stations may work each other, and will send RS(T) and county. Out-of-state stations send RS(T), ARRL section or country. Suggested frequencies: CW 1810, 3555, 7055, 14055, 21055, 28130; PHONE 1820, 3935, 3975, 7235, 14280, 21380, 28575; NOVICE 3730, 7130, 21130, 28130; VHF 50.115, 145.015, FM simplex (no repeaters). For additional information about rules, logs, and awards, contact Dale Clement, WA1FSZ, Bela View Drive, RFD #2, Concord, N. H. 03301. Send logs to: C. Halloway, 9 Via Tranquilla, Concord, N. H. 03301.

ROCHESTER Hamfest & NY State ARRL Convention, May 25-27. Add your name to mailing list. Send QSL to Rochester Hamfest, Box 1388, Rochester, NY 14603. Phone (716) 424-1100.

WORKSHOPS: a new and expanded series of Four 3 day hands on workshops on 8080/8085 Design, Microcomputer Interfacing, Software Design and Digital Electronics are being given by the authors of the popular Bugbooks. Participants have the option of retaining equipment used in these courses. Dates are March 19 to 28, *1979. For more information contact Dr. Linda Leffel,* C.E.C., Virginia Polytechnic Institute and State University, Blacksburg, Virginia 24061. (703) 961-5241.

WASHINGTON: Pacific Northwest Hamfest, July 14 & 15, HAM Inc., Box 78442, Seattle, WA 98178.

NEW HAMPSHIRE: 3rd Annual Interstate Repeater Society Auction and Hamfest, February 10 starting at *9:00 AM, Manchester Armory (just across the Amoskeag* Bridge from I-93). Bargains galore. Commercial exhibits. Free admission. Free parking. Talk-in on 146.52, 146.25/85 and 224.86/223.46.

ANNUAL QCWA MEMBERSHIP QSO CONTEST: sponsored by the Long Island Chapter, to be held on two separate weekends in 1979! CW: 0001 GMT Saturday, February 10th to 2400 GMT Sunday, February 11th. PHONE: 0001 GMT Saturday, March 10th to 2400 GMT Sunday, March 11th. Three global areas have been established for additional interest and point-scoring purposes, while the separate weekends will encourage greater participation. Keep an eye on QCWA NEWS for frequencies, confirmation texts, and related contest rules and guidelines.

More Details? CHECK — OFF Page 126

OHIO: The Mansfield Mid-Winter Hamfest/Auction February 11, 1979 at Richland County Fairgrounds, Mansfield, Ohio. Prizes, flea market, auction. Large heated building. Open 8:00 AM. Talk-in 146.34/.94 MHz. Tickets $1.50 in advance, $2.00 at the door. For information, advanced tickets, and tables contact Harry Frietchen, K8HF, 120 Homewood, Mansfield, Ohio 44906 or phone (419) 529-2801 or (419) 524-1441.

IOWA: Annual Davenport Radio Amateur Club Hamfest February 25, 1979 at the Masonic Temple in Davenport. Admission $2.00 advance, $2.50 at the door. Talk in on 28/88 and 52 simplex. Refreshments and tables available. For information and tickets, SASE to John S. Birmingham, WBØQCC, 2022 Brown St., Davenport, Iowa 52804.

PENNSYLVANIA: 7th Annual Lancaster Hamfest Sunday, Feb. 18, 1979 at the Guernsey Sales Pavillion, U.S. Rt. 30 & PA Rt. 896, Lancaster, PA. Doors open at 8:00 AM, prize drawing at 2:00 PM. Admission $3.00. Table reservation $2.00 advance. New, larger indoor flea market area. Food and soft drinks available. Talk-in 146.01/.61.

FLORIDA: The Martin County Amateur Radio Association's annual FIFTY CENT HAMFEST, Saturday, February 17, 8 AM to 4 PM. Tri-County Rehabilitation Center, 4461 So. Federal Hwy. (US 1) three miles south of Stuart. Admission: 50 cents. Swap tables: $3.00 ea. For table reservations or other info: MCARA, P.O. Box 1901, Stuart, FL 33494 or K4ZK, Hamfest Chairman, (305) 334-7418.

NEW JERSEY: Cherryville Repeater Assocation's annual Hamfest. March 17, 10 AM to 5 PM. Field House, Hunterdon Central High School, north of Flemington on Route 31. Over 200 tables, seminars, door prizes. Admission: $2.50 per person. For info: Sandford G. Franklin, Chairman, RD #3, Box 336, Milford, N. J. 08848.

TEXAS: Texas VHF/FM Society Winter Convention February 9, 10, 11, sponsored by Temple Amateur Radio Club, Inc. Activities include exhibits, transmitter hunt, communications programs, full ladies' activities plus Hotel facilities. FCC examinations. Tickets: Adv. $3.00 or $4.00 at the door. Grand Prize: Kenwood 700SP; Pre-registration prize; Transmitter Hunt prize; Hourly Door prizes. Full info and reservations for Ponderosa Motor Inn contact: Convention Committee, P.O. Box 4037 Temple, Texas 76501.

ILLINOIS: Sterling Rockfalls Amateur Radio Society's annual Hamfest, March 4, Sterling High School Fieldhouse, 1608 Fourth Ave., Sterling. $1.50 advance, $2.00 door. Large indoor Flea Market, free parking. Area for campers and mobile trailers. Commercial reservations only. Nearby shopping centers and theaters. For tickets: Don VanSant, WA9PBS, 1104 Fifth Avenue, Rockfalls, IL 61071. Make checks payable to Sterling Rockfalls ARS, please include SASE. Talk-in on 146.94 simplex.

VIRGINIA: Vienna Wireless Society's annual WINTERFEST, Sunday, February 25, Vienna Community Center. Indoor tables and sales, prizes, food. Frostbite tailgating: $1.00. Opens 6:30 AM for vendors; 8 AM for general admission. Admission: $3.00 (includes one prize ticket) $2.00 for extra ticket. Preteens with parents free. Tables $5.00 (not including admission fee) for first; $3.00 for second; $2.00 each over two, limit six. Reservations: Carroll N. Guin, 7533 Oak Glen Court, Falls Church, VA 22042. Information: Vienna Wireless Society, P.O. Box 418, Vienna, VA 22180. Reservations close February 15.

INDIANA: LaPorte ARC's winter Hamfest, Sunday, February 25, LaPorte Civic Auditorium. Tables $1.00. Donation: $2.00 gate. Talk in on .01/.61 and .52 simplex. For info: LARC, Box 30, LaPorte, IN 46350.

MISSISSIPPI: Old Natchez ARC Hamfest, Sunday, April 1, 1979 at the Natchez Convention Center. Indoors, airconditioned, FREE admission and swap tables. Talk-in on 146.31/.91 and 146.52 simplex. For information, write ONARC, 1226 Magnolia Avenue, Natchez, Mississippi 39120.

OHIO: Cuyahoga Falls Amateur Radio Club's 25th annual Electronic Equipment Auction and Flea Market, 9 AM to 4 PM, Sunday, February 25, North High School, Akron. Tickets: $2.00. Bring own tables, some available for $2.00 ea. Refreshments, prizes, Grand Prize — Triton IV. Tallmadge Avenue off-ramp, North Expressway (Rt. 8) to major interstates and Ohio Turnpike. Check in 146.52 simplex or 146.04/.64 repeater. Details from CFARC, P.O. Box 6, Cuyahoga Falls, OH 44222. Phone K8JSL (216) 923-3830.

MASSACHUSETTS: The Algonquin Amateur Radio Club's annual Electronic Flea Market, February 24, 10 AM to 4 PM, St. Mary's School Hall, Broad St., Marlboro. Easy access via I-495 to Rt. 20 East. Talk-in on 52 direct. Sellers contact: Charlie (W1BK) (617) 562-5622.

FLORIDA: The Treasure Coast Hamfest, March 17 and 18 at the Vero Beach Community Center. Admission $3 per family. Talk-in on 146.13/.73 and 222.34/223.94 repeaters or 146.52/.52 simplex. For information write P.O. Box 3088, Vero Beach, Florida 32960.

More Details? CHECK—OFF Page 126

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	1N4005	600v	1A	.08
	1N4007	1000v	1A	.15
	1N4148	75v	10mA	.05
	1N4733	5.1v	1 W Zener	.25
	1N753A	6.2v	500 mW Zener	.25
	1N758A	10v	"	.25
	1N759A	12v	"	.25
	1N5243	13v	"	.25
	1N5244B	14v	"	.25
	1N5245B	15v	"	.25

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	20-pin	pcb	.35	ww	.95
	22-pin	pcb	.35	ww	.95
	24-pin	pcb	.35	ww	.95
	28-pin	pcb	.45	ww	1.25
	40-pin	pcb	.50	ww	1.25
	Molex pins		.01	To-3 Sockets	.25
	2 Amp Bridge		100-prv		.95
	25 Amp Bridge		200-prv		1.50

TRANSISTORS, LEDS, etc.

QTY.			
	2N2222	(2N2222 Plastic .10)	.15
	2N2222A		.19
	2N2907A	PNP	.19
	2N3906	PNP (Plastic Unmarked)	.10
	2N3904	NPN (Plastic Unmarked)	.10
	2N3054	NPN	.45
	2N3055	NPN 15A 60v	.60
	T1P125	PNP Darlington	1.95
	LED Green, Red, Clear, Yellow		.15
	D.L.747	7 seg 5/8" High com-anode	1.95
	MAN72	7 seg com-anode (Red)	1.25
	MAN3610	7 seg com-anode (Orange)	1.25
	MAN82A	7 seg com-anode (Yellow)	1.25
	MAN74	7 seg com-cathode (Red)	1.50
	FND359	7 seg com-cathode (Red)	1.25

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QTY.			QTY.		
	9301	.85		9322	.65
	9309	.35		9601	.20
	9316	1.10		9602	.45

MICRO'S, RAMS, CPU'S, E-PROMS

QTY.			QTY.		
	8T13	1.50		2107B-4	4.95
	8T23	1.50		2114	9.50
	8T24	2.00		2513	6.25
	8T97	1.00		2708	10.50
	74S188	3.00		2716 D.S.	34.00
	1488	1.25		2716 (5v)	59.00
	1489	1.25		2758 (5v)	23.95
	1702A	4.50		3242	10.50
	AM 9050	4.00		4116	11.50
				6800	13.95
	MM 5314	3.00		6850	7.95
	MM 5316	3.50		8080	7.50
	MM 5387	3.50		8212	2.75
	MM 5369	2.95		8214	4.95
	TR 1602B	3.95		8216	3.50
	UPD 414	4.95		8224	3.25
	Z 80 A	22.50		8228	6.00
	Z 80	17.50		8251	7.50
	Z 80 PIO	10.50		8253	18.50
	2102	1.45		8255	8.50
	2102L	1.75		TMS 4044	9.95

C MOS

QTY.		
	4000	.15
	4001	.15
	4002	.20
	4004	3.95
	4006	.95
	4007	.20
	4008	.75
	4009	.35
	4010	.35
	4011	.20
	4012	.20
	4013	.40
	4014	.75
	4015	.75
	4016	.35
	4017	.75
	4018	.75
	4019	.35
	4020	.85
	4021	.75
	4022	.75
	4023	.20
	4024	.75
	4025	.20
	4026	1.95
	4027	.35
	4028	.75
	4029	1.15
	4030	.30
	4033	1.50
	4034	2.45
	4035	.75
	4037	1.80
	4040	.75
	4041	.69
	4042	.65
	4043	.50
	4044	.65
	4046	1.25
	4048	.95
	4049	.45
	4050	.45
	4052	.75
	4053	.75
	4066	.55
	4069/74C04	.35
	4071	.25
	4081	.30
	4082	.30
	4507	.95
	4511	.95
	4512	1.10
	4515	2.95
	4519	.85
	4522	1.10
	4526	.95
	4528	1.10
	4529	.95
	MC14409	14.50
	MC14419	4.85
	74C151	1.50

LINEARS, REGULATORS, etc.

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	MCT2	.95		LM323K	5.95		LM380 (8-14 Pin)	1.19
	8038	3.95		LM324	1.25		LM709 (8-14 Pin)	.35
	LM201	.75		LM339	.75		LM711	.45
	LM301	.45		7805 (340T5)	.95		LM723	.40
	LM308	.65		LM340T12	.95		LM725	2.50
	LM309H	.65		LM340T15	.95		LM739	1.50
	LM309K (340K-5)	1.50		LM340T18	.95		LM741 (8-14)	.35
	LM310	.85		LM340T24	.95		LM747	1.10
	LM311D	.75		LM340K12	1.25		LM1307	1.25
	LM318	1.75		LM340K15	1.25		LM1458	.65
	LM320H6	.79		LM340K18	1.25		LM3900	.50
	LM320H15	.79		LM340K24	1.25		LM75451	.65
	LM320H24	.79		LM373	2.95		NE555	.45
	7905 (LM320K5)	1.65		LM377	3.95		NE556	.85
	LM320K12	1.65		78L05	.75		NE565	.95
	LM320K24	1.65		78L12	.75		NE566	1.25
	LM320T5	1.65		78L15	.75		NE567	.95
	LM320T12	1.65		78M05	.75			
	LM320T15	1.65						

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QTY.			QTY.			QTY.			QTY.		
	7400	.10		7482	.75		74221	1.00		74LS02	.30
	7401	.15		7483	.75		74367	.95		74LS04	.30
	7402	.15		7485	.55		75108A	.35		74LS05	.35
	7403	.15		7486	.25		75491	.50		74LS08	.35
	7404	.10		7489	1.05		75492	.50		74LS09	.35
	7405	.25		7490	.45		74H00	.15		74LS10	.35
	7406	.25		7491	.70		74H01	.20		74LS11	.35
	7407	.55		7492	.45		74H04	.20		74LS20	.30
	7408	.15		7493	.35		74H05	.20		74LS21	.35
	7409	.15		7494	.75		74H08	.35		74LS22	.35
	7410	.15		7495	.60		74H10	.35		74LS32	.35
	7411	.25		7496	.80		74H11	.25		74LS37	.35
	7412	.25		74100	1.15		74H15	.45		74LS38	.45
	7413	.25		74107	.25		74H20	.25		74LS40	.40
	7414	.75		74121	.35		74H21	.25		74LS42	.75
	7416	.25		74122	.55		74H22	.40		74LS51	.45
	7417	.40		74123	.35		74H30	.20		74LS74	.45
	7420	.15		74125	.45		74H40	.25		74LS76	.50
	7426	.25		74126	.35		74H50	.25		74LS86	.45
	7427	.25		74132	.75		74H51	.25		74LS90	.65
	7430	.15		74141	.90		74H52	.15		74LS93	.65
	7432	.20		74150	.85		74H53	.25		74LS107	.50
	7437	.20		74151	.65		74H55	.20		74LS123	1.20
	7438	.20		74153	.75		74H72	.35		74LS151	.85
	7440	.20		74154	.95		74H74	.35		74LS153	.85
	7441	1.15		74156	.70		74H101	.75		74LS157	.85
	7442	.45		74157	.65		74H103	.55		74LS160	.95
	7443	.45		74161	.55		74H106	.95		74LS164	1.20
	7444	.45		74163	.85		74L00	.25		74LS193	1.05
	7445	.65		74164	.60		74L02	.20		74LS195	.95
	7446	.70		74165	1.10		74L03	.25		74LS244	1.70
	7447	.70		74166	1.25		74L04	.30		74LS367	.95
	7448	.50		74175	.80		74L10	.20		74LS368	.95
	7450	.25		74176	.85		74L20	.35		74S00	.35
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	7460	.40		74190	1.25		74L55	.65		74S05	.35
	7470	.45		74191	1.25		74L72	.45		74S08	.35
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	7474	.30		74194	.95		74L75	.85		74S20	.25
	7475	.35		74195	.95		74L93	.55		74S40	.20
	7476	.40		74196	.95		74L123	.85		74S50	.20
	7480	.55		74197	.95		74LS00	.30		74S51	.25
	7481	.75		74198	1.45		74LS01	.30		74S64	.15
										74S74	.35
										74S112	.60
										74S114	.65
										74S133	.40
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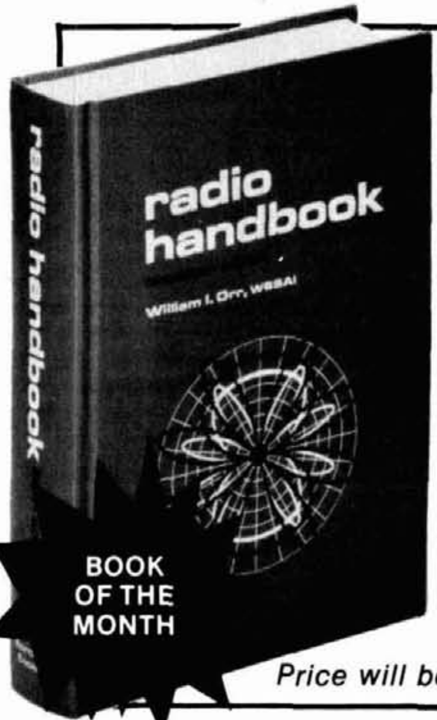

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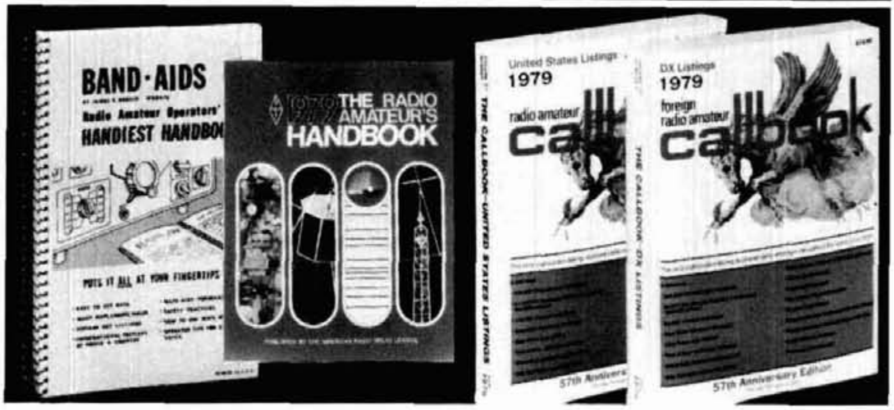

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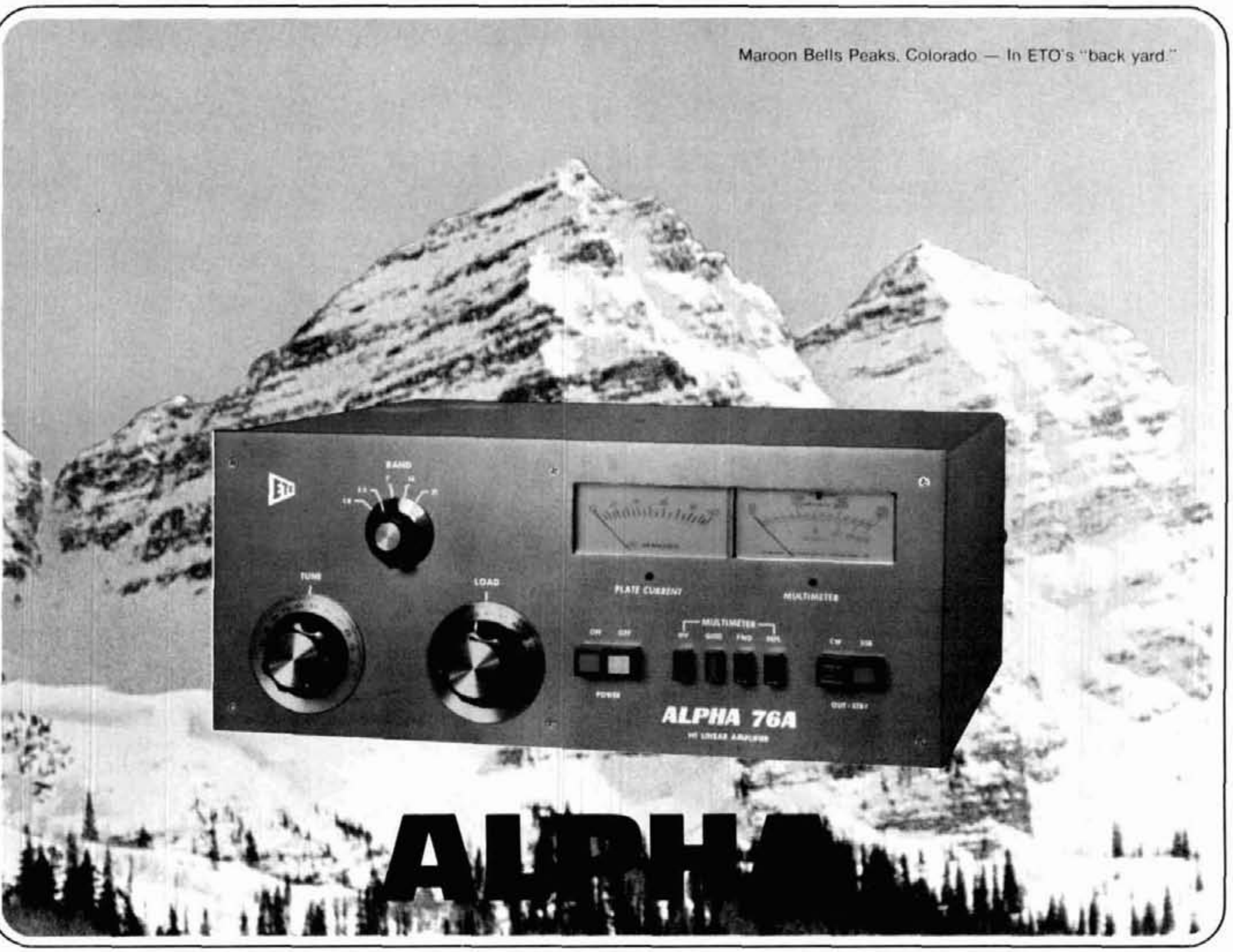

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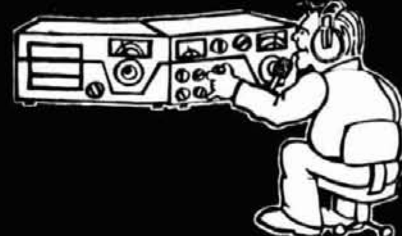

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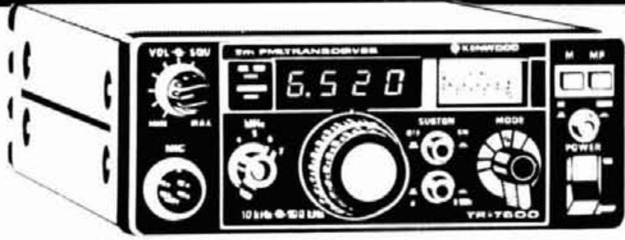

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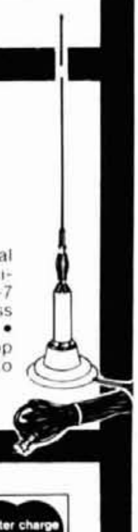

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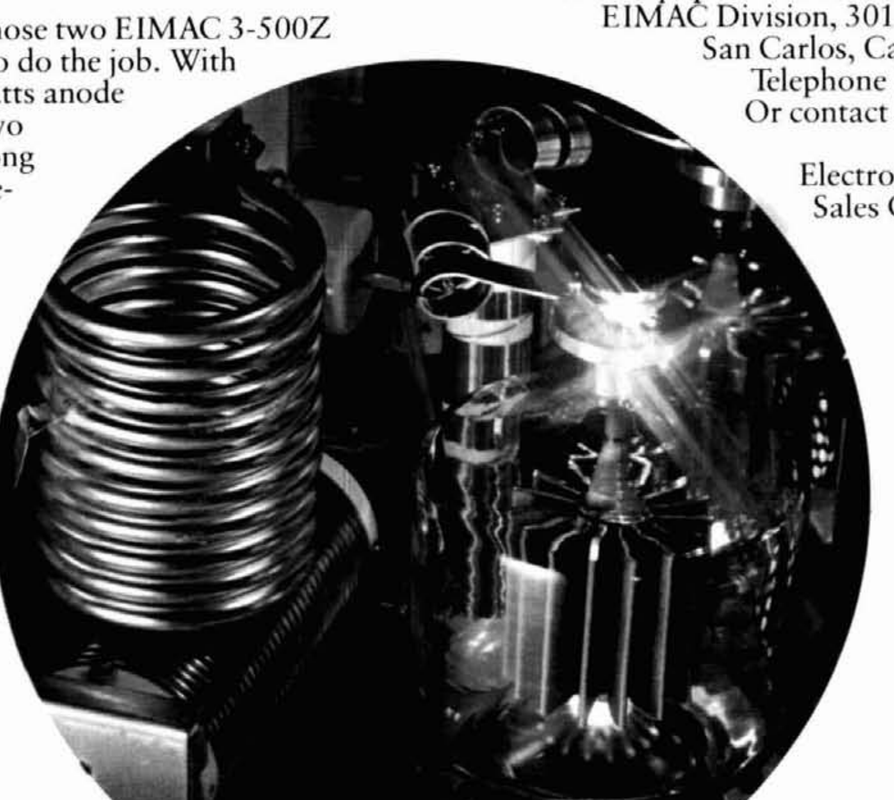

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