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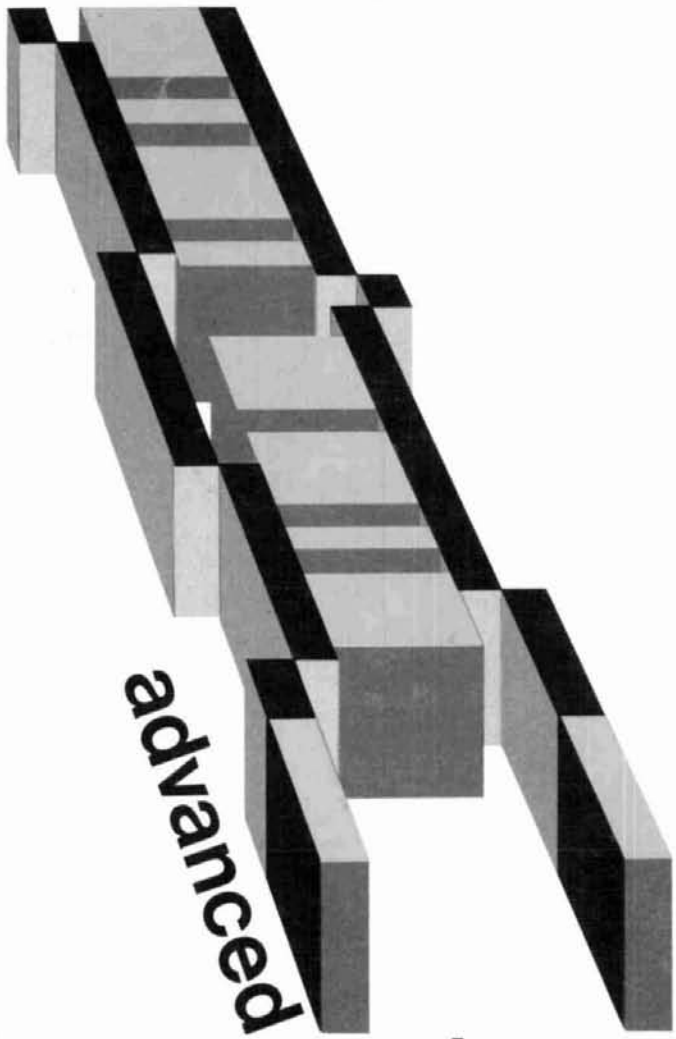
ham radio

magazine

APRIL 1978



advanced electronic keyer

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Tempo 130A10	10W	130W	$179.
Tempo 130A02	2W	130W	$199.
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Tempo 80A10	10W	80W	$139.
Tempo 80A02	2W	80W	$159.
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More Details? CHECK—OFF Page 126

This NEW MFJ Versa Tuner II . . .

has SWR and dual range wattmeter, antenna switch, efficient airwound inductor, built in balun. Up to 300 watts RF output. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines.

BRAND NEW

$79.95

Antenna matching capacitor. 208 pf. 1000 volt spacing.

Sets power range, 300 and 30 watts. Pull for SWR.

Meter reads SWR and RF watts in 2 ranges.

Efficient airwound inductor gives more watts out and less losses.

Transmitter matching capacitor. 208 pf. 1000 volt spacing.

Only MFJ gives you this MFJ-941 Versa Tuner II with all these features at this price:

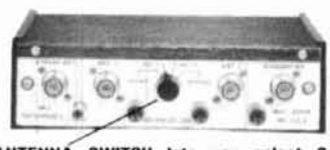
A SWR and dual range wattmeter (300 and 30 watts full scale) lets you measure RF power output for simplified tuning.

An antenna switch lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

A new efficient airwound inductor (12 positions) gives you less losses than a tapped toroid for more watts out.

A 1:4 balun for balance lines. 1000 volt capacitor spacing. Mounting brackets for mobile installations (not shown).

With the NEW MFJ Versa Tuner II you can run your full transceiver power output — up to 300 watts RF power output — and match your transmitter to **any** feedline from 160 thru 10 Meters whether you have coax cable, balance line, or random wire.



ANTENNA SWITCH lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

You can tune out the SWR on your dipole, inverted vee, random wire, vertical, mobile whip, beam, quad, or whatever you have.

You can even operate all bands with just one existing antenna. No need to put up separate antennas for each band.

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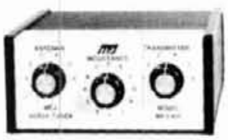

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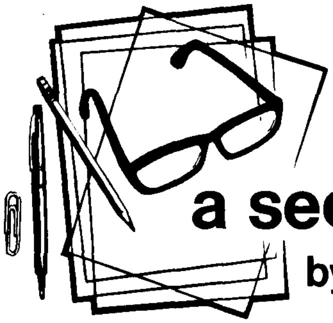

a second look

by Jim Fisk

If you look at technological advances with the eye of a historian, you'll see periodic peaks in new technology, followed by nulls where everything seems to come to a standstill. This phenomenon is most noticeable in the semiconductor industry during the past few years, but if you look back to the 19th century, you'll see the same thing happened then, with George Stephenson's steam-powered locomotives and Eli Whitney's cotton gin early in the century, and later with the development of mass production techniques, the electric telegraph, and Marconi's wireless. Every few years, it seems, a major breakthrough in technology occurs which produces an avalanche of new products on the market place. If you analyze it carefully, in most cases you'll find that the "major breakthrough" didn't happen all at once as it first appears, but was the culmination of years of research by many different workers in many diverse fields.

In many ways the cyclic rise and fall of technological achievement is an extension of the well-known domino theory — the apparent lulls in activity occur when the dominoes are being lined up; the breakthrough comes when the last domino is set in place and the whole line is knocked down, one domino after the other. It takes but one missing or misplaced domino to prevent the whole line from going down.

In the field of gallium-arsenide field-effect transistors (GaAs fets or "gas" fets, for short), the last of the dominoes has been set in place, and in the near future there will be devices on the market which will provide noise figures of less than 1 dB at 4000 MHz, and power outputs of 6 watts or more at 10 GHz. At a conference last summer at Cornell University, researchers from Bell Labs reported on a GaAs fet amplifier which yielded 10 dB gain at 4 GHz with a noise figure of 0.7 dB; at 6 GHz the noise figure increased to 1.25 dB and gain dropped to 9 dB. At the same conference, scientists from Rockwell International were talking about a device which gave a 2.2 dB noise figure at 10 GHz with 8.5 dB gain. These devices and similar ones from other manufacturers will be available on the commercial market this year, but be prepared to pay a pretty healthy price for the privilege of building a circuit around them. As manufacturing processes are improved and yields go up, however, I expect device cost will drop within the amateur price range. If you can't wait, in this issue JH1BRY describes a high-performance GaAs fet preamp for 432 MHz that has a noise figure below 1 dB.

At Texas Instruments the accent is on power GaAs fets with claimed power output of 4.2 watts at 10 GHz; and RCA has announced 140 mW output at 22 GHz. The most impressive of the power GaAs fets, however, is a device from Fujitsu which provides 10 watts output at 4000 MHz. In theory 14 or 15 watts should be possible at 4 GHz with GaAs fet devices, and 6-7 watts at 10 GHz, but there are still a number of problems to be solved, so this capability may be five years in the future.

Some of the same technology that has produced high-performance microwave GaAs fets is also being used in other areas. At the International Solid State Conference in February, an engineer from Hewlett-Packard described a monolithic 4-GHz integrated amplifier which provides 28 dB gain from dc to 2000 MHz (gain is 7 dB from dc to 4 GHz). The three-stage amplifier is built on a single GaAs chip and uses reverse biased Schottky diodes as capacitors and MESFETs as resistors. At the same conference, Bell Labs reported on a uhf operational amplifier which has a unity gain frequency in excess of 1000 MHz. Silicon NPN transistors are used in the design, which provides 20 dB gain at 300 MHz; the operational frequency range is 10-500 MHz.

While GaAs fets threaten to displace bipolars above 3000 MHz, silicon MOSFETs are quietly moving in on microwave bipolars from the lower end of the frequency spectrum. Attention has been focused recently on advances in VMOS devices for vhf and uhf applications. Since the first commercial VMOS device was announced by Siliconix two years ago (*ham radio*, September, 1976), a number of semiconductor manufacturers have gotten on the VMOS bandwagon including Intersil, Motorola, and Westinghouse. One firm has reportedly obtained 10 watts at 1000 MHz and more than 5 watts at 1500 MHz with a VMOS transistor. One of the big advantages of VMOS is its negative temperature coefficient which eliminates thermal runaway and secondary breakdown. This means that emitter resistors, temperature sensing diodes, and other protection circuitry required for bipolars can be eliminated, resulting in a substantial cost savings. VMOS also offers better linearity with third-order IMD typically 3 to 5 dB better than bipolar power amplifiers.

As I said earlier, technological breakthroughs seem to come in spurts. During the past year or so there have been some remarkable achievements in the world of digital electronics, but it has been pretty quiet on the analog front. With the latest rf devices just now emerging from the laboratory, how long will we have to wait before all the dominoes are gathered for the next breakthrough?

Jim Fisk, W1HR
editor-in-chief

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ICOM's matchless mobile **IC-22S** has, since its introduction, been the standard against which all other VHF mobile transceivers have been measured. It is a prime example of ICOM's quality engineering, peak performance and phenomenal flexibility blended into one splendidly simple and affordable radio. Small wonder that the **IC-22S** has been the most popular radio that ICOM has ever offered.

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presstop

TEN-METER AMATEUR LINEARS WERE BANNED by the FCC Commissioners in an early February meeting, and it seemed evident that most had already decided on the ban before the meeting began. The ban covers the commercial manufacture, distribution and sale of any RF power amplifier covering the 24-35 MHz range. It appears that linears in the marketing "pipeline" — in manufacturers' stock or on dealers' shelves — will receive some sort of "blanket" waiver to be sold providing that they meet all the Type Acceptance criteria other than frequency coverage. Any that don't will have to be considered individually. Amplifier sales between individual Amateurs are still permitted, and individual Amateurs are still permitted to build their own 10-meter linears.

LIMITED TYPE ACCEPTANCE of Amateur amplifiers below 144 MHz was also adopted with technical standards similar to those already in Part 97; actual implementation of Type Acceptance won't take place until detailed specifications and procedures are ready in several months.

To Meet The New Type Acceptance requirements an amplifier will have to be capable of accepting 50 watts (mean power) of drive, have less than 15 dB power gain, require external transmit-receive switching (no RF sensing) — and, of course, not operate in the 24-35 MHz spectrum. How difficult a manufacturer is supposed to make it for an individual Amateur to modify his product to cover 10 meters is still unclear.

It's Expected that the ban and Type Acceptance will go into effect at the same time, but just when won't be known until a Report and Order on the two dockets is released. Further clouding the issue is what action industry plans, with several manufacturers, ARMA, and the ARRL all likely candidates for filing petitions for reconsideration. In both industry, and among individual Amateurs, the feeling is that the Type Acceptance requirements in themselves would have been sufficient to drive the illegitimate amplifiers off the market, so the ban on top of Type Acceptance serves only to unnecessarily penalize Amateurs.

AMATEUR SECONDARY STATION licenses for individuals will be phased out as their present terms expire as a result of the first Report and Order on Docket 21135, effective March 24. Special events stations and callsigns were also eliminated as the Commissioners attempted to reduce the Amateur Radio workload at Gettysburg to within budgetary limitations. For Amateurs who presently hold more than one station license, the effect will be a decision on their part as to which callsign to keep when it is time to renew; the freeze on issuance of new secondary licenses, in effect for almost a year, is now permanent.

Extra Class Requests for specific 1x2 or 2x2 callsigns will no longer be honored under the newly adopted rules, though they will still be able to request an unspecified 2-letter-suffixed callsign. Eventually, but well in the future, it's planned to offer 1x3 callsigns to all the holders of Advanced or Extra Class licenses who wish them. Also, the present process by which a 1x3 callsign holder receives a 1x2 "preferred" callsign when changing call areas is to be eliminated. However, as the rules now permit an Amateur to retain his former callsign even after moving permanently to a new call area, it's likely many present 1x3 and 1x2 callsign holders will elect that option. Along that line, future initial callsign assignments are now going to be based on the applicant's mailing address rather than station location.

Club, RACES, and Military Amateur station licenses are to be further considered under a new Notice of Proposed Rule Making on Docket 21135. Distinctive prefixes for club stations (WK) and military stations (WM) are also proposed in that NPRM, while RACES stations would continue with WC callsigns as at present. It's also proposed that the criteria for a club station license be tightened considerably, requiring that the organization demonstrate a "compelling need" (distinctive use, past and present) for a club license before one would be issued or renewed. RACES licenses would also be reduced to only one per Civil Defense organization, and licenses for all three classes of organizazations would be issued to the organization, not an individual trustee. Due date for Comments on this NPRM is June 2, with Reply Comments June 30.

REPEATER LICENSES AREN'T REQUIRED to put a repeater on the air under terms of a recent waiver. No prior FCC approval or additional license is now required to operate a repeater, auxiliary link, control, or remotely controlled Amateur station — all may be operated with the individual's primary station license, callsign, and appropriate ID. The waiver is effective until the Commission releases an Order on Docket 21033.

KH1 THROUGH KHØ are the new prefixes for U.S. Pacific islands, while KP1-KPØ will identify Caribbean areas; the change was effective March 24. As some spots such as KX6 and KC6 are not FCC administered and would apparently continue with present prefixes, there seems to be enough in each block to handle the needs.

THE 2-METER DX RECORD has been stretched again, this time to 6300 km (3940 miles), as KP4EOR worked LU8DIN. YV5ZZ has heard LU3AAT on 432.1 MHz, and they plus KP4EOR are attempting schedules on that band as well as on 2 meters.

The TR-7500 is a 100 channel PLL synthesized 146-148 MHz mobile transceiver offering the dependability you've come to expect from Kenwood products.

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TR-7500

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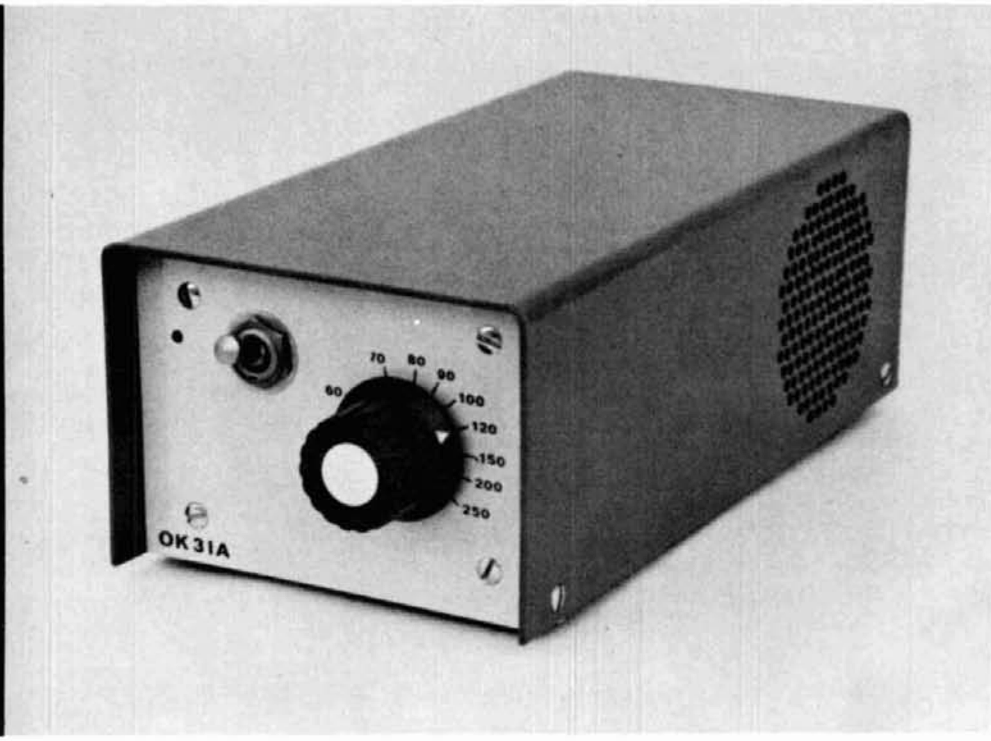


advanced electronic keyer

By defining keying intervals the author has developed a practical keyer that permits maximum time for character keying — enabling you to send letter-perfect code.

About thirty years ago, the bug was the most popular keying device, being widely used by hams all over the world. Only a small minority were using a new device, the electronic keyer, which had just begun to appear on CW bands, to produce fast and perfect machine keying. In the late 1940s, W6OWP described in *QST*[1] a new and simple principle for the electronic keyer. In Europe, this keyer was modified and popularized on the amateur bands by OZ7BO.

The W6OWP electronic keyer was simple, yet reliable. It did not use any clock-pulse generator, with the timing of dots, dashes and spaces performed by RC circuits, two triode vacuum tubes, and two relays. The dot, dash, and space time ratio was practically constant over the whole speed range. The most important features, though, were the self completing dots and dashes (including the following space), instant response when the paddle was closed, and sufficient time to release the paddle when the character generation was to be stopped.

electronic keyer requirements

Are there new requirements for the electronic keyer today? The optimum timing of characters, (dots, dashes, and spaces) over the entire speed

By Pavel Horvath, OK3IA, Institute of Physics, Slovak Academy of Sciences, 899 30 Bratislava, Czechoslovakia

range, and simple but reliable control of the keyer with the paddles are, of course, still expected. This can be easily understood by examining the timing of characters as shown in **fig. 1**. The time durations of the dash, dot, space within a character, space between characters, and space between words are defined as 3, 1, 1, 3, and 7 elementary time intervals. This timing of characters seems to be optimal for fast speeds. A different timing could be required at very low speeds; this problem, however, is not discussed here. Even though the keyer must accept imperfect keying, the output characters should be generated without errors.

Control of the keyer with the paddles is very important. Two types of paddles are generally used; the single paddle for standard keying and the dual paddle

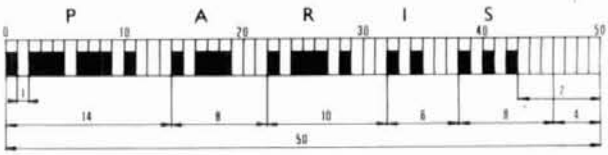


fig. 1. Timing of characters and spaces.

for squeeze or iambic keying. The single paddle, when released, slips into the neutral position and can be closed for a dot or for a dash. The dual paddle can be closed simultaneously for a dot and for a dash. Keying with a dual paddle is somewhat different than that with a single one.

Using the single paddle, there are two distinct groups of characters. First, those characters having an alternate sequence of dots and dashes (C for example) are generated by alternatively closing the

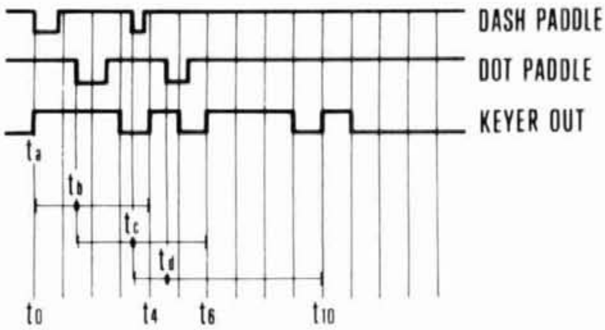
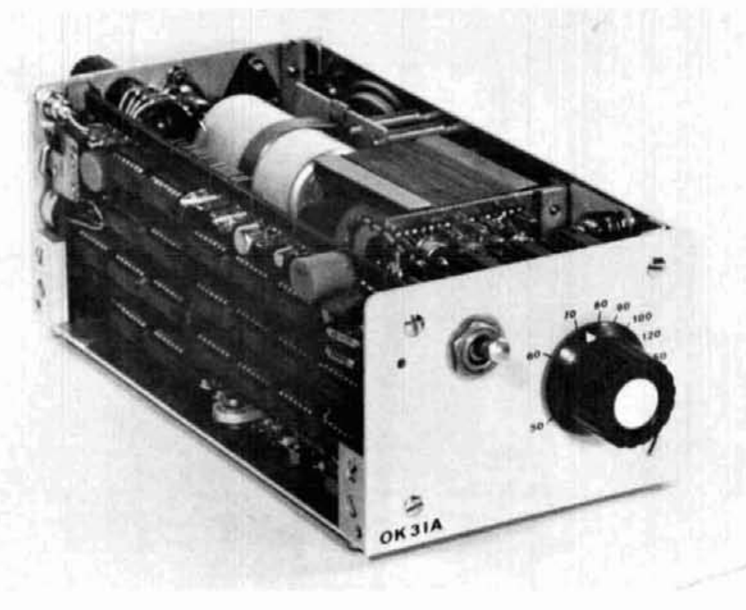


fig. 2. Permitted time intervals for closing the paddles to generate the character C.

paddle for a dash and then a dot. The first dash starts at the instant $t_a = t_\emptyset$, when the paddle is closed. No delay time should occur. By using dot and dash memories, the paddle can be moved to the dot side after the first dash begins. **Fig. 2** shows the permitted time intervals for closing the paddle to generate the character C without errors.

The second group of characters are those which have a series of dots or dashes. **Fig. 3** shows the keying required for the character = (BT). The paddle is first closed to generate a dash at the instant $t_a = t_\emptyset$, but yet, must be released and moved to the dot side before t_4. During the interval t_c to t_{10}, the paddle must be repositioned to form the last dash. For example, the paddle could be closed during t_c to t_{10} and released during t_d to t_{14}.

A somewhat different situation arises if you use dual paddles for iambic keying. The output with both paddles closed is called an iambic sequence. **Fig. 4** shows the keying for CQ. You start by closing the dash paddle, followed by closing the dot paddle. The



Front view showing the construction and placement of the printed circuit boards, power supply, speaker, and other components.

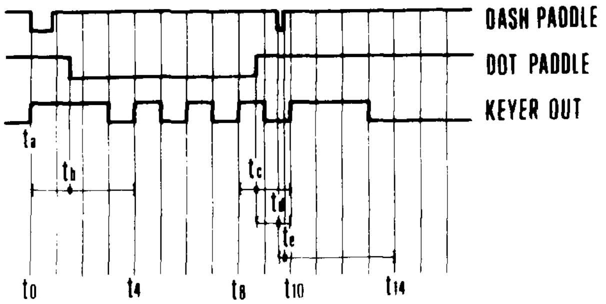


fig. 3. Permitted time intervals for closing and releasing paddles to generate the character = (BT).

dot paddle can be closed any time during the t_0 to t_4 interval. Both paddles are held until you recognize the last dot in the character C. Then you release both paddles. The sequence for releasing the paddles is not important, but both must be released during t_{10} to t_{12}. If the keyer also completes the spaces between characters, you can re-close the dash paddle anytime during the t_{12} to t_{14} interval. After the second dash in the Q has started (t_{18}), you must close the paddle for a dot during t_{18} to t_{22}, keeping both paddles closed to produce the character in the iambic mode. If both paddles are released anytime during t_{24} to t_{28}, the Q will be completed. When both paddles are released, the iambic sequence must stop after one bit (dash or dot) is completed. No second bit should be generated.

The permitted time intervals previously discussed are the maximum intervals in which the paddles can be closed or released, allowing the characters to be generated without errors. Any shorter time intervals make the keying worse. A summation of the previous points produces a list of features which an electronic keyer should incorporate:

1. Keying by both single and dual paddles.

2. Maximum possible permitted time intervals to make the keying easy.

3. At the beginning of the message, the character should start the instant the paddle is closed.

4. Optimum timing of dots, dashes, and spaces, throughout the entire speed range.

5. Stable clock generator, without clock pulse variations after triggering.

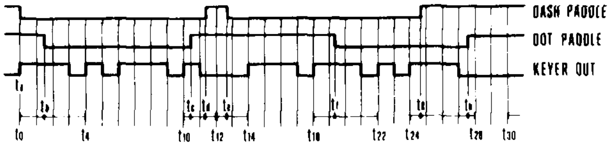


fig. 4. Keying of the word CQ in the iambic mode.

I began my design with the circuit described by WB2DFA.[2] In my version, I use very similar timing circuits, modified for a triggered clock and complete timing of spaces. My efforts were concentrated on improving the control of the keyer with paddles, making keying easy and convenient.

A block diagram of this keyer, with the significant signal names, is shown in **fig. 5**.

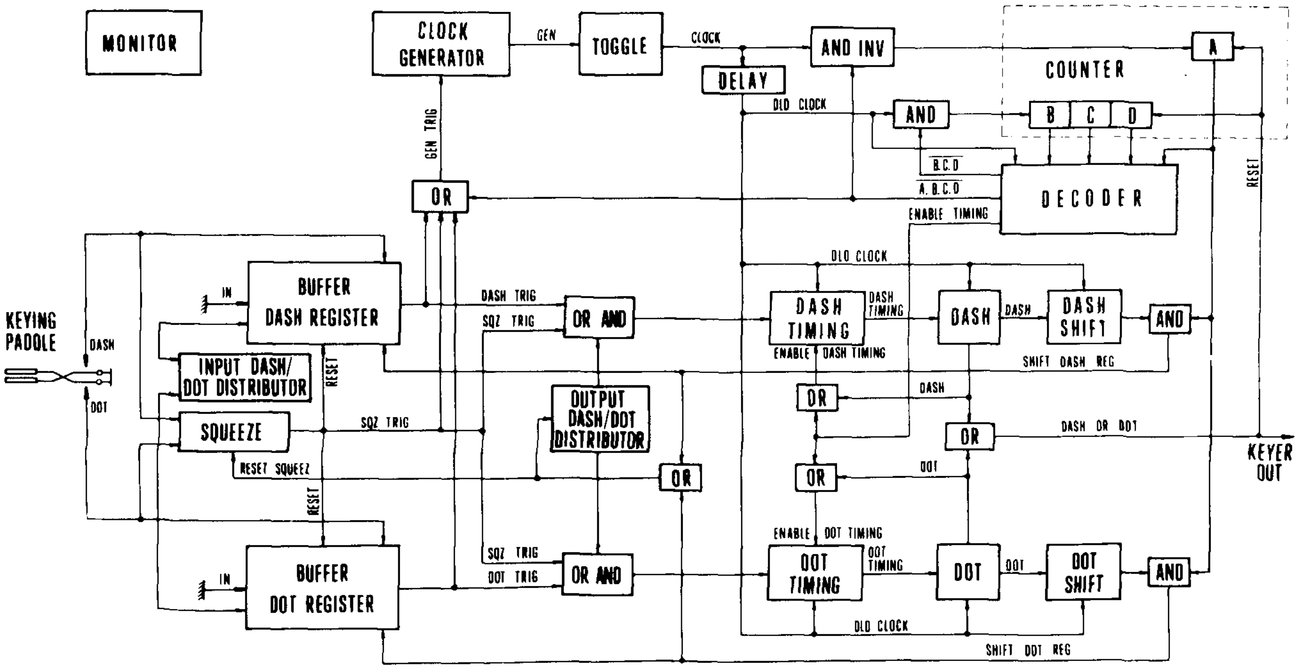


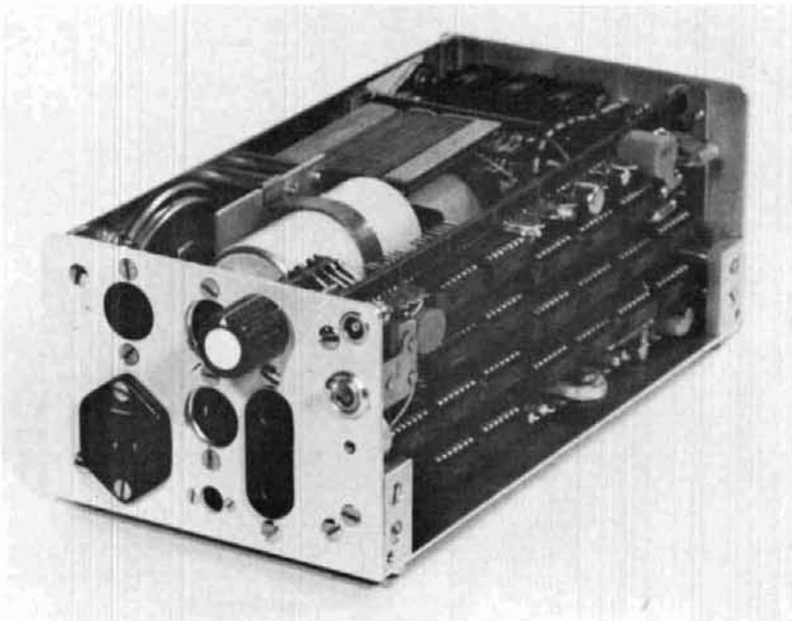
fig. 5. Block diagram of the advanced electronic keyer showing the most significant signals.

Clock generator. As shown in **fig. 6**, the schematic diagram, the clock pulses are generated by transistors T_4, T_5, and the associated logic. In the initial state, the clock generator is off since GEN TRIG is low (0). The signal GEN TRIG is the logical sum of DASH TRIG, DOT TRIG, SQZ TRIG, MESSAGE START, and A•B•C•D. Because of the low from GEN TRIG, T4 is turned on by the current flowing through D3 and subsequently T5 is also turned on. At the instant the paddle is closed, GEN TRIG goes high, enabling the clock generator. Simultaneously, a short, negative pulse is formed by the gates connected to D4. This negative pulse turns off T4 and starts the generation of clock pulses. This circuit philosophy allows the generator to start in synchronization with the paddle. The clock pulses (GEN) are perfectly formed, the first pulse being the same as the following ones. The clock frequency is continuously variable and corresponds to speeds of 10 to 50 words per minute. The toggle and delay circuits divide the GEN signal by two, producing CLOCK and DLD CLOCK signals, with a delay of about 200 nS.

Space generation timing circuits. The circuitry in this portion of the keyer is used to provide a constant-width space following each element of a character, whether a dot or a dash. This self-completing space corresponds to the length of one dot. In addition, the timing of the space between characters and words can also be made self-completing, corresponding to 3 and 7 elementary-time intervals.

In the initial state (completion of a space), the four outputs of the binary counter are all high, causing the ENABLE TIMING line to also be high. When this line is high, the circuits for generating a dot or dash are not inhibited, and can be activated by pressing either paddle. Now, if a dash is to be sent, for example, the paddle is pressed, causing the clock generator to be activated, and the DASH signal to be generated. This signal, DASH, is the actual eventual output from the keyer. And, in addition to being the output, it is used to reset the binary counter for the duration of the signal. When DASH resets the counter, through the RO(1) and RO(2) inputs, the B•C•D and A•B•C•D are high, thus feeding the clock pulses to the binary counter. But, the pulses will not be counted until the completion of the dash since the resets override the normal clock input.

When the dash is completed, the CLOCK and DLD CLOCK pulses will then be counted. The gates connected to the B and D outputs of the binary counter will give an ENABLE TIMING pulse at the completion of the first or third timing interval; if there is a dot or dash waiting in the buffers, it will be generated after the first interval and the process will start again. If a dot or dash is not waiting, another ENABLE TIMING pulse will occur after the third timing interval (character space). If a character is still not in the input buffers, the word space will be generated (seven time intervals) by the gates connected to the B, C, and D outputs of the counter. At the completion of the word space, the A•B•C•D line will go low, stopping the pulse generator. If a character had been keyed in as the beginning of a new word, however, the pulse generator will not stop. Thus, if you are keying properly, the clock generator runs without interruption through the whole message.



Rear panel view showing connectors for the ac line, 12-volt battery, paddles, manual key, external speaker, message memory, keyer output, S1 and S2, and speaker volume control.

Buffers and registers. The input buffers (or dot and dash memories) are used to store the code elements as they are put into the keyer. The dash and dot memories are actually 2-bit shift registers set up as FIFO registers, composed of D-type flip-flops. During keying, the dots and dashes are not only stored, but are also simultaneously shifted into the timing circuits. Therefore, it is possible to buffer characters in which the sum of the dots and dashes is greater than four.

Dash and dot distributors. The input and output dash/dot distributors remember which kind of a bit, dash or dot, was stored first. The input distributor stores the dots and dashes in a sequence depending on the first bit keyed in; the output distributor correctly loads the content of the buffers into the timing circuits.

The character C, if stored in the buffers, is processed to the keyer output in the following manner. The output dash/dot distributor, being enabled for a dash, allows the DASH TRIG signal to pass through the gates into the dash timing circuit, triggering the first dash. When the dash and the following single width space are completed, the signal SHIFT DASH REG shifts the dash register and clocks the output distributor to the dot memory, thus enabling the DOT TRIG signal through the gates into the dot timing circuit, triggering the dot. When the dot and the single width space are completed, the signal SHIFT DOT REG shifts the dot register, turning the output distributor back to the dash buffers. The same procedure is followed until the character C is shifted out of the buffers.

Dash and dot timing. The timing of a dash will start if the DASH TRIG or SQZ TRIG signal is low, the output dash/dot distributor enabled for a dash, and the signal ENABLE DASH TIMING high. The pulses, DLD CLOCK, transfer the high level on the D input to the output of the flip-flop, causing DASH TIMING to go high. With this signal high, a four-state binary counter receives the DLD CLOCK pulses. The counter states are then decoded into the signal DASH, which is three elementary-time intervals long. While the DASH is executed, the DOT circuit is disabled.

Dash shift and dot shift. The signals SHIFT DASH REG and SHIFT DOT REG, shown in **fig. 7**, are short pulses positioned at the end of the single width space following the dash or the dot. By having these pulses at the end of the space, you have the maximum amount of time in which to key in another dot or dash.

The block diagram in **fig. 5** illustrates how these signals are formed. The signal CLOCK is delayed 200

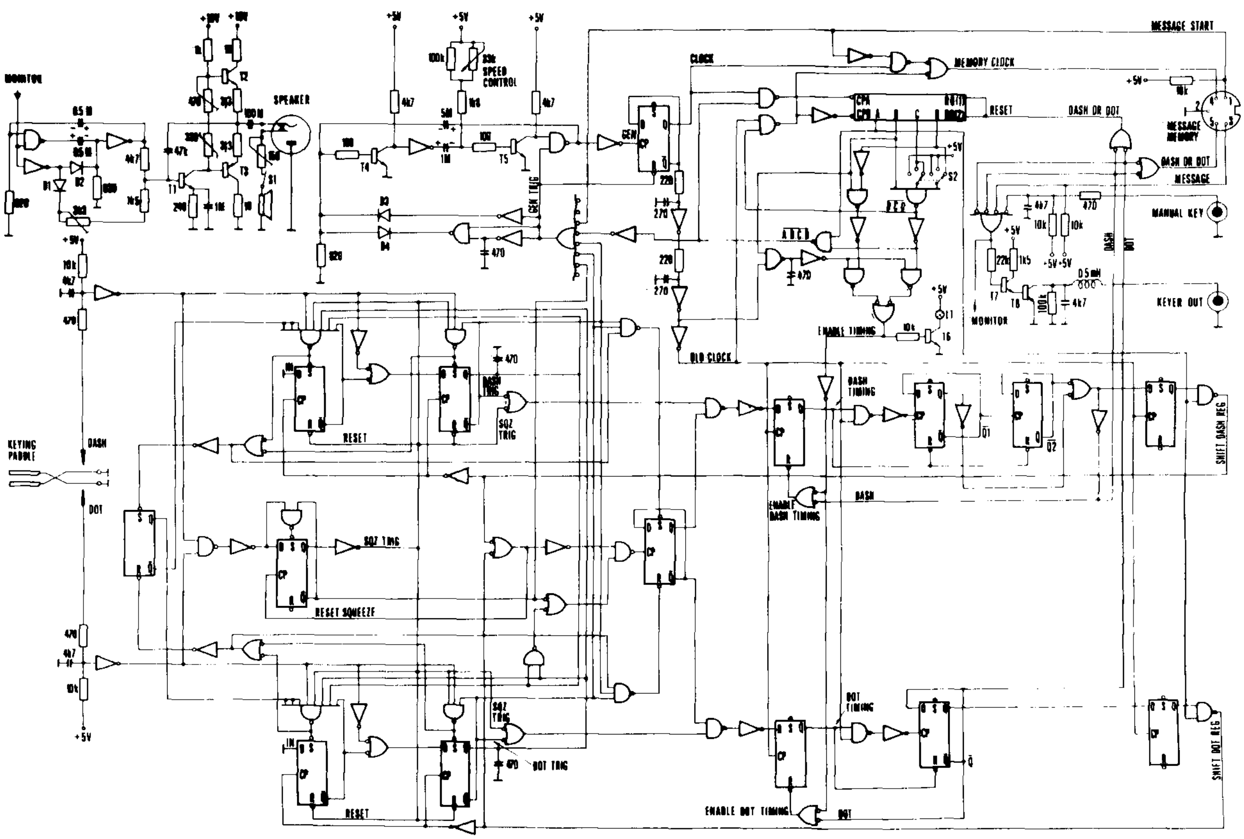


fig. 6. Schematic diagram of the advanced electronic keyer.

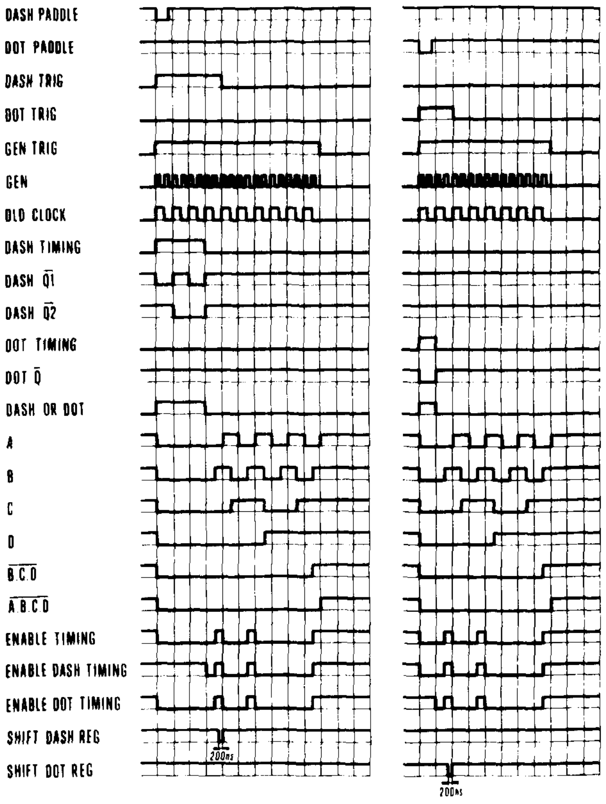


fig. 7. Timing diagrams for the signals shown in *figs. 5* and *6*.

nS, producing DLD CLOCK. Also, CLOCK is used to drive the A section of the binary counter. After the DASH and DOT signals are formed, they're shifted into another D-type flip-flop, which is clocked by DLD CLOCK. The dash or dot therefore initiates the shift register signal, while the DLD CLOCK stops it, 200 nS later.

Squeeze. During the initial phase of keying, the buffers, dash and dot registers, and input and output dash-dot distributors are involved. When both paddles are squeezed, SQZ TRIG goes low, clearing both buffer registers, enabling the clock generator, and executing the iambic sequence. When both paddles are released, RESET SQZ resets the squeeze circuit forcing SQZ TRIG to go high at the end of the first completed single-width space.

Monitor. The keyer contains an internal side-tone oscillator, which consists of a multivibrator, an integrating amplifier, and internal speaker. The multivibrator generates a 1-kHz rectangular waveform, which is integrated in the amplifier. The triangularly shaped sidetone drives an internal 8-ohm permanent magnet speaker.

Message memory. Provisions have been made to connect the keyer to an external random-access memory. Four signals are available on a rear panel connector for operation of the memory. During the "message write operation" the MEMORY CLOCK and DASH or DOT are used for synchronous storing of the message. The MEMORY CLOCK automatically starts when the paddle is activated and continues to run during the manual keying. When the manual keying stops, the MEMORY CLOCK runs for seven clock pulses after message generation ceases. MEMORY CLOCK can be enabled to run through the remaining addresses by taking MESSAGE START low. During the "message read operation" MESSAGE represents the memory content previously stored and is transmitted through the OR gate to the keyer output.

construction

The electronic keyer is built in a 7 x 10 x 21 cm (2 3/4 x 4 x 8 1/4 inch) steel cabinet. The overall view is shown in the photographs. The ICs are mounted on a universal printed-circuit board with wire wrap used for the circuit connections. The parts list is shown in **table 1**.

concluding thoughts

This advanced electronic keyer could seem to some readers to be rather complicated. However, even an inexperienced homebrewer can expect perfect performance by exercising reasonable care during construction. Additionally, there are no special circuit adjustments necessary.

table 1. Semiconductors used in the electronic keyer

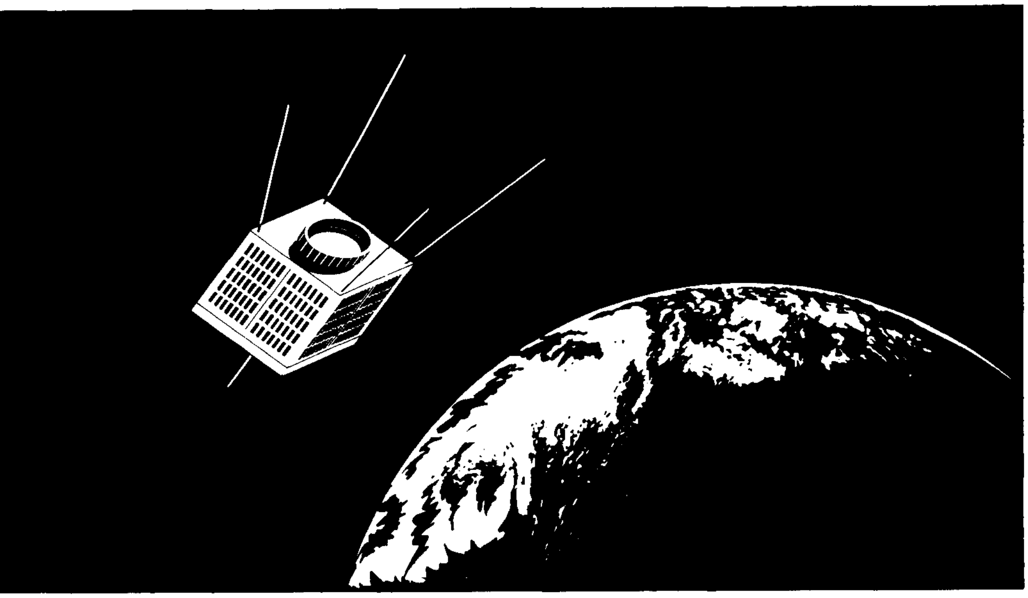
integrated circuits		transistors
7400	9 each	KC507 Silicon, 0.3 W T1, T4 - T7
7404	6 each	KF508 Silicon, 0.8 W T2, T8
7410	2 each	KF517 Silicon, 0.8 W T3
7420	1 each	
7430	3 each	
7474	8 each	Diodes
7493	1 each	KA221 Silicon, Switching D1 - D4

Only inexpensive TTL ICs were used in this project. The more experienced builder could change some circuit components or integrated circuits to ones more available in the United States. Or he could use NOR gates, monostable multivibrators, or MSI shift registers to minimize the number of parts and energy consumption. This should not be in contradiction with the aim of this article, however, which is to provide optimum keyer control and timing.

references

1. F. A. Bartlett, W6OWP, "Further Advances in Electronic Keyer Design," *QST*, October, 1948, page 27.
2. J. W. Pollock, WB2DFA, "Cosmos IC Electronic Keyer," *ham radio*, June, 1974, page 6.

ham radio



AMSAT-OSCAR D

The AMSAT-OSCAR D spacecraft is scheduled for launch sometime this month — here are the complete operating parameters for this new amateur satellite

AMSAT-OSCAR D, the next spacecraft in the OSCAR series, is a Phase II spacecraft which was built over the past two years by radio amateurs in the United States, Canada, Japan, and West Germany and is the first spacecraft in which AMSAT, Project OSCAR, and the ARRL have joined together to build flight hardware. The spacecraft makes extensive use of parts left over from the OSCAR 7 and Phase III programs, and was built primarily because the Phase III spacecraft will not be available until 1979. By stretching its resources almost to the limit, AMSAT has been able to work on both the Phase III spacecraft (with lots of publicity) and OSCAR D (with little publicity).

The new spacecraft carries transponders for two modes of operation. There is a conventional 145.9 MHz to 29.4 MHz Mode A transponder, and a new 145.9 MHz input, 435.1 MHz output, Mode J transponder — a similar frequency combination was used on the short-lived OSCAR IV spacecraft in 1966. In addition, six channels of telemetry are provided to monitor the onboard status of the satellite.

mission objectives

The principal objective of AMSAT-OSCAR D is its use as an educational tool in schools. Other objec-

By Perry Klein, W3PK, and Joe Kasser, G3ZCZ, AMSAT, Post Office Box 27, Washington, D.C. 20044.

tives include the continuation of communications demonstrations by means of stations in the amateur-satellite service, of the feasibility of using satellites with small amateur terminals for emergency communications, communications between medical centers and isolated areas, aeronautical, maritime, and land-mobile communications, direct satellite-to-home voice "broadcasting" to simple amateur receivers, and other similar applications. Further objectives are to demonstrate special operating techniques that enhance the usefulness of low orbits for these satellite applications, and to test the suitability of a new communications transponder frequency combination (Mode J) for small terminal users.

AMSAT-OSCAR D will permit the continuation of the education program, which began with OSCARs 5, 6, and 7, over the next several years, the anticipated lifetime of the satellite. OSCAR satellites have begun to play an important role in a new approach to science education. Used as remote laboratory tools, these satellites represent a pioneering utilization of an active space system in the classroom. Using inexpensive ground terminals for OSCAR satellites in schools, students can gain firsthand experience in space science. This type of direct, active involvement has relevance to the study of communications, astronomy, engineering, physics, mathematics, and meteorology. The low-cost OSCAR ground terminal puts an active satellite system at the disposal of the instructor and student.

spacecraft description

OSCAR D is a communications satellite in the Phase II (low-orbit) series which is designed to operate with small stations in the amateur-satellite service on a non-commercial basis. The spacecraft contains two communications transponders and command and telemetry systems; it is solar powered, weighs 27 kg (60 pounds), and is a 38 cm (15-inch) rectangular solid 33 cm (13 inches) high. Its anticipated useful operating lifetime is three years.

Two types of communications transponders are aboard the spacecraft. Normally, only one transponder will be operated at a time because of spacecraft battery constraints.

Two-to-ten meter transponder. The *Mode A* transponder is a two-to-ten meter unit similar to the one used on OSCAR 7 (input frequency passband 145.85-145.95 MHz, and output frequency passband between 29.40 and 29.50 MHz). A 250 mW telemetry beacon provides telemetry data in Morse code at a frequency of 29.402 MHz. Approximately −95 dBm is required at the transponder input terminals for an output of one watt. This corresponds to an effective radiated power from the ground of 80 watts for a distance to the satellite of 1930 km (1200 miles) and a polarization mismatch of 3 dB. The transponder translation frequency (input frequency minus output frequency) is 116.458 MHz. Thus, the relationship between the uplink (f_u) and downlink (f_d) is

$$f_d = f_u - 116.458 \pm Doppler$$

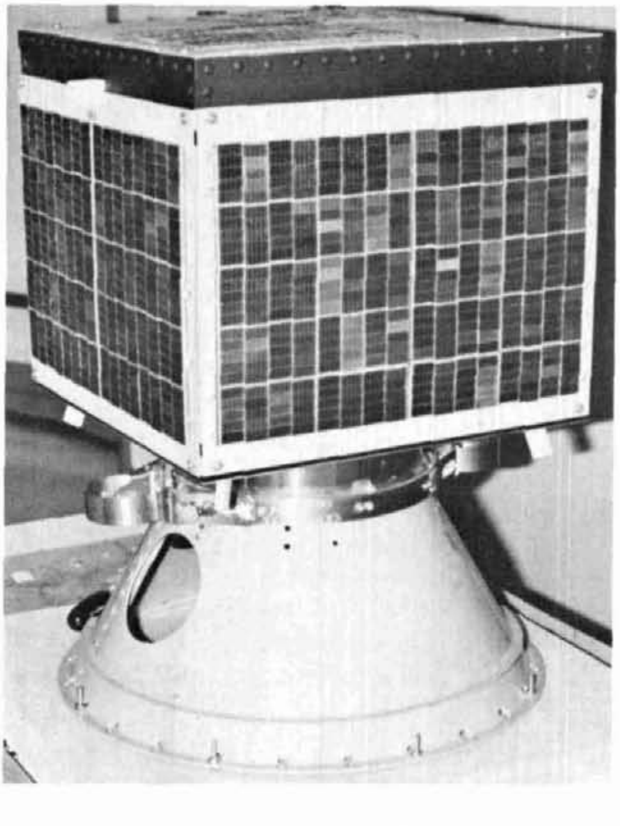
where both f_d and f_u are in MHz.

An uplink signal at 145.900 MHz, for example, will produce a down-link signal from the transponder on 29.442 MHz ± Doppler. As in the two-to-ten meter transponders in OSCAR 6 and 7, the passband is *not* inverted, and upper-sideband uplink signals become upper-sideband downlink signals. Output power is 1 to 2 watts.

Note that the downlink frequency will be slightly different (8 kHz) from that of the equivalent OSCAR 7 Mode A transponder which has an equivalent frequency relationship of

$$f_d = f_u - 116.450 \pm Doppler$$

AMSAT OSCAR D spacecraft to be launched in March, 1978, by NASA. When it is in orbit it will be designated AMSAT OSCAR 8.



Two meter-to-70cm transponder. The Mode J transponder, constructed by members of the Japan AMSAT Association in Tokyo, uses a two-meter input, 70 centimeter output combination which has not yet been flown in the AMSAT Phase II series. This transponder operates with an input frequency passband of 145.90-146.00 MHz, and an output frequency passband of 435.10-435.20 MHz. Power output is about 1-2 watts PEP, and the output passband is *inverted, i.e.*, upper-sideband uplink signals become lower-sideband downlink signals. The transponder translation frequency (input frequency plus output frequency) is 581.1 MHz ± Doppler. Uplink sensitivity for one-watt output is − 105 dBm, corresponding to an effective radiated power from the ground of 8 watts for a distance to the satellite of 1930 km (1200 miles).* Note the greatly improved sensitivity of this mode, and keep your power down. A 100-milliwatt beacon carries telemetry at a frequency of 435.095 MHz. The relationship between the uplink (f_u) and downlink (f_d) is

JA1CBL of Japan AMSAT Association (JAMSAT) building parts of the Mode J transponder.

$$f_d = 581.1 - f_u \pm Doppler$$

where both f_d and f_u are in MHz.

antenna system

Both the Mode A and Mode J transponders use the same receiving antenna, a canted turnstile comprised of four 48 cm (19-inch) lengths of 12.5 mm (1/2-inch) carpenter's rule fed by a hybrid and matching network so as to develop circular polarization. One port of the hybrid feeds the Mode A receiver; left-hand circular polarization is required by users in the Northern Hemisphere and right-hand circular polarization is required in the Southern Hemisphere. A second port of the hybrid is connected to the Mode J receiver; right-hand circular polarization is required in the Northern Hemisphere, and left-hand circular polarization in the Southern Hemisphere. The antenna gain should approach 5 dB in the $-Z$ direction (*i.e.*, toward the bottom of the satellite).

The Mode A ten-meter downlink antenna is a linearly-polarized dipole which is oriented perpendicular to the stabilization magnets in the spacecraft as in OSCAR 6 (but unlike OSCAR 7, which has the ten-meter antenna parallel to the axis of the magnets).

The 435-MHz Mode J downlink antenna is a simple monopole, linearly polarized, and located on the top of the spacecraft. Note that its location may result in some radiation shielding at high Southern Hemisphere latitudes.

*Sensitivity may decrease by a factor of ten (10 dB) under different conditions of battery voltage and satellite operating temperature, so that as much as 80 watts may be required at certain times.

telecommand system

A five-function telecommand system of a new design is carried on OSCAR D. The system is based on the best features of the OSCAR 6 and 7 telecommand systems, and is designed to be virtually immune to noise and interference. The command functions are:

Mode A Select (Two-to-ten meter transponder ON)
Mode J Select (2 meter to 70 cm transponder ON)
Mode D Select (recharge mode; both transponders OFF)
Ten-meter Antenna Deployment
Ten-meter Antenna Reset

telemetry system

OSCAR D contains a six-channel Morse code telemetry system similar to the units flown in OSCARs 6 and 7. Telemetry is sent at 20 words per minute as three-digit numbers; *A1* emission is used in keying the Mode A or Mode J telemetry beacons, depending upon which transponder is in use. The six telemetry parameters are:

Channel 1. Total Solar Array Current —
$I_T = 7.15\,(101\text{-}N)\ ma.$

Channel 2. Battery Charge-Discharge Current —
$I_{Bat} = 57\,(N\text{-}50)\ ma.$

Channel 3. Battery Voltage —
$V_B = 0.1\,N + 8.25\ Volts$

Channel 4. Baseplate Temperature – $T_{bp} = 95.8 - 1.48\,N\ (^{\circ}C)$

Channel 5. Battery Temperature – $T_{Bat} = 95.8 - 1.48\,N\ (^{\circ}C)$

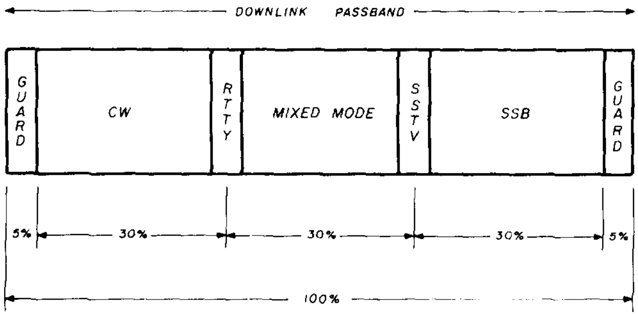
Channel 6. RF Power Output, Mode J – $P_{JT} = 23N$ *milliwatts*

Note that, unlike OSCAR 6 and 7 telemetry, OSCAR D has only one parameter per line (OSCARs 6 and 7 had 4). As a result, a complete telemetry frame is sent in approximately 20 seconds.

power supply

The spacecraft contains solar panels on its four sides (along the $+X$, $-X$, $+Y$, and $-Y$ axes), and on the top (the $+Z$ axis). No panels are contained on the bottom ($-Z$ axis) since this is where the spacecraft is attached to the launch vehicle. The solar cells, combined with a 12-cell, six-ampere-hour rechargeable nickel-cadmium battery should be adequate to power the spacecraft with a positive power budget in Mode A for several years, even considering solar cell degradation in the radiation environment. The power drain in Mode J, however, is somewhat larger, so the Mode J transponder probably cannot be operated continuously.

A battery charge regulator is also contained which converts from the 28-30 volt solar array voltage to the 14-16 volts required by the battery. It also tapers the charge rate so the battery trickle-charges as the battery approaches full charge (as indicated by the battery voltage).



1. Guard bands to avoid interference to beacons. These frequencies are available for emergency and bulletin stations.
2. RTTY and slow-scan television are placed at the edge of the CW and ssb passbands. This conforms to their high-frequency use where RTTY is present within the CW space, and SSTV is transmitted in the ssb sub-band.
3. Mixed-mode area is recommended for crystal-controlled stations, DXpedition stations, or anyone wishing to work both CW and ssb stations.

fig. 1. Basic satellite band plan proposed by G3ZCZ and adopted by the AMSAT Board of Directors in October, 1977. This band plan allocates a percentage of the available radio frequency spectrum as seen on the downlink to different modes of communication. The relative amount of spectrum for each mode is thus the same for any transponder in any satellite. Band plan frequencies for AMSAT OSCARs 7 and 8 are shown in *fig. 2*.

stabilization system

Four permanent magnets located inside the spacecraft and aligned along the Z axis provide stabilization, as in OSCARs 6 and 7. The polarity of the magnets is such that the top ($+Z$ axis) of the spacecraft always points toward the magnetic north pole of the earth. Hysteresis permalloy damping rods mounted behind the $+X$, $-X$, $+Y$, and $-Y$ solar panels are designed to reduce the spin of the spacecraft about the Z axis; they function in a manner similar to a shorted transformer turn as it cuts the lines of flux of the earth's magnetic field. The permalloy rods are left over from OSCAR 7, which successfully used the same type of stabilization system.

launch interface and orbit

The OSCAR D spacecraft is being launched from the NASA Western Test Range as a secondary payload with the NASA Landsat-C earth resources technology satellite and the NASA PIX (Plasma Interaction Experiment). The spacecraft will be ejected from the second stage of the two-stage Thor-Delta 2910 launch vehicle 5120.6 seconds after lift-off, at an approximate position of 78 degrees N. latitude and 15 degrees W. longitude. Programmed orbital parameters are:

Apogee 928 km (577 statute miles)
Perigee 884 km (549 statute miles)
Period 103 minutes
Inclination 99.0 degrees
Time of Descending Node 9:30 AM (launch window from 1754-1824 UTC)

The orbit is planned to be sun-synchronous with passes repeating at the same time each day on a one-day cycle (as opposed to the two-day cycle of OSCARs 6 and 7).

spacecraft initialization

AMSAT-OSCAR D will automatically be powered up upon ejection from the Thor-Delta launch vehicle over northern Greenland at which time it will assume the next available number in the OSCAR series. It is designed to initialize itself in Mode J (two-meter-to-70-cm transponder ON). The two-to-ten meter (Mode A) transponder will be initialized OFF and should be kept off until the spacecraft is nearly completely stabilized, which may require as much as a

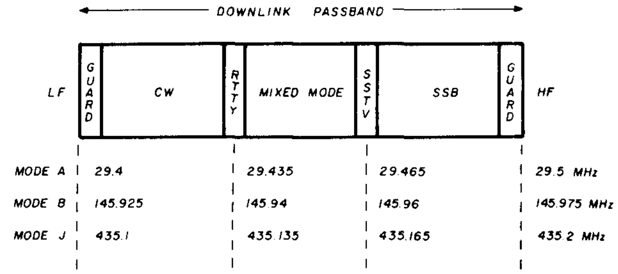


fig. 2. Satellite band plan for AMSAT OSCAR 7 and 8. As shown in *fig. 1*, this plan is based on percentages of the downlink passband, and applies to both inverting and non-inverting transponders. In Modes A and J the guard channels are 5 kHz wide; the guard channels in Mode B are 2.5 kHz wide.

week. Because of the non-rigidity of the deployable ten-meter dipole antenna, this antenna must *not* be deployed until the spacecraft spin rate is less than 1 rpm; otherwise, the antenna may be severely damaged or may not deploy properly. OSCAR D's ten-meter antenna is comprised of tubular extendable members which are slowly unreeled from the spacecraft by small motors. The deployment process takes approximately 15 seconds and is non-reversible (*i.e.*, the antenna elements cannot be retracted once they are deployed). During the time when the antenna is being deployed, the telemetry beacon switches from its normal Morse code format to a series of keying pulses, the rate of which is a function of the tip-to-tip length of the ten-meter dipole; the rate-of-change of the telemetry pulses will be carefully tape-recorded during deployment of the antenna to permit analysis later to verify success.

telecommand verification

OSCAR D's telecommand and telemetry systems have been designed to provide two means to verify whether the spacecraft is accepting commands. First, when the telecommand system has been enabled and is ready to accept a command, the Morse code telemetry will be interrupted and an unmodulated carrier will be heard on the beacon frequency. The beacon will revert back to Morse code telemetry when the telecommand system is no longer enabled.

The second method of telecommand verification is to use the "Ten-meter Antenna Deployment" command. This will cause a series of keying pulses to be heard on the telemetry beacon in place of the Morse code telemetry if the command has been accepted. The "Ten-meter Antenna Reset" command should be sent soon afterward to restore the beacon to the Morse code telemetry mode.

telemetry interpretation

The most important telemetry channel that will affect operations decisions is *channel 3* (battery voltage). In Mode A the spacecraft should maintain a positive power budget so there should not be a net discharge of the battery over an orbit average. Mode J operation, however, requires somewhat more power, which may result in a net discharge of the battery, especially under conditions of high transponder loading; therefore it will be necessary for telemetry and telecommand stations to keep a close watch on the battery voltage so that action can be taken as necessary to command the spacecraft into Mode D (the recharge mode) before the battery discharges too far. Three cutoff levels are specified below:

Red Level A (1.2 volts/cell) Channel 3 = 61 counts
Red Level B (1.1 volts/cell) Channel 3 = 50 counts
Red Level C (1.0 volts/cell) Channel 3 = 38 counts

Red Level **A** should be used during the first year or so of the spacecraft's life as the cutoff point below which telecommand stations should command the satellite into Mode D for recharging. Later in the spacecraft's life, as the battery discharge characteristic curve changes, Red Level **B** should be used; Red Level **C** should be used if there is evidence of battery deterioration or if it is desired to recondition the battery.

Channel 1 (solar array current) provides an indication of whether the spacecraft is in the sun or eclipse (it should read in the nineties in counts when in eclipse). Fluctuation in channel 1 telemetry is the best indicator of the rate of spin of the spacecraft, along with observations of fading, particularly of the 435-MHz Mode J downlink signal from the quarter-wave 435-MHz monopole antenna.

Channel 2 (battery charge-discharge current) gives information on whether the battery is charging or discharging. A reading larger than 50 counts indicates that the battery is charging, while a reading of less than 50 counts means the battery is discharging. There is a two-second integration time associated with the current telemetered on this channel. The total power drain of the spacecraft can be determined by observing channel 2 while the spacecraft is in darkness (as indicated by channel 1, which should read in the 90s in darkness).

Channels 4 and 5 (baseplate temperature and battery temperature) should generally track within a few degrees (except perhaps in the first day or so after launch when the spacecraft has not yet stabilized at thermal equilibrium). Experience from OSCARs 6 and 7 indicate that the battery can overcharge and overheat during periods of the year when the spacecraft sees the most sunlight. If this is the case, channel 5 may exceed channel 4 in temperature by 10 degrees celsius or more, and action should be taken to reduce this overheating. This can be accomplished by keeping the spacecraft in Mode J to consume any extra charge current from the battery.

Channel 6 is a measure of the Mode J transponder 435-MHz rf power output. Associated with the telemetered readings is an integration time of 2.5 seconds, so that it is average power rather than peak power that is telemetered. There is no telemetry of the Mode A transponder. The Mode A transponder power consumption (largely determined by power amplifier current) can be measured by observing channel 2 telemetry as noted above.

operating schedule

Since the prime mission of the AMSAT-OSCAR D spacecraft is to use the Mode A transponder for the ARRL OSCAR educational program in schools, the spacecraft may be left in Mode A during weekdays (Monday through Fridays, United States time), and put in Mode J on weekends. Note that all communications should conform to the G3ZCZ band plan shown in **figs. 1** and **2**. Additionally, if not an excessive burden on the telecommand stations, evening orbits in the Western Hemisphere (morning orbits in the Eastern Hemisphere) can be switched to Mode J, battery permitting. In any case, all operation in Mode J will require careful monitoring of the battery charge level (as indicated from channel 3 telemetry). The power budget may not support the Mode J transponder for continuous fulltime operation in this mode over an entire weekend. In any event, details of the operational modes of the spacecraft will be announced by AMSAT in the *Newsletter*, with late updates on the AMSAT Nets.

OSCAR D will operate in a 900 km (560 statute mile) orbit, *i.e.*, at just over half the altitude of the orbit of OSCAR 7. Thus, communication ranges will be different. The usable time on an overhead pass will be about 18 minutes instead of the 22 minutes provided by OSCAR 7, and the horizon range will be 3220 km (2000 miles) instead of the 3940 km (2450 miles) of OSCAR 7. This means that transatlantic communications will still be possible, but not as often as with OSCAR 7.

Keeping track of this satellite is going to be much simpler than for OSCAR 7. It will come into range at approximately the same time each day; the overhead descending node pass is planned for 9:30 AM local time.

credits

It is impossible to single out all those who contributed to the construction of the spacecraft, but a few calls can be listed.

JAMSAT — Mode J Transponder: JA1ANG, JA1CBL, JG1CDM, JA1VDV, JA1JHF, JR1SWB

AMSAT — Mode A Transponder: WA4DGU, W3PK

Morse Code Telemetry System: W5CAY, WA4DGU

Telecommand System: W3GEY, WA3LND, WA3ZCE, W3HUC, W3ITO, K1RT/WA1JZC

Antenna and Antenna Deployment Module: W3GEY, W3HUC, W3ITO, K1RT, WA3LND

Power System: DJ4ZC, JA1TUR, JF1DMQ, K1RT, W3HQ

Structure and Module Containers: K6GSJ and Project OSCAR, K1JX/WA1JLD, K1RT, WA4DGU, VE3DPB, W3HSO, WBØGIM, Henry Smith, David Vanderbeke, W3ZKI

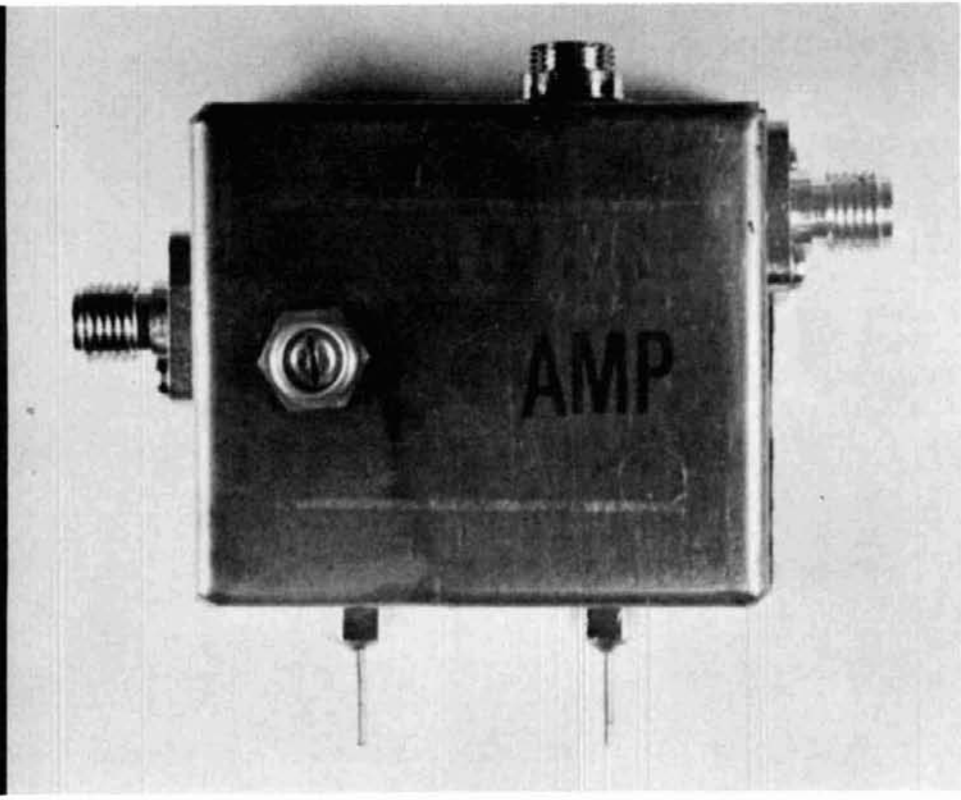
Cables and Wiring: Marie Marr, W3TMZ

Engineering Drafting: WB4GIB

To others who contributed or assisted, our thanks and the thanks of thousands of radio amateurs, school children, and educators. Let's use the satellite wisely so that it will further help the educational program until the Phase III satellites are flying high.

ham radio

"Alfred was never any good in a crisis!"



very low-noise GaAs FET Preamp for 432 MHz

Construction details for a uhf GaAs fet preamp that will provide a 0.7 dB noise figure with 18 dB gain at 432 MHz

Every amateur who is active on the uhf bands is looking for ways to improve his system performance. One of these ways is to use a receiver preamplifier with lower noise figure and higher gain; some excellent preamplifier circuits have been published in *ham radio*. The low-noise uhf preamplifier described by W1JR in the March, 1975, issue is perhaps the best known of these circuits,[1] and is widely used by EME operators and others who are interested in long distance, weak-signal uhf communications.

This article describes a low-noise, state-of-the-art preamplifier for 432 MHz which uses an NEC NE24406 (2SK85) GaAs fet. This preamp is capable of a 50°K (0.7 dB) noise figure at 423 MHz, so it can improve the performance of your uhf receiving system by about 100°K, as compared with a conventional bipolar transistor.

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field-effect transistors

The history of the fet is older than most people realize. In fact, it's one of the oldest three-terminal solid-state devices. Dr. Shockley first proposed the fet in 1952, but because of a variety of technological and fabrication difficulties, fets did not become practical until the early 1960s.

Basically, there are three different types of fets. The junction fet or *jfet* is the simplest of the three types and became commercially available about the same time as the first bipolar microwave transistors.

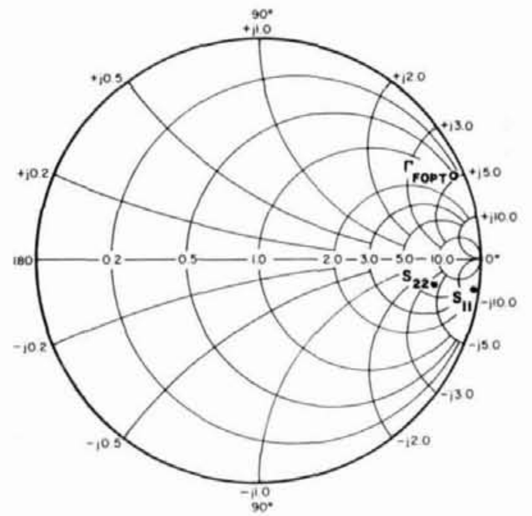


fig. 1. Smith chart plot of the NE24406 impedance characteristics at 432 MHz. Γ_{FOPT} ***is the optimum source impedance for noise figure at 423 MHz.***

Advances in fabrication techniques and requirements for lower power fostered the development of the metal-oxide-semiconductor fet or *mosfet*. Both the jfet and mosfet are widely used in applications which require high input impedances, such as in the input stages of test instruments. The same fabrication processes developed for mosfets are also useful in the manufacture of integrated circuits.

In the microwave region the bipolar transistor has reigned supreme for a number of years — it wasn't threatened by either the jfet or the mosfet. However, a third type of fet has changed that. This new fet, which uses a Schottky barrier at the gate electrode, is called the metal-semiconductor fet or *mesfet*. Mesfets use Gallium Arsenide (GaAs) as the semiconductor material and are usually referred to as GaAs fets (pronounced *gas* fets).

Gallium arsenide has high electron mobility (5 to 7 times as high as silicon), and offers significant advantages over silicon at microwave frequencies. The intrinsic characteristics of GaAs result in shorter transit times and lower resistance, thereby providing higher gain, lower noise figures, and extremely high cut-off frequencies — all important characteristics for microwave transistors.

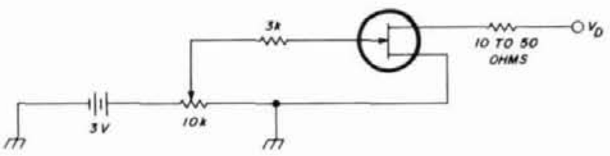
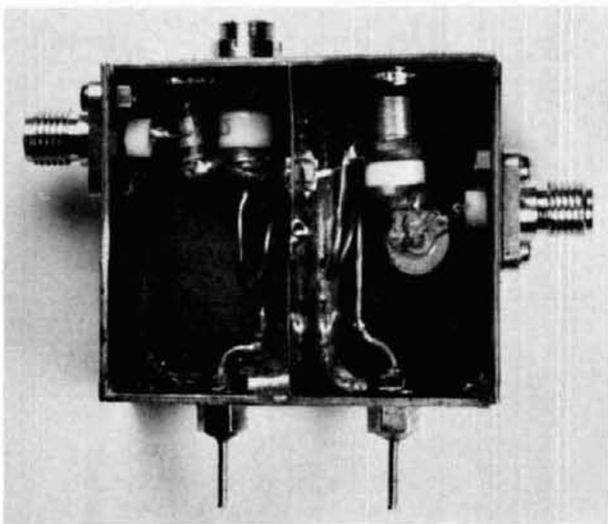


fig. 2. Recommended GaAs fet bias circuit uses a series resistor in the gate circuit to protect the device from transients.

Gallium arsenide has been under development for several years, and practical microwave GaAs fets are now available off the shelf. Recently I had an opportunity to experiment with the NEC NE24406 (2SK85) GaAs fet. Although there have been many published reports which describe the performance of the NE24406 in amplifiers up to X band (about 12 GHz), there have been no published data on their use below 500 MHz. This article describes the first experimental results of the NE24406 GaAs fet on the amateur 70 cm band.

noise figure

Obviously a transistor with a low noise figure is required in a low noise preamplifier, but that's not the



Layout of the low-noise 432-MHz preamplifier. The input is to the left, output to the right. The GaAs fet is installed in the center shield partition. Capacitors C1 and C3 are supported by their own leads; C2 (left compartment) and C4 (right compartment) are mounted in the side walls of the chassis.

only requirement. Also to be considered are the effects of the next stage's noise figure and the requirements for a low-loss matching circuit, low feedback, and good stability.

The overall noise figure of a receiving system is given by the following formula

$$NF_T = NF_1 + \frac{NF_2 - 1}{G_1} \quad (1)$$

Where

NF_T = Total overall noise factor

NF_1 = Noise factor of the first stage

NF_2 = Noise factor of the second stage

G_1 = Available power gain of the first stage

Note that the noise factors and power gains must be in ratios, *not* in dB. For example, consider that you have a preamplifier with a 1.0 dB noise figure ($NF_1 = 1.259$) and 10 dB power gain ($G_1 = 10$); the noise figure of the second stage is 5 dB ($NF_2 = 3.162$). The calculated overall noise factor NF_T is 1.475 (or a noise figure of 1.688 dB).

Note that the system noise figure has increased about 17 per cent as compared to the preamplifier's noise figure. For best results, it's recommended that you use a lower noise second stage with a higher gain preamp.

One question that often arises is, "Can the noise figure of the preamplifier be as low as the noise figure of the transistor itself?" No, it cannot, even on the

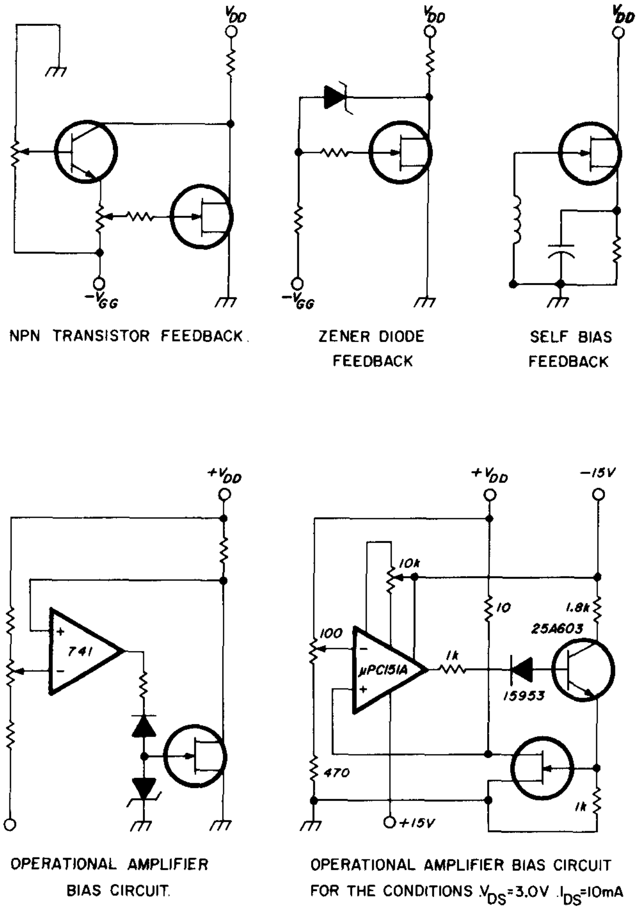


fig. 3. Several circuits that may be used to provide bias in fet circuits. Note that all circuits must be bypassed at rf.

assumption that there are no matching or circuit losses. One way to describe the optimum noise figure of a transistor is *noise measure* which is expressed as

$$M + 1 = \frac{NF - 1}{1 - \frac{1}{G}} \quad (2)$$

where

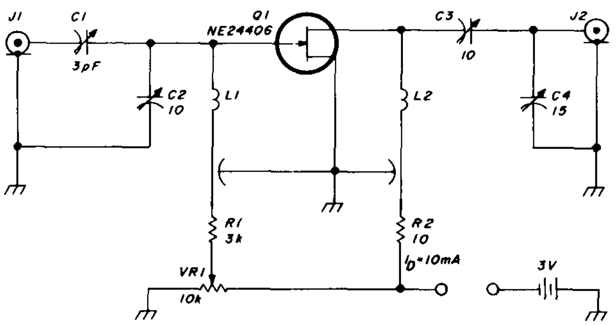
NF = Noise figure of the transistor

G = Associated gain of the transistor

The value of $M + 1$ shows the ideal total system noise figure, and indicates the same result where an infinite number of transistors with the same characteristics are used in cascade.

Assume you have a transistor which has a specified noise figure of 1.2 dB ($NF = 1.318$) and 14 dB gain ($G = 25.12$). From **eq. 2**

$$M + 1 = \frac{1.318}{1 - \frac{1}{25.12}} = 1.351 \text{ or } 1.243 \text{ dB}$$



C1 3 pF piston trimmer (Johanson 7274 or 7284)
C2, C3 10 pF piston trimmer (Johanson 5202)
C4 15 pF piston trimmer (Johanson 5402)
L1, L2 1 turn no. 20 (0.8mm), 1/4" (6mm) ID (see **fig. 5**)
J1, J2 SMA type coaxial connectors
Q1 NEC NE24406 GaAs fet

fig. 4. Schematic diagram of the low-noise preamp for 432 MHz which provides a 0.7 dB noise figure and 18 dB gain. Lowest noise figure occurs at drain current flow of 10 mA (see fig. 7).

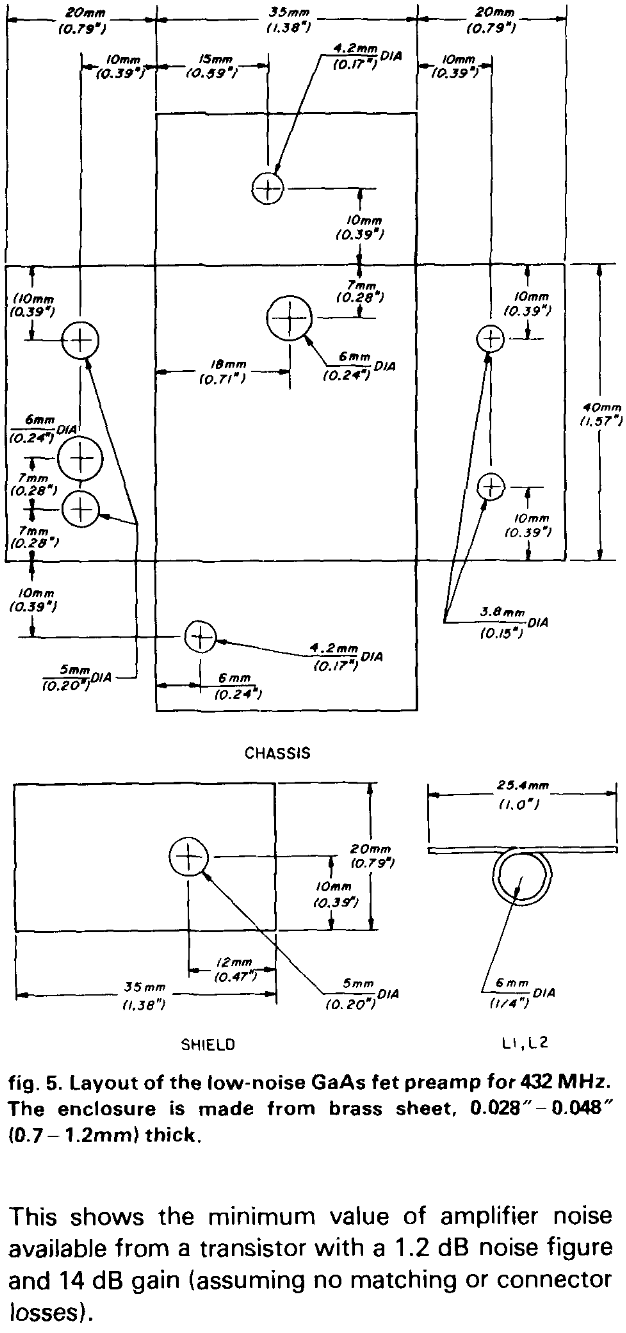


fig. 5. Layout of the low-noise GaAs fet preamp for 432 MHz. The enclosure is made from brass sheet, 0.028″ – 0.048″ (0.7 – 1.2mm) thick.

This shows the minimum value of amplifier noise available from a transistor with a 1.2 dB noise figure and 14 dB gain (assuming no matching or connector losses).

circuit description

The GaAs fet has several disadvantages which make it somewhat more difficult to use than the bipolar transistor. For one thing, as shown in the Smith chart plot of **fig. 1**, the impedances are very high. This also indicates that the GaAs fet is a high Q device and not easy to match to low impedances such as 50 ohms. Also to be considered is the bias circuit. Once an operating point has been established for the GaAs fet, a bias circuit must be chosen which will provide stable operation over the required temperature and frequency range.

The importance of affording adequate protection against transients and change in I_{DSS} cannot be over emphasized. As every designer who uses GaAs fets will eventually discover, transients are the leading "fet killer." The most likely burn-out mode is a short from gate to drain or from gate to source which is caused by high field or high current transients. The highest field in common-source operation is between the drain and gate, and should never exceed 10 volts.

The best way of applying bias is to use a battery through a series resistor to the gate, as shown in **fig. 2**. Although two power supplies are normally required, a source resistor may be used to develop the necessary reverse bias, but it may reduce both gain and noise figure. Several bias circuits are shown in **fig. 3**.[2] It should be mentioned that all bias circuits must be bypassed to rf; a series gate resistor from 1000 ohms to 10k will protect the gate from high-frequency transients.

For best performance a low-loss input matching circuit is very important; low-loss components should be used in the simplest possible low Q circuit. As can be seen in the Smith chart plot of the NE24406's S-parameters, $|S_{11}|$ and $|S_{12}|$ are very large; the optimum source impedance for noise figure, Γ_{FOPT}, is also very high. Fortunately, the optimum source impedances for gain and noise figure are not greatly different. Therefore, the input matching circuit can consist of a series capacitance or inductance. Since the series reactance must be carefully selected, the capacitor has the advantage of being easily tuned.

Theoretically the preamplifier will work fine with only a series capacitance in the input circuit, but the NE24406 can provide 12 dB gain in the 4-GHz band, so there's a possibility of oscillation outside the 430-MHz band. The parallel resonant circuit, L1-C2, in **fig. 4** suppresses these oscillations. L1 is also used to supply bias to the gate of the GaAs fet. The resonant impedance of L1-C2 is very high, and input matching is actually provided only by C1.

As was shown in **eq. 2**, the noise figure of a transistor amplifier depends greatly on the available gain of the device being used. Since a gain increment of 4.0 to 4.5 dB ($|S_{22}| \cong 0.8$) can be anticipated at 432 MHz for the NE24406 when the output is matched,

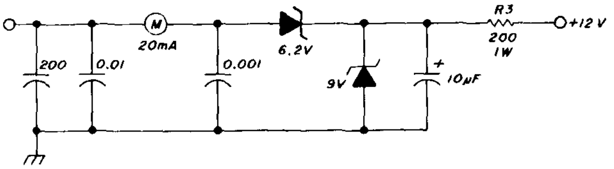


fig. 6. Recommended drain voltage supply circuit for the NE24406 GaAs fet preamplifier.

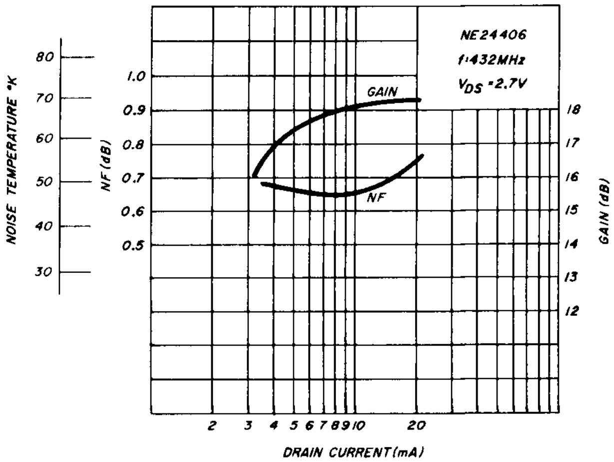


fig. 7. Preamp noise figure and power gain as a function of drain current. Note that lowest noise figure occurs at I_D = 10 mA.

improper output matching will degrade noise figure. Since the output impedance of the GaAs fet is extremely high, I installed a simple parallel resonant circuit at the output (L2, C2, C3), and matched to the 50-ohm output with the capacitive portion of the network. Both the input and output circuits have relatively high *Q*, so stable operatin is obtained over a narrow bandwidth.

construction

Basic construction of the low-noise 432-MHz preamplifier is shown in **fig. 5**. The enclosure is made from no. 18 to no. 22 (0.028-0.048 inch or 0.7-1.2mm thick) brass sheet. Do not omit the center shield — it's absolutely necessary to obtain stable operation.

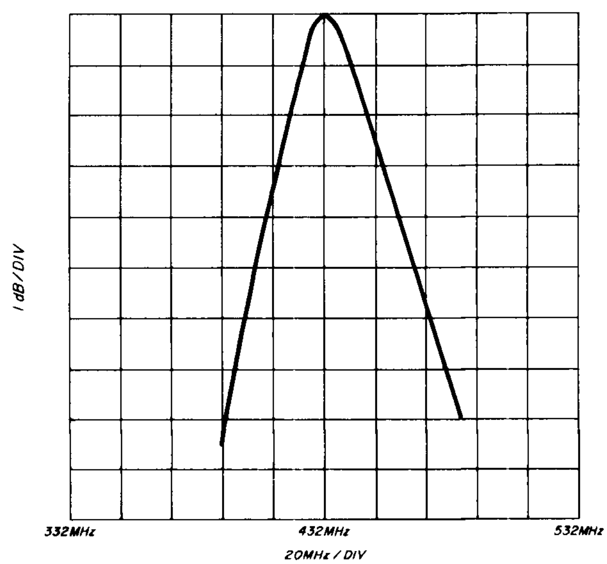


fig. 8. Passband response of the low-noise 432-MHz preamp; gain is 3 dB down at approximately 414 and 456 MHz.

Although no particular care is required when mounting the GaAs fet, the source leads should be separated and soldered directly to the holes in the shield plate. Also, when soldering the drain and gate leads to the input and output coils, take care not to pull on them too strongly.

Any type of coaxial connector may be used at the input and output, but for best noise performance it's important that the connector have good uhf characteristics. The losses of BNC connectors will degrade the noise figure. I used type SMA connectors. Type N or TNC connectors could also be used, but SMA types are much smaller and therefore more suitable

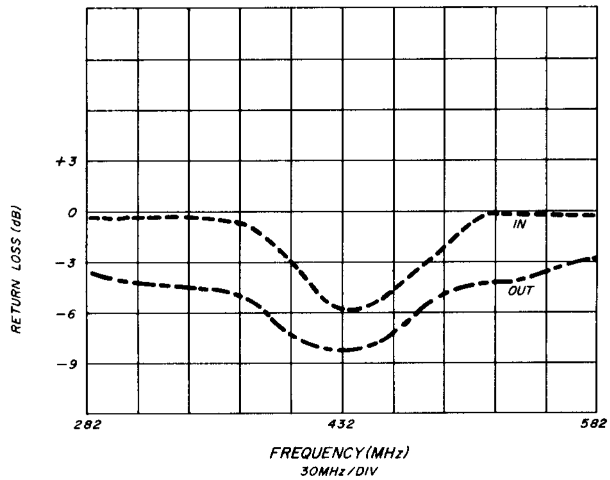


fig. 9. Return loss (vswr) of the GaAs fet preamp.

for a miniature low-noise preamplifier such as this one.

operation and test

After construction is completed, inspect each part to make sure you haven't made any wiring errors. The components in the bias circuit should be checked and double checked. When you are satisfied that the circuit is correctly built, the bias and drain supply voltages should be applied to the GaAs fet in the following manner:

1. Voltage is initially applied to the gate circuit with a 3-volt battery (see **fig. 4**). This reverse biases the gate and prevents current flow in the drain circuit, which may reach the magnitude of I_{DSS}. There is no problem in allowing current flow up to I_{DSS}, but the intention here is to suppress any transient phenomenon due to this current (there are examples where I_{DSS} reaches 100 mA).

2. Next apply the drain voltage, but make sure that the voltage between drain and source, V_{DS}, does not exceed 3 volts. When the drain voltage approaches 2.7 volts, the drain current should be set to 10 mA by adjusting the 10k pot in the bias circuit.

With the completion of these two steps, amplifier bias is established. A recommended drain bias circuit is shown in **fig. 6**.

Note that when a reverse bias is applied to the gate and drain, and current flows with only a slight application of drain voltage, either the gate circuit is open or a breakdown has occurred in the GaAs fet.

After completing the bias adjustments, adjust the input tuned circuit to resonance with a grid dipper. Apply a *weak* signal in the 430-MHz band and tune C1, C3, and C4 for maximum gain. When the preamp is adjusted for maximum gain, the noise figure will deteriorate slightly, but not seriously.

performance

The performance of the GaAs fet preamplifier is shown in the graphs of **figs. 7, 8**, and **9**. The plot of noise figure *vs* drain current (**fig. 7**) shows that lowest noise figure occurs at $I_D = 10\ mA$. The bandpass characteristic of the preamp is shown in **fig. 8**; when the preamplifier is peaked up to 432 MHz, gain is 3 dB down at approximately 414 and 456 MHz. In most applications no external bandpass filter should be required. **Fig. 9** shows the vswr at the input and output of the preamplifier. The third-order IMD products are shown in **fig. 10**; 1 dB compression occurs at about 0 dBm — the third-order intercept point is at a very respectable +20 dBm (100 mW).

When this preamplifier is adjusted for maximum

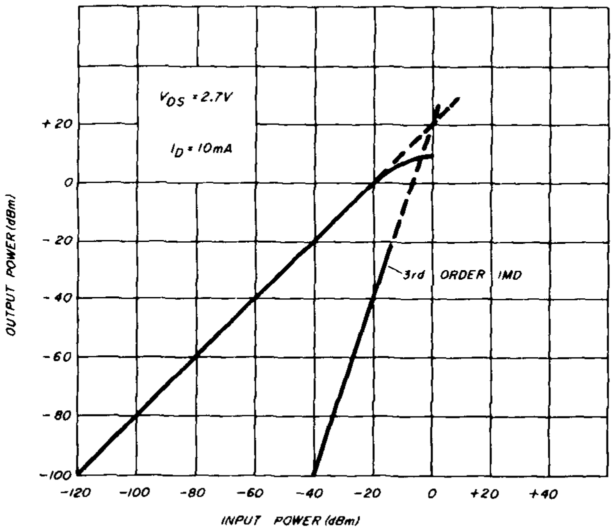


fig. 10. Input-output characteristics of the GaAs fet preamp. Third-order intercept point is at +20 dBm (100 mW); 1 dB compression is at approximately 0 dBm (1 mW).

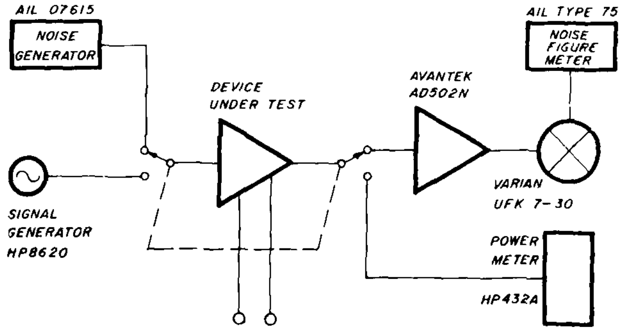


fig. 11. Block diagram of the noise-figure measurement set-up used by JH1BRY when evaluating preamplifier performance.

gain, the measured noise figure is about 0.75 dB. If an automatic noise figure meter is available, the input and output circuits can be adjusted for best noise figure — my measurements indicate that a noise figure improvement of about 0.1 dB is possible.

When making noise-figure measurements, I use an AIL noise-figure meter with a *solid-state* noise source (**fig. 11**). Gaseous-discharge noise sources are also available for this frequency range, but they should *never* be used with fragile GaAs fet circuits which are susceptible to damage from surge transients.

summary

This GaAs fet preamplifier should bring you right up to the state-of-the-art in noise figure at 432 MHz. The NE24406 GaAs fets are available in the United States from California Eastern Laboratories* or one of their sales representatives.

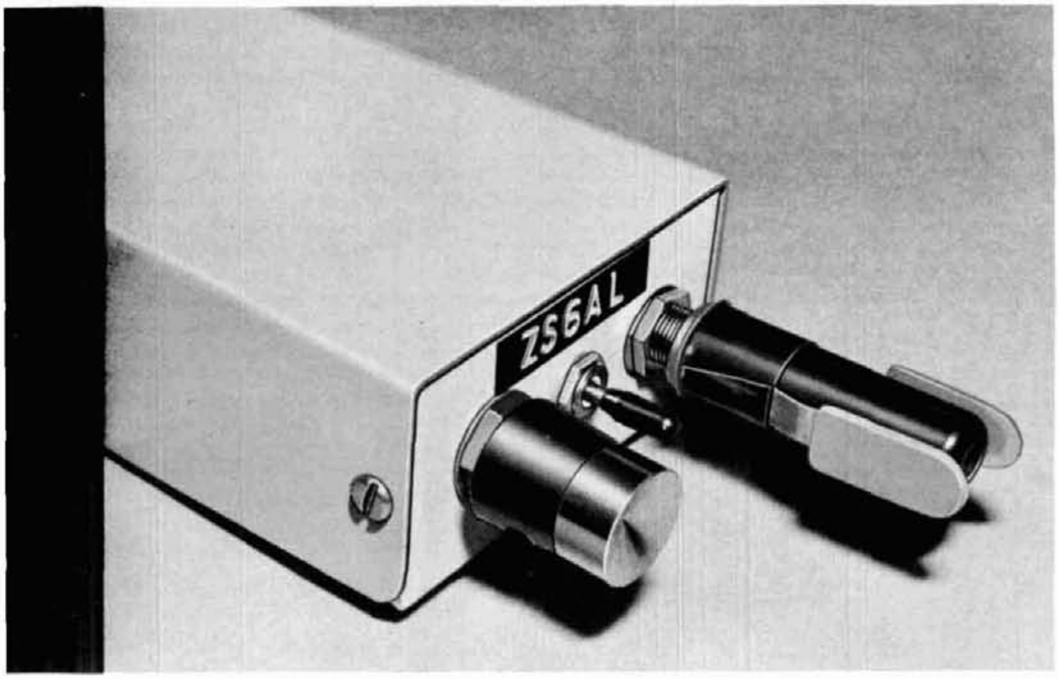
I would like to express my deep appreciation to Aki Munezuka, JA1VDV, for providing me with detail information on 432-MHz EME and kindling the fire of uhf ssb, and to Carl Peterson, K6VJN, who gave me an opportunity to publish this article. I also wish to thank the Nippon Electric Company for the use of test equipment and devices necessary to design this preamplifier. Special thanks go to Haruo Yoneda, JA1ANG, for all his helpful suggestions.

*California Eastern Laboratories, Inc., One Edwards Court, Post Office Box 915, Burlingame, California 94010; telephone (415) 342-7744. The NE24406 is priced at $190.00 in small quantities. After this article was written, NEC announced the basic GaAs fet in a smaller package at a lower price; this device is designated the NE24483 and is priced at $120.00.

references

1. Joseph Reisert, W1JR, "Ultra Low-Noise UHF Preamplifier," *ham radio*, March, 1975, page 8.
2. Jerry Arden, "The Design, Performance, and Application of the NEC V244 and V388 GaAs FET," California Eastern Labs, Burlingame, California, June, 1976.

ham radio



simple paddle

for electronic keyers

Construction details for the Ambidextrous Paddle for Electronic Keyers or APEK

The need may arise for the serious CW operator to take a battery operated electronic keyer on a mobile trip or to environments where a heavy fixed station dual-lever paddle simply has to stay home. I have improvised several keying devices in the past which provide comfortable keying of the mobile or portable rig under the varied conditions typical of such excursions. These devices were, however, all of the single lever variety that either had to be wedged, clamped, or clipped on to something, and moreover, had an extra lead going to the keyer — a nuisance at the best of times.

The keying paddle described here is essentially a dual-lever device which has proved to be so versatile that most of the problems typical of portable or mobile operation could be surmounted. As a matter of fact since inventing this Ambidextrous Paddle for Electronic Keyers, which I call APEK, the heavy dual-lever paddle at the fixed station has hardly been used at all.

The APEK is basically a three-contact jack plug with a few bits and pieces added, the total cost of which is hardly worth mentioning. Should a particular keyer, however, not be fitted with a jack type paddle socket, such a socket will have to be mounted on the front panel in a convenient position and connected in parallel with the existing socket. Since the

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APEK plugs directly into the keyer, it can be rotated to have the dot lever at the right or left — a real ambidexter! As a matter of fact, it can be rotated into any convenient angle depending on the orientation of the keyer, which, during mobile or portable work can be just about anything imaginable.

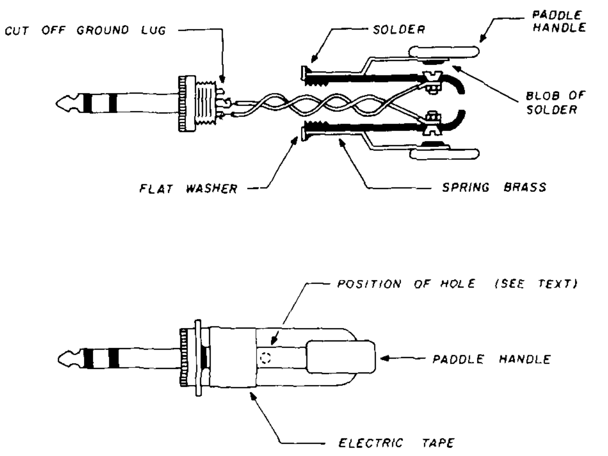


fig. 1. Construction of the Ambidextrous Paddle for Electronic Keyers or APEK. The paddle is built completely from junk-box parts, as described in the text.

The keyer shown in the photograph was built into a box which fits snugly into the car's door box. The APEK sticks out vertically so you can reach it without effort. Another attractive feature is the fact that the APEK is compatible with dual-lever keying devices. Squeeze keying features, if provided, don't have to be sacrificed when using the APEK, a feature impossible to achieve with single-lever keying devices.

construction

Obtain a standard three-connector jack plug and remove the shell. Cut off the ground connection close to the jack body. Drill and countersink two opposing holes in the plug shell about 8 mm (5/6 inch) from the unthreaded end. The holes should be suited to take two 2.3 mm by 4 mm (3/56 x 5/32 inch) countersink machine screws.

Take two small solder lugs and connect about 60 mm (2-1/4 inch) of thin flexible hook-up wire to each. Fit one of these lugs to each screw on the inside of the shell and secure with a nut. Find a suitable flat washer which will fit over the threaded part of the jack body and having an outside diameter slightly larger than that of the shell.

Prepare two strips of spring brass approximately 0.5 mm (1/64 inch) thick, 5 mm (3/16 inch) wide and 40 mm (1-1/2 inches) long. De-burr the edges and bend them into the shape shown in **fig. 1**. Solder these strips to the rim of the flat washer in opposing positions. This step can be simplified by temporarily assembling the plug. Mark off, on the underside of each strip, the positions of the screw heads fitted to the shell.

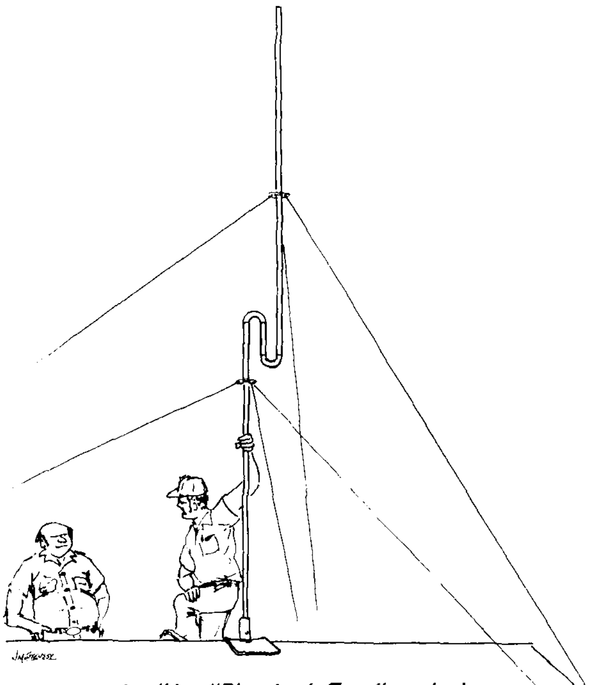
Disassemble the plug, clean and tin the positions marked off on the strips. Form a small blob of solder at these points and file them lightly to give a flat contact surface. Solder the two pieces of hookup wire, already connected to the screws on the shell, to the inner and outer lugs of the jack. Leave these wires long enough to survive the twist they will get when putting the jack together.

Shape two pieces of old PC board according to taste or as shown in **fig. 1** to form the paddle handles. Fix them to the brass strips with a small amount of epoxy. Finally, one or two wraps of electric tape cut to the correct width should be wound onto the jack shell. This will provide the necessary damping of the otherwise too springy brass strips.

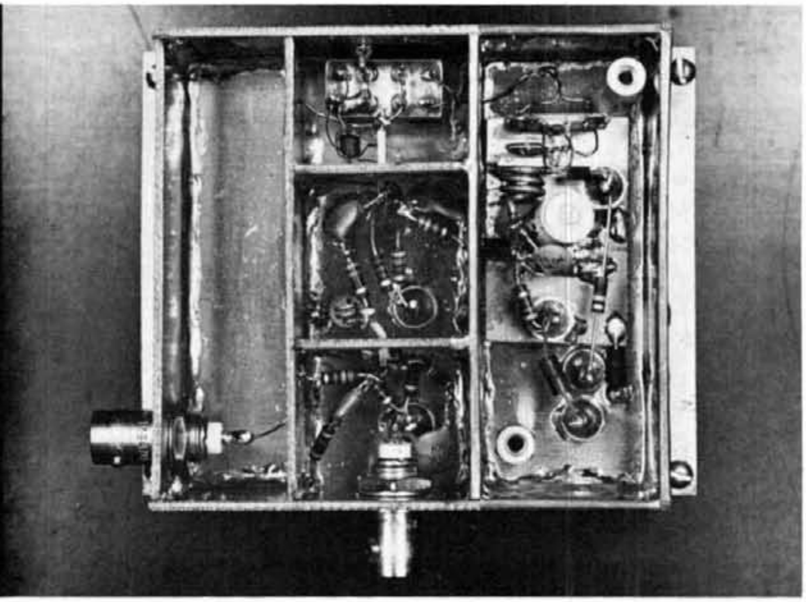
final adjustment

Contact spacing and spring tension can be adjusted by means of long nosed pliers. If you find that the spring tension is too high, a small hole drilled through the spring at the position indicated should do the trick. Then determine by experiment which side of the APEK activates dots or dashes and mark it for your convenience.

ham radio



I call it a "Plumber's Trap" vertical.



spectrum analyzer tracking generator

What is a tracking generator? When used with a spectrum analyzer it generates a CW signal corresponding to the frequency at which the spectrum analyzer is tuned. It's useful for looking at filter response. Filter blowby and undesired responses can be readily observed.

The generator described here was built for use with the spectrum analyzer described in reference 1. A tracking generator identical to this one is now being used with a Hewlett-Packard 8554/141. The tracking generator can be used with almost any spectrum analyzer that provides first-local oscillator output and has a first i-f of 200 ± 20 MHz.

Because of the tracking generator's oscillator instability, narrow-bandwidth measurements can't be made, such as the bandpass response of crystal filters. Nevertheless, it's useful for measuring responses at other frequencies (parasitic resonances) often found in crystal filters.

circuit description

The schematic is shown in **fig. 2**. Q1 and Q2 provide some gain and isolation between the 200-MHz oscillator, Q3, and the first i-f amplifier of the spectrum analyzer. R1 provides fine tuning and should be located for easy access by the operator. MX1 mixes the 200-MHz output with the signal from the first local oscillator to provide the 100-kHz to 100-MHz tracking signal. An optional 130-MHz lowpass filter is shown. The lowpass filter attenuates the 400-500 MHz component generated by the mixer in the tracking generator.

The tracking generator is built in a box made from 1/16-inch thick (1.5mm) copper-clad board. The same board is used as separators between stages. Paper-thin copper, available from hobby shops, is wrapped over the surface where the cover for the 200-MHz oscillator attaches. The other stages do not have shield covers. A blank compartment is available for the optional 130-MHz lowpass filter. **Fig. 1** shows the layout.

The oscillator is built on a separate piece of copper-clad board, 7/8 x 1-1/2 inch (22x38mm) and

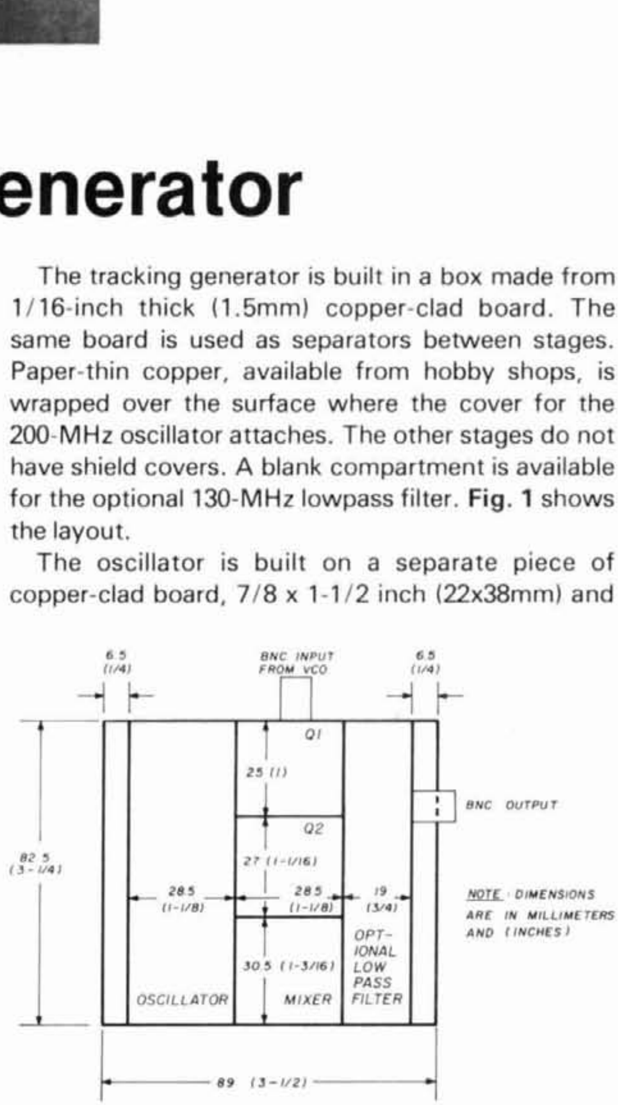


fig. 1. Circuit-board layout. Shielding is important for the 200-MHz oscillator compartment (see text).

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is held in place with double sticky-back tape. Two ferrite beads are strung on the ground wire that connects the oscillator ground to the compartment ground to minimize ground-loop currents, which can cause radiation from the oscillator. The mixer can be a standard mixer or the home-made mixer described in reference 1.

Dimensions shown inside are inside dimensions. The assembly is 1 inch (25.5mm) high. The 1/4 inch (6.5mm) overhang at each end is for mounting.

operation

The spectrum analyzer vfo output is connected to the tracking generator input of the spectrum analyzer.

1. Set spectrum-analyzer bandwidth to 250 or 300 kHz and scan width to 100 MHz.

2. Set R1 to the center of its range.

3. Adjust C1 so that the baseline on the spectrum analyzer shifts up.

4. Move L2 away from L1 until the baseline starts to move down.

5. Install the shield on the 200-MHz oscillator and adjust R1 so that the baseline is as high as possible, consistent with a flat response. As the tracking generator is being used, the 200-MHz oscillator will drift, and some readjustment of R1 will be required.

design considerations

One of the more challenging problems here is preventing the 200-MHz oscillator from radiating into the first i-f. If this happens, the baseline on the spectrum analyzer will shift up without a signal input. The signal from the oscillator will leak through MX1 backward through buffers in the tracking genera-

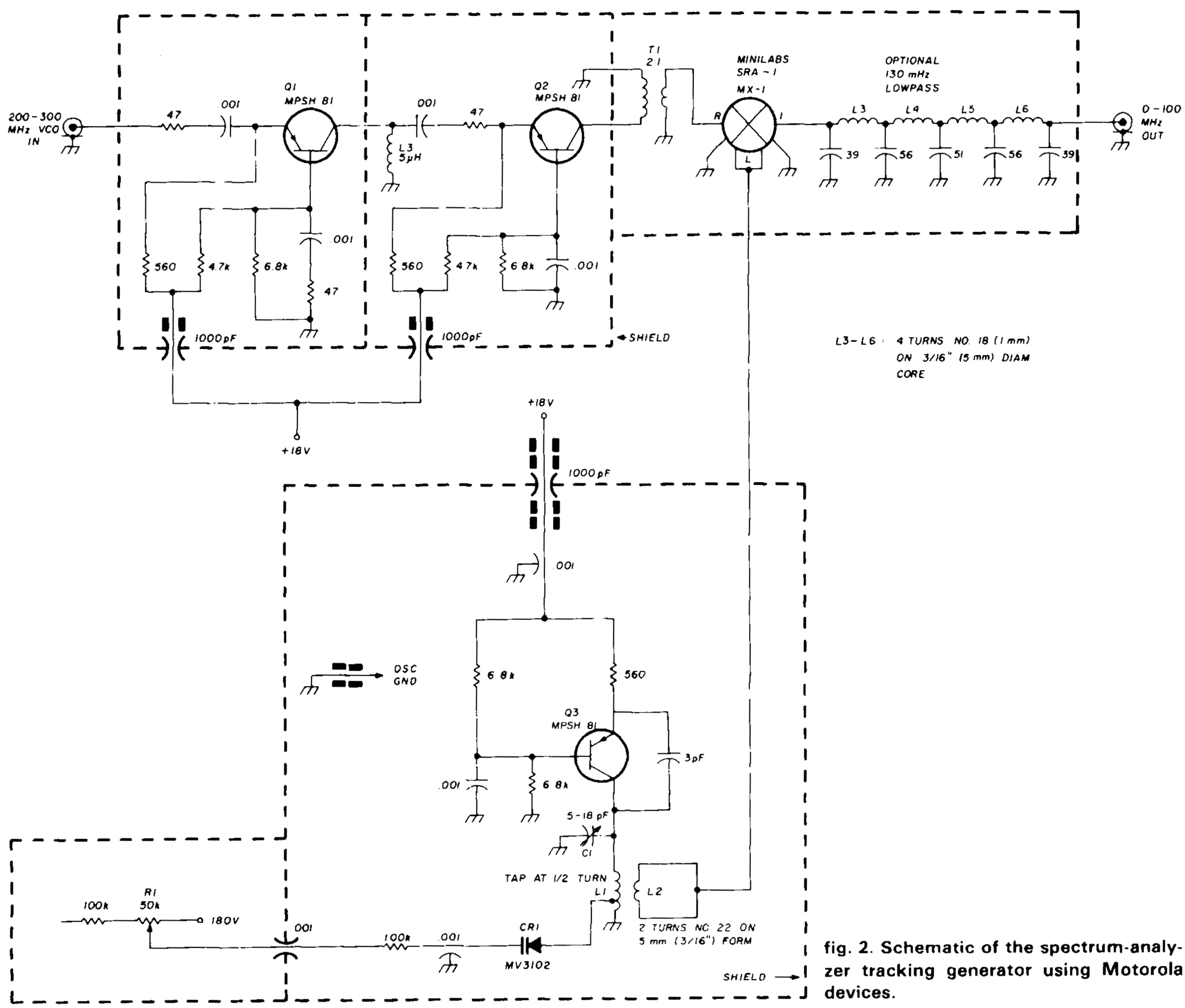

fig. 2. Schematic of the spectrum-analyzer tracking generator using Motorola devices.

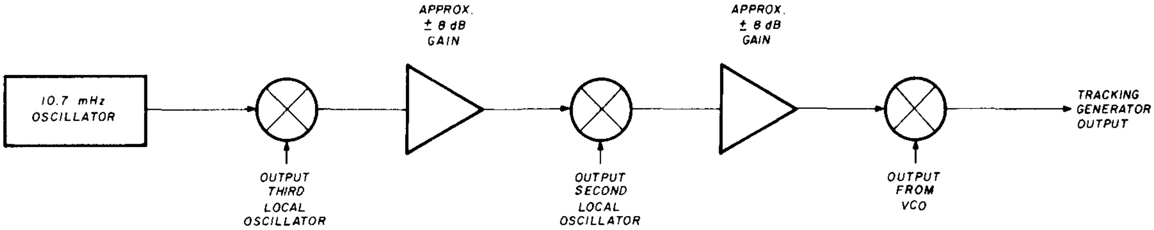

fig. 3. Suggestion for a tracking generator for measuring narrowband signals. Oscillators within the spectrum analyzer are used. A highly stable signal is provided.

tor, backward through any buffer in the vco and will follow the path of the vco to the first mixer in the spectrum analyzer and leak through the first mixer into the first i-f. This is not surprising, since the output from the oscillator is about 0.5 volt and the spectrum analyzer sensitivity is only a few microvolts. Also, the 200-MHz oscillator can radiate directly into the spectrum analyzer if both are not carefully shielded.

The tracking generator described here was mounted on the outside back of the spectrum analyzer to minimize coupling. If your spectrum analyzer has a phase-locked vco, or if the vco is stabilized in some other way, you might consider a different form of tracking generator for measuring narrow bandwidth. This method uses the local oscillators within the spectrum analyzer and provides a more stable tracking-generator signal. The block diagram (**fig. 3**) assumes that the last i-f is 10.7 MHz.

reference

1. Wayne F. Ryder, W6URH, "High-Performance Spectrum Analyzer," *ham radio*, June, 1977, pages 16-30.

ham radio

zip-cord feedlines

Many years ago I watched as someone set up a rig in the desert. The power plant was set out, the rig was set up, and the antenna was strung out between two convenient cactus plants (cactus attain a respectable height here). The thing that caught my eye was the feedline. It consisted of a long length of garden variety zip cord. My funny look at it gained a quick assurance that lamp cord was a perfectly good feedline.

Everything seemed to work just fine and the operators had no trouble working out on the band of the day, which was 75-meter phone. In those days it was common to run 100 to 200 watts input with high level a-m.

After a recent move I decided to put up a 75-meter antenna at home and use the most economical feedline. In the process of getting on the air, intermediate forms of antennas were used. That means some wire of about the resonant length was thrown up on the roof, and the near end was run into the rig. It was somewhat of a disappointment to discover that the intermediate antenna worked better than the well elevated final installation. A little checking showed that when the feedline itself was loaded up, it worked better than the antenna it was supposed to be feeding. In short, it appeared that something was not quite right.

A little further investigation revealed some rather interesting facts. RG-8/U, RG-58/U, and lamp cord were tested at 4 and 21 MHz. The rig was tuned up into a terminating type of wattmeter and the feedline under test was inserted between the rig and the meter. Power out with 60 cm (24 inches) of feedline *vs* power out with 9 to 18 meters (30 to 60 feet) of feedline was measured and the results were tabulated.

RG-58/U coax showed a 58 per cent loss at 21 MHz; it showed almost no attenuation at 4 MHz. RG-8/U gave about 12 per cent loss at 21 MHz.

Zip cord looked like it was best suited for use on the other end of a lamp or soldering iron. If you really want to know, it showed about a 60 per cent loss at 4 MHz. There was no need to measure it at 21 MHz!

About 20 meters (65 feet) of coax was used in the tests, and only 9 meters (30 feet) of zip cord (the rest of the zip cord was still attached to the antenna). Obviously, more lamp cord would have shown more loss. These tests were not conducted under laboratory conditions, but the variables were held to a reasonable level so that it was possible to reach a reasonable conclusion.

In any case, the zip cord came down and the RG-58/U went up. A crosstown telephone call got K7OXS on the air, again. Bill indicated that my 3-5 watt rig was back up to its normal signal strength at his location. He was almost as glad as I was that my antenna problems were finally resolved.

Evert Fruitman, W7RXV

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AM-366

More Details? CHECK – OFF Page 126

constant-current battery charger for portable operation

Tired of pulling out your nicads for charging? Here's a neat method for keeping your batteries fully charged at all times in any location

The use of portable battery-operated equipment requires fully charged batteries to obtain maximum usefulness of the equipment. For handheld transceivers, keeping the batteries fully charged would normally require frequent overnight charging. If the handheld is operated in a mobile environment it must be removed often from the car to the battery charger. However, it's possible to charge the equipment's batteries while in use, thus always assuring a full charge and maximum lifetime from the batteries between charges.

A charging system has been developed that permits constant-current charging from an automobile electrical system to a 10-cell nicad battery pack, which is commonly found in handheld and similar portable transceivers. Since a 12-volt nicad battery is not easily charged from an automobile 12-volt system, a special circuit was designed to furnish a charging current to the nicads.

nicad charging systems

The most obvious method of charging a nicad battery is shown in **fig. 1A**. A source potential, $V1$, delivers current into a battery pack whose potential is $V2$. Resistor R limits the charging current to a safe and maximum value. The charging current, I, is defined by

$$I = \frac{V1 - V2 - 0.6}{R}, \quad V1 > V2 \qquad \textbf{(1)}$$

This simple circuit works only if $V1$ is greater than $V2$, which may not be true in all instances. For example, if a nicad battery has a discharge potential of 10 volts, a 12-volt source can be used as a charging source. However, as the nicad battery becomes charged, its potential will increase to 13.5 volts, in which case the charging current would actually stop flowing during the charge cycle. In an automobile environment, source potential $V1$ is not constant but may vary from 11 to 14 volts depending on the condition of the automobile battery and engine rpm. Although the worst-case maximum voltage condition can be assumed, and the maximum charge current

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defined and limited, this simple circuit will not charge a nicad battery with a constant current all the time.

If a large value of $V1$ is available, the circuit in **fig. 1B** works well. Since $V1$ is much larger than $V2$, the battery potential, the change in $V2$ versus charge time has little effect on the charging current. A lamp is used to decrease the voltage to the nicad at a specified current. This circuit is representative of many chargers on the market today. Its disadvantage is that the charging current is not exact and $V1$ must be much greater than $V2$, which is not the case in an automobile system.

Fig. 2 shows an excellent method of obtaining a constant-current from a common three-terminal voltage regulator. The three-terminal voltage regulator would normally have its common terminal grounded and would deliver a constant voltage between the output and common terminal. However, if the common line is not grounded, but left floating, and a fixed resistance is connected from the output to the common terminal, the regulator will try to furnish a fixed voltage across the resistance. The current through the resistor is given by Ohm's law:

$$I_{OUT} = V/R$$

where I_{OUT} is current through resistor R; V is the voltage regulated by the device.

Once the circuit is completed between the three-terminal regulator's common and power return (ground connection), current will flow through this connection, even if a resistance or voltage exists in the path to ground. The only requirement is that the input voltage must be equal to or greater than the full charge battery potential plus 5 volts (for a 5-volt regulator) plus 2 volts (overhead voltage). Thus, for a 13-volt full-charge nicad potential and a 5-volt-type voltage regulator, the input voltage to the current regulator circuit must be greater than 20 volts.

In the circuit shown, a transformer and rectifier were used to provide 30 volts dc. A 50 mA, 6-volt lamp was used in series with the regulator to indicate when the 50 mA charge current was flowing; it also dropped the power-supply voltage to 24 volts dc.

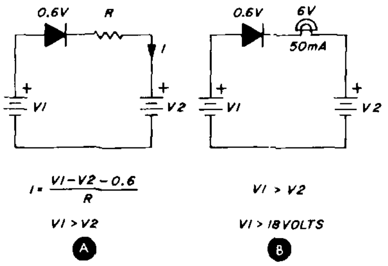

fig. 1. Simple charging systems for nicads. In *A* a resistor is used to limit the current source; the current is not constant and is nonlinear. Sketch *B* shows a lamp current-limited source. Voltage *V1* must be much larger than *V2* for good regulation. This is a very popular charging method.

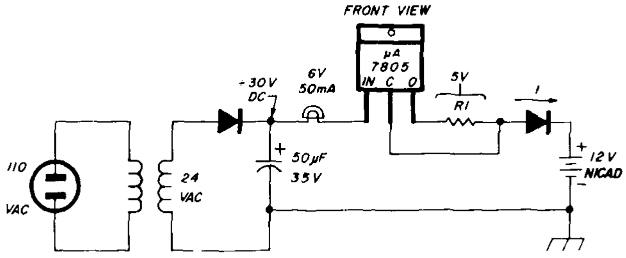

fig. 2. An excellent constant-current regulator using a popular 3-terminal voltage-regulator IC. Current source is determined by R1. For a 7805 IC, V is 5 volts. Thus for R = 50 ohms, I = 50 mA constant current.

The use of three-terminal regulators is an excellent technique for defining a constant current level. The current level is easily adjusted by changing resistor $R1$ in the output circuit. The current is essentially constant regardless of the discharge state of the nicad battery.

voltage-doubler circuit

To use a constant-current source as described it's necessary to provide an input source voltage of at least 20 volts. In an automobile situation, a 20-volt source is not available, so the circuit of **fig. 3** was designed. This circuit uses a NE-555 universal timer IC and two power transistors in a voltage doubling circuit. The output voltage is roughly twice the input voltage, minus any diode voltage drops. Thus, a 10-volt source will be converted to 19 volts and a 13-volt source to 25 volts. This doubled voltage is then used to drive a source current into a three-terminal current regulator.

The operation of the voltage doubler is as follows, referring to **fig. 3**.

The NE-555 is used in the common astable multivibrator configuration. The oscillation frequency is determined by $R1$, $R2$, and $C1$ and is equal to

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

To have a near 50 per cent duty cycle the ratio of $R1$ to $R2$ should be around 1 to 4 as shown.

The 555 astable oscillator drives a pair of complementary transistors. High-current power transistors were used to switch the large charge and discharge currents. Transistor Q1 charges capacitor C2 to the input voltage during the first part of the astable cycle. During the second half of the cycle, transistor Q2 puts the fully charged capacitor C2 in series with the supply voltage. Typical values for C2 are

$$C2 = \frac{I_{OUT}}{F} \qquad (3)$$

where F was defined earlier and I_{OUT} is the constant output current required. A diode and capacitor (C3) filter the pulsating dc to a value nearly equal to two times the supply voltage. The value of C3 should equal C2.

Transistors Q1 and Q2 are plastic-cased power transistors. Power transistors are used because of the high peak currents during the charge and discharge cycle of C2. Although MJE 2955 and MJE 3055 types are shown, most common power transistors will work. The transistors are not heat sinked because both are operated in the saturated mode, thus power dissipation is held to a minimum. The diodes are 1-amp silicon types and may also be substituted with similar devices.

The current regulator operates as the one previously described. It may be attached to a heat sink, but is not mandatory. If the regulator's input terminal voltage is 25 volts, and the output terminal is at 17 volts, only 400 mW of heat must be dissipated in the worst case. The oscillation frequency of the voltage-doubling circuit can be increased, but transistor switching becomes less efficient with the increased switching speed. A too-low frequency requires larger values of capacitance for C2 and C3. A good compromise is to make the frequency between 1 and 10 kHz.

In the circuit shown the switching frequency was set to 1.4 kHz. Since ten 500 mA-hr nicads were being charged, the charging current (I_{OUT}) was set to 50 mA. For an input voltage range of 10-15 volts, the charging current was a constant 50 mA.

practical approach

To make these charging schemes work in a mobile charging system, charging current may be delivered

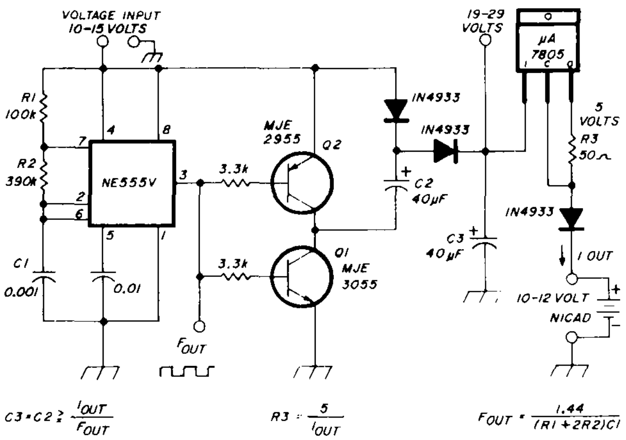

fig. 3. Voltage-doubler circuit suitable for developing enough voltage from an auto battery to run the current regulator. Power transistors may be substituted for those shown.

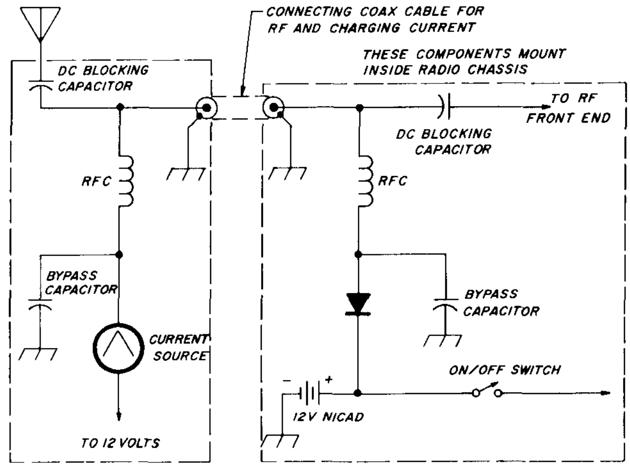

fig. 4. Use of the constant-current charging source requires only the connection of the antenna lead to the radio in a mobile or base station. The current source may be located inside your power amplifier or within an ac-operated charger. For 2 meters, the rf choke, RFC, is ten turns no. 26 AWG (0.3mm) copper wire on a 1-megohm 1/2-watt composition resistor.

to the portable unit through the antenna coaxial cable. This approach makes it easy to connect the handheld or other equipment to the charging source as well as to the antenna.

It's quite common to use small power amplifiers in a mobile system. This creates a convenient location for the current source circuits. **Fig. 4** shows a typical mobile setup. The current source is located inside the amplifier housing. A small rf choke isolates the incoming and outgoing rf energy from the current-source circuits. A small capacitor in series with the antenna confines any charging potential to the inside of the automobile. The coax connecting the handheld to the amplifier or antenna carries the charging current to the batteries through an isolating rf choke. A dc blocking capacitor should be used in series with the rf circuitry. Most transceivers employ such a capacitor; check your schematic. The charging current source should be connected across the battery at all times.

advantages

By charging the nicads through the coax line, the batteries can be maintained at full charge at all times.

By using the a-c operated charger and the automobile charger system, the nicads can be charged at any location. The current source could be switched to a lower trickle charge if desired. A 3-position switch could be used for this purpose and for turning off the regulator.

ham radio

how to modify linear amplifiers for full break-in operation

With the introduction of the Ten-Tec *Triton IV*, which provides full break-in operation on CW, I needed a companion linear amplifier which also had this very desirable operating feature. At the present time, however, there is only one commercially produced amplifier on the market which meets this requirement — and it's priced at almost $3000. The purpose of this article is to outline the theory and give some typical circuits which can be used to modify any power amplifier to provide full break-in capability. The circuitry is for use with grounded-grid triode linear amplifiers; class B or AB is assumed.

While class C might seem to be a better choice for a CW power amplifier, there are several reasons to maintain linear operation. First, and most important, since class C operation is not linear, the CW waveform supplied by the exciter will be distorted and the resulting output wave may produce serious key clicks and other unwanted spurious signals.

For the modification described here the grid of the amplifier tube must be at chassis ground, with a positive voltage applied as cutoff bias to the cathode. The classical biasing scheme, with the cathode at ground, **fig. 1A**, is rearranged to place the grid at both rf and dc ground, **fig. 1B**. The operating or cutoff bias is in the form of a positive voltage which is applied to the cathode; the high-voltage B+ plate supply is above ground by this potential.

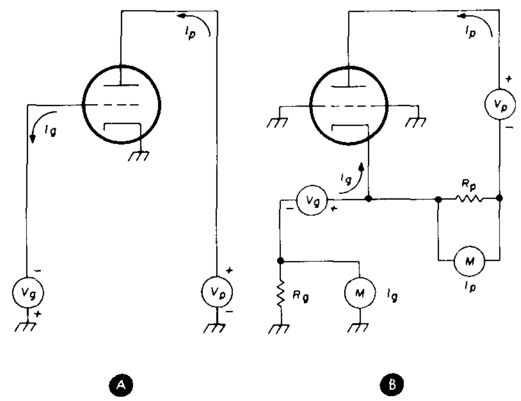

fig. 1. In the classical bias circuit, A, the cathode is grounded and a negative bias is applied to the grid; in the grounded-grid circuit a positive bias is applied to the cathode, B. In these circuits V_g is the bias supply, V_p is the plate supply, R_g is the grid meter shunt, and R_p is the plate meter shunt.

When the amplifier is used for break-in operation, it is essential that the tube be biased to cutoff — with no plate current flowing — so the tube doesn't generate noise which would mask weak received signals.

cutoff bias

Of the several types of tubes commonly used for linear amplifier service, there are two basic types: those with hot cathodes and those with indirectly heated cathodes (see **fig. 2**). For the purposes of this discussion the main difference is that the bias on a hot cathode is applied through the center tap of the filament transformer (**fig. 2A**); bias for indirectly heated tubes is fed directly to the cathode through an rf choke (**fig. 2B**).

Figs. 3 and **4** show two common ways to bias rf power amplifiers to cutoff. The circuit in **fig. 3** (from the ARRL *Handbook*), uses a 10k resistor in series with the cathode circuit to develop self bias. It is the least acceptable method, however, because it requires a certain amount of plate current flow to provide the bias. Therefore, the tube is not completely cut off.

Another cutoff bias circuit is shown in **fig. 4**. It uses a + 150 volt bias supply which cuts off all plate current and powers the T/R relay, K1. Actually, for B + supplies up to about 3500 volts, 75 volts of bias is enough to cut off a pair of 3-500Zs — 75-volt relays, however, are hard to find.

Fig. 5 shows a circuit which uses an rf-sensing transistor switch to remove the cutoff bias from the amplifier. Under standby conditions Q2 is an open circuit and 50 volts cutoff bias is applied to the cathode. When the drive signal is applied, Q2 turns on

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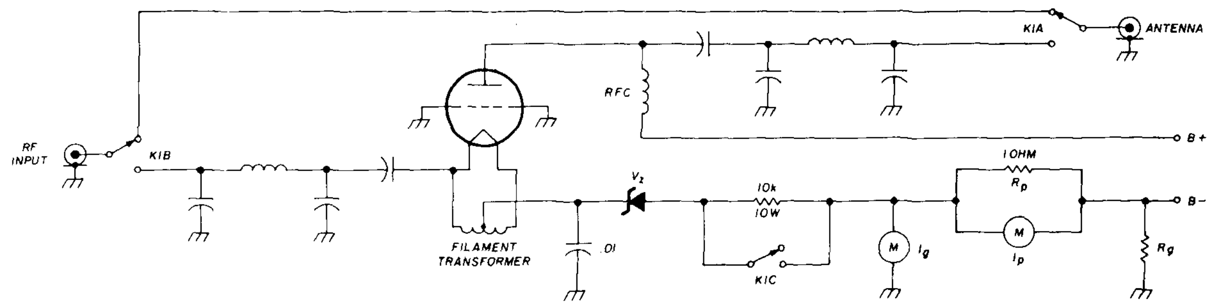

fig. 3. Typical linear amplifier bias and control using self bias developed across the 10k resistor in the B- line. Relay K1 is shown in the operating position with the amplifier on.

and the operating bias becomes the sum of the zener voltage, V_Z, Q2's collector-emitter saturation voltage, $V_{CE(SAT)}$, and the voltage drop across the grid meter shunt, R_g. $V_{CE(SAT)}$ for the specified 2N5321 resistor is 0.8 volt at 500 mA.

When selecting Q2, two criteria must be met: a low $V_{CE(SAT)}$ because this adds to the operating bias, and a maximum collector-voltage rating that is greater than the applied voltage bias voltage. The Darlington configuration is used because transistors which meet the two voltage requirements seldom have the necessary current gain for this circuit. The power requirement for Q2 is quite low because the voltage impressed across it is low when it carries the grid current. A power dissipation rating of 10 watts is more than adequate. The resistance of the grid meter shunt resistor, R_g, should be as low as possible; usually 0.5 ohm is sufficient with a grid current meter which has a 1 mA movement.

The basic bias circuit of **fig. 5**, modified to accommodate the directly heated cathodes of two 3-500Zs, has been used successfully in a Heath SB-220. Note that the zener diode no longer has to pass the full plate current so it is permissible to use a less expensive 20-watt zener diode. An added feature of this circuit is that if any of the devices which supply operating bias fail open, cutoff bias will be applied to the tube — this prevents any damage that might otherwise occur.

The 1N4004 diode shown in the negative high-voltage line is a clamp which prevents the B− from going negative with respect to the chassis in case of bias supply failure.

When the linear amplifier is in the standby condition virtually no current is drawn from the bias supply. When Q2 turns on, resistor R1 limits the current flow. This current will show as grid current, so it's a good idea to keep R1 as large as possible. When testing this circuit I found it was useful to provide a means for turning on Q2 without applying rf drive power; S1 is a test button which does this.

In normal operation, the linear appears to be running class C because plate current is drawn only when an rf drive signal is present. The amplifier is not operating class C, of course, but since quiescent plate current is no longer being drawn under no-signal conditions, the average power dissipated by the tube due to quiescent bias (200 to 600 watts in a typical class AB amplifier) is greatly reduced; this increases tube life, improves reliability, and results in a cooler ham shack.

receiver switching

When the power amplifier can be operated without using relays to switch the bias, one step remains: receiver antenna switching. Electronic T/R switches have been available for years, but have never been widely accepted because, when the transmitter is tuned to resonance, the tank circuit acts as a "suck out" filter which reduces received signal strength. The electronic T/R switch was not the culprit — its placement in the system caused the problem. The solution is to place the electronic T/R switch at a high impedance point in the system. If you want to eliminate noisy relays which are slow and prone to failure and go to full break-in operation, the antenna

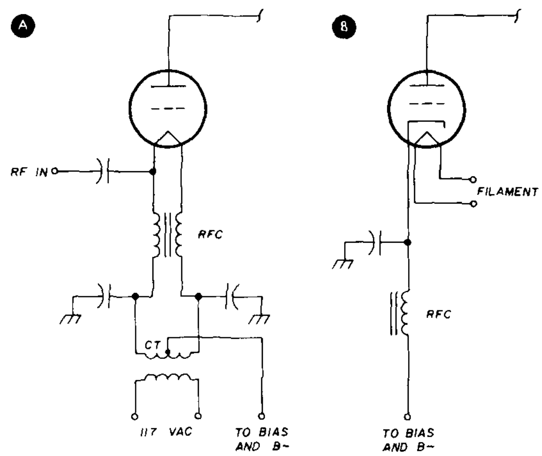

fig. 2. Grounded-grid biasing arrangements for hot cathode tubes, A, and indirectly-heated cathodes, B. Typical hot-cathode tubes are the 3-500Z, 3-1000Z, and 3CX1000; the 8877 and 8874 series of power amplifier tubes have indirectly-heated cathodes.

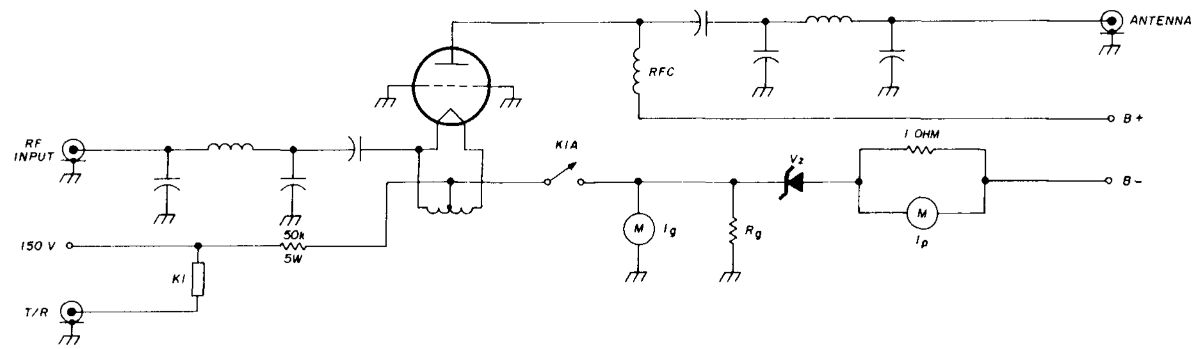

fig. 4. Linear bias and control using an external bias supply; this method provides better operation than the arrangement of *fig. 3*. Relay K1 is shown in the operating position with the amplifier biased to cutoff.

must be connected to the transmitter at all times. Thus, to find a high impedance point to connect an electronic T/R switch, the plate tank of the amplifier is a good choice.

Coupling to the plate tank is not without its problems, however. First of all, it's dangerous because of the high voltage that is present. Secondly, since there is a large amount of rf voltage present, the coupling capacitor C1 (**fig. 6**), must be small both to reduce its effect on the amplifier tuning and to assure that the T/R switch is not overdriven. This presents a problem because the optimum value of C1 for receiving purposes, about 5 pF, is too much capacitance when used at the kilowatt level on transmit. Therefore, a compromise is made on the side of safety and reliability.

The capacitor specified is made from RG-8/U coaxial cable with the center conductor overlapping the braid for approximately 12 mm (1/2 inch). Two 25 mm (1 inch) square tabs, spaced 1 cm (3/8 inch), will also work if the mechanical layout of the amplifier will permit placement of this arrangement. Be sure that this capacitor is placed directly across the plate tuning capacitor *after* the plate blocking capacitor. This compromise results in a slight loss of gain on receive, but most transceivers have more than enough front-end gain to make up for the loss.

construction

Since most modern kilowatt linears use some sort of input tuning, the placement of S1 in **fig. 6** is not too difficult. The switch wafer can usually be added to an existing switch shaft, or the shaft can be extended to provide room for the wafer. When pur-

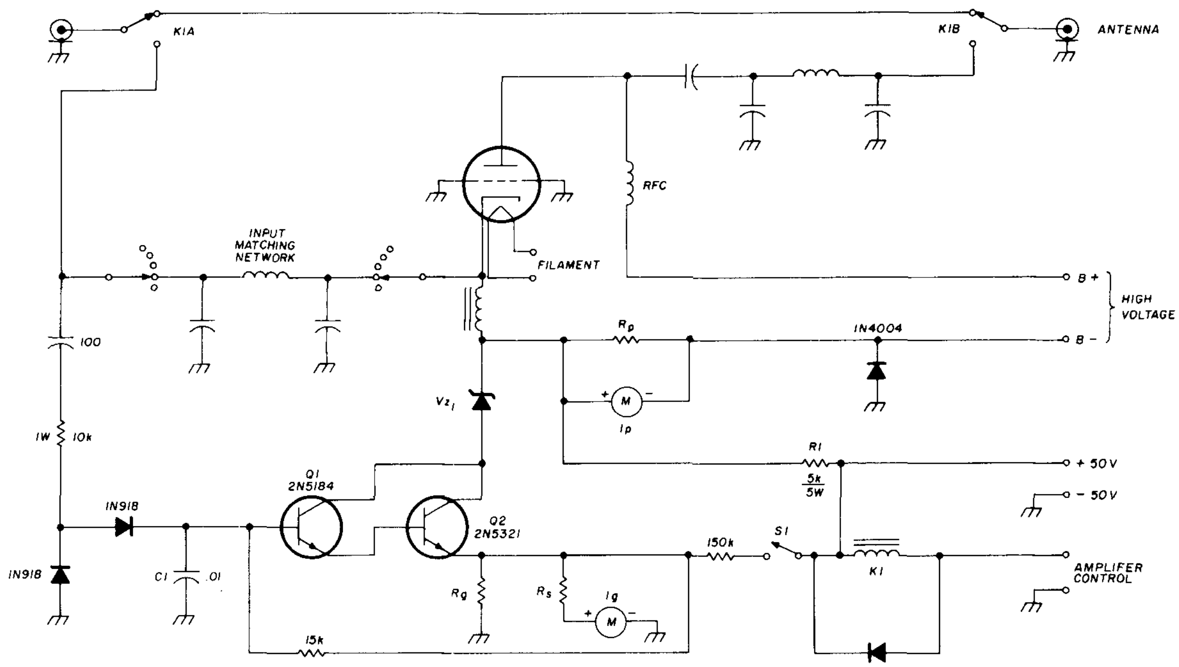

fig. 5. Cutoff bias circuit for a linear amplifier which uses an rf sensing transistor switch. The zener voltage, V_z, provides the proper operating bias and was part of the original amplifier circuit. Cutoff bias circuit operation is described in the text.

chasing the wafer switch for S1, be sure to note whether the amplifier uses 30, 45, or 60 degree indexing between switch positions.

The location of Q1 and its associated circuitry is not too critical except that it must be reasonably close to both C1 and S1 to minimize losses. It should be shielded if it is located inside the plate tank enclosure and all dc leads must be shielded. Since the dc current drawn is very small, the + 18 volts can be obtained with a voltage doubler from the amplifier filament supply, or from a zener regulated drop from the positive bias supply. The switched coils form broadly resonant tuned circuits with the existing circuit capacitance, so some adjustment of the given values may be necessary.

Triton IV modifications

There are also modifications to the Triton IV which will improve performance when using a linear amplifier. First is a simple change to eliminate the delay of the control relay while using ssb. As built at the factory, the time constant capacitor, C3 in **fig. 7** and page 3-34 of the manual, is tied directly to ground on the circuit board. By lifting the ground end of C3 and taking it to the unused center pin on the board, a wire can be run to the mode switch, S1E (CW1 and CW2 positions), which will activate the delay only when using CW. S1E is part of the rear wafer, closest to the chassis.

The second modification involves removal of the control relay, K1. This should be done only if the amplifier uses a positive voltage to key its control and changeover relay. The reason for doing this is to eliminate the turn-on delay which is caused by using one relay to activate another. Relay K1 has no function in the operation of the Triton by itself. When the key is closed, it takes about 25 milliseconds for this relay to close, and perhaps another 23 ms for the amplifier relay to pick up, before the output of the Triton is amplified. This causes key clicks on the first dot. The transistor used for Q4 depends on the voltage used on the amplifier control relay and the relay current. With K1 removed, there is plenty of room for the added components on the board.

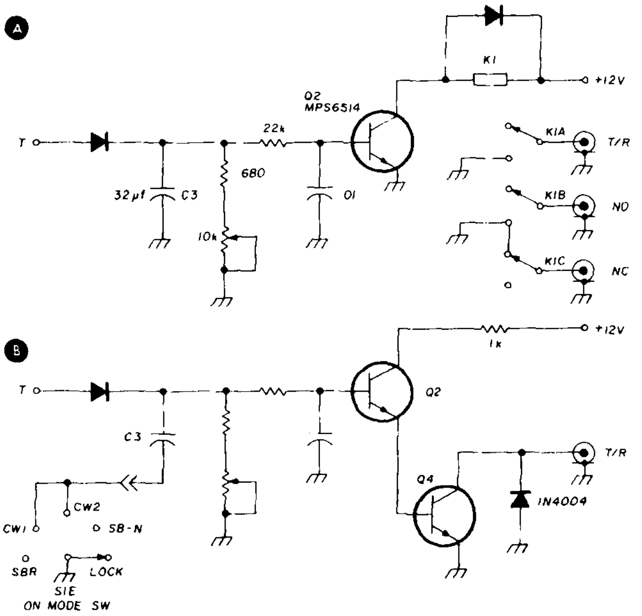

fig. 7. Modification of the Triton IV to remove the control relay and to eliminate unwanted delay times during CW break-in operation.

Some Triton users have complained of an ac hum in the receiver when using a linear amplifier — especially in the narrow CW-1 position. This is caused by ac ripple (or raw ac) on the relay control line. When attached to the Triton T/R relay's normally-open jack, unshielded wires going to K1 pass underneath the CW filter which picks up the hum. The solution is simple: reroute the wiring between K1 and T/R normally-open jacks on top of the chassis. This change has been incorporated in late production Tritons.

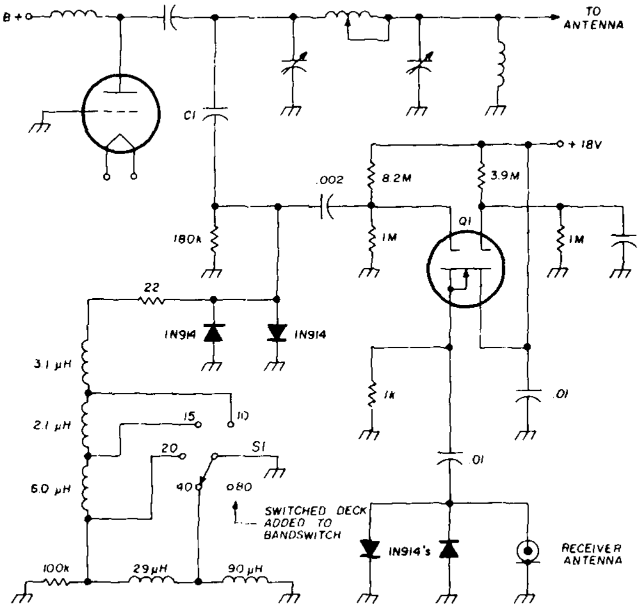

fig. 6. Electronic T/R switch for the receive antenna. The station antenna is connected to the linear at all times. Capacitor C1 is less than 1 pF and consists of 12 mm (1/2 inch) of RG-8/U coaxial cable (see text).

conclusion

What has been presented here are notes and basic technical information needed to modify an existing rf power amplifier for full break-in operation with a Triton IV; the circuits can also be adapted to other transceivers. Ten-Tec in no way assumes responsibility for the use or misuse of this information, nor for any damage to other manufacturer's equipment resulting from implementation of these circuits.

ham radio

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Model 1359 Drake UV-3E (144-430)*

(Models above include factory installed modules for bands as listed, standard dynamic mike, and mobile mounting bracket.)

* *This model tuned for European fm bands. See your dealer for price details.*

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Model 1504 Drake PS-3 AC Power Supply..... **$ 89.95**
Model 1525 Drake 1525EM Encoding Mike.... **$ 49.95**
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NOTE: Certain of the above models will be available before others. Check with your dealer for specific availability.

how to design matching networks

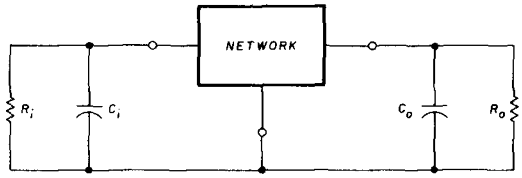
Six basic impedance matching networks and how to design them for your own applications

Common L, T, and pi networks work well for impedance matching but they lack the selectivity required for amplifiers or frequency-multiplier chains. Adding a component to the network and rearranging allows it to be selective on both sides of band center. Presented here are six simple matching circuits and their design equations that allow adjustment of design *Q* for selectivity and different source or load impedances.

Fig. 1 shows the basic network with its source and load interfaces. The assumption is that R_i is less than R_o. Simply reverse the input, output, and the network if R_o is less than R_i. Both source and load are assumed to have shunt capacitances; this is usually true and will include stray capacitance as well.

The value of *Q* is the design *Q* of the network and determines selectivity. It is *not* component *Q* which should be at least five times design *Q*. Selectivity around band center is treated the same as a tuned circuit with a certain *Q*. Each of the six networks has different attenuation far from center; examples of this are shown later.

The constants listed in **fig. 1** reduce the size (and complexity) of the network equations, and will be used with all six networks. It must be emphasized that all capacitive reactances must calculate *negative*, inductive reactances *positive*. Any exception to this rule with a network indicates that particular network cannot be used.



R_i LESS THAN R_o

CONSTANTS

$X_i = -1/(2\pi F C_i)$ $\quad$ $X_o = -1/(2\pi F C_o)$

F = CENTER FREQUENCY

$N = \frac{R_i X_i}{R_i^2 + X_i^2}$ $\quad$ $R_x = X_i N$

$Q_i = R_i + X_i Q$ $\quad$ $Q_o = R_o + X_o Q$

fig. 1. The basic matching network. Constants shown here will be used in calculating the values for the six different networks discussed in this article.

different sources and loads

Fig. 1 shows a parallel capacitive reactance across the input and output terminals. Sometimes the end impedances are given (or measured) in series form. This may be converted to parallel form either by impedance-to-admittance plus inversion of resulting conductance and susceptance. Any of the calculators with rectangular/polar conversion can handle this easily. The following conversion formula can be

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used if those functions are not available on your calculator:

$$R_p = R_s + (X_s^2/R_s)$$
$$X_p = X_s + (R_s^2/X_s)$$

where
R_s = *Series resistance*
X_s = *Series reactance*
R_p = *Parallel resistance*
X_p = *Parallel reactance*

Note that the sign of the reactance is preserved in conversion.

An inductive reactance is a special problem. Compensation of this is done by capacitive shunting so that the total reactance at the band center becomes capacitive. Some of the networks will have shunt inductors at the ends. This condition allows the physical inductor to be the parallel difference with stray capacitance (always present) forming C_i or C_o. If the end reactance changes rapidly around the band center, it is better to add a physical capacitor and use calculated inductance directly. In any case, end reactances must be capacitive.

handling data from the spec sheet

Transistor data is invariably given in admittance or S-parameters. Admittance is already in parallel form so the conductance and susceptance values are taken directly and inverted to yield end resistance and X_i or X_o (watch the sign of susceptance, it is positive when capacitive).

S-parameters are a bit different and are found on Smith chart representations. These are normalized to 50 ohms or 0.02 mho and can be taken directly,

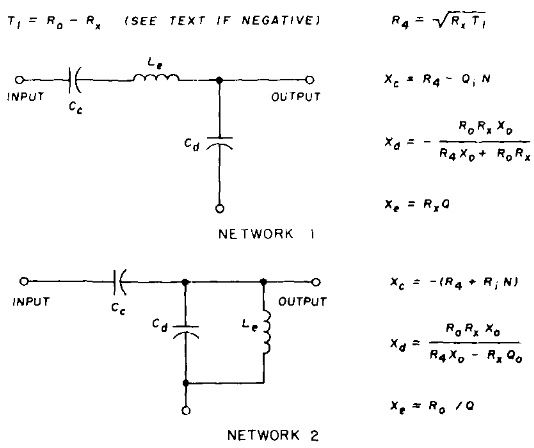


fig. 2. Networks 1 and 2.

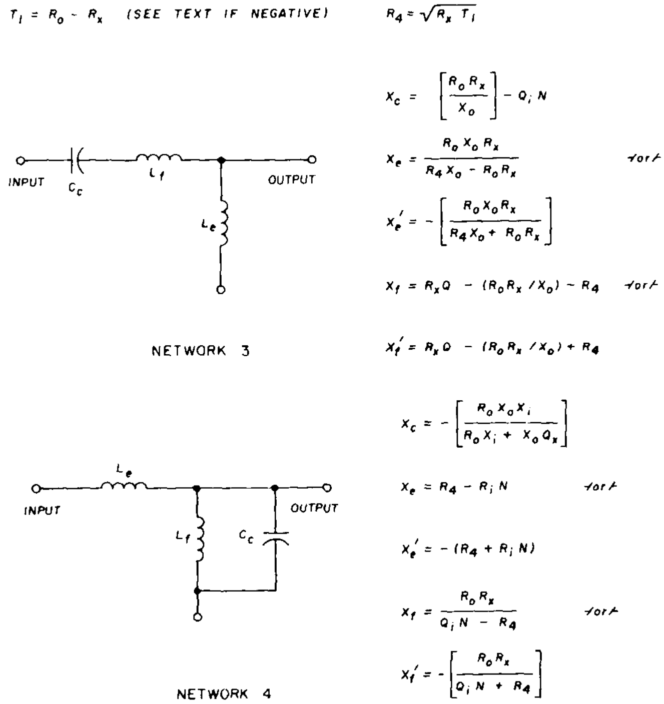


fig. 3. Networks 3 and 4.

then un-normalized. S_{11} is the input impedance while S_{22} is the output impedance; both are complex numbers. Knowledge of Smith chart interpretation is required.[1]

Data in admittance or S-parameter form is invariably frequency sensitive. They also represent the average data of a production run so individual devices can vary. Under such conditions it is well to keep the design Q relatively low. There may be an advantage to swamping the end with a parallel resistance (small carbon composition resistor) to reduce sensitivity of values. This has two cautions: power loss and about 2 pF extra capacitance with each resistor.

examples

Numerical examples all assume matching a 50-ohm line to the input of a Motorola MC1590 amplifier IC (single-ended) at 7.15 MHz with a design Q of 20. Stray capacitance of 3 pF is assumed at each end; the MC1590 is insensitive with a parallel, single-ended impedance of 5k and 5 pF. The line is assumed perfect. **Fig. 1** constants are then:

R_i	$= 50$	R_o	$= 5000$
C_i	$= 3\,pF$	C_o	$= 8\,pF$
X_i	$= -7419.81$	X_o	$= -2782.43$
N	$= -6.73841\ 10^{-3}$	R_x	$= 49.9977$
Q_i	$= -148346$	Q_o	$= -50648.6$

An HP-25 calculator was used for these and following calculations.

networks 1 and 2

These are shown in **fig. 2**. Variable T_1 is a test value and is also used for Networks 3 and 4. If T_1 results in a negative, none of the first four networks can be used with the given constants. Design Q and/or end resistances can be changed to make it positive. The negative situation only comes about with low R_o:R_i ratios.

For the first four, $T_1 = 4950$ so $R_4 = 497.483$. Values of Network 1 are then:

$$X_c = -502.135 \qquad C_c = 44.33\ pF$$
$$X_d = -613.263 \qquad C_d = 36.30\ pF$$
$$X_e = 999.555 \qquad L_e = 22.26\ \mu H$$

Note that all reactance signs are correct. Values for Network 2 are:

$$X_c = -497.146 \qquad C_c = 44.77\ pF$$
$$X_d = -605.847 \qquad C_d = 36.74\ pF$$
$$X_e = 250.000 \qquad L_e = 5.565\ \mu H$$

Reactance signs are correct here, too.

networks 3 and 4

Test variable T_1 and R_4 apply here. Both inductors of both networks have two possible solutions. The reactance sign rule still applies so *both* X_e and X_f must be positive or *both* X_e' *or* X_f' must be positive. Do not mix X_e' and X_f or vice-versa; use either all non-prime or all prime values. Network 3 will have:

$$X_c = -1089.46 \qquad C_c = 20.43\ pF$$
$$X_e = 425.637 \qquad L_e = 9.474\ \mu H$$
$$X_f = 592.318 \qquad L_f = 13.18\ \mu H$$

Non-prime values obey the rules as do Network 4 values:

$$X_c = -274.782 \qquad C_c = 81.01\ pF$$
$$X_e = 497.819 \qquad L_e = 11.08\ \mu H$$
$$X_f = 497.853 \qquad L_f = 11.08\ \mu H$$

The inductors in Network 4 came out very close to the same value. This happened with this particular example, but is not true of other conditions.

network 5

This one is shown in **fig. 4**. It must be noted that the two inductors have zero coupling and must be physically separate. The schematic may appear to be a conventional tapped-inductor circuit but such would need a different set of equations plus measurement of mutual coupling. Networks 3 and 4 must also have separate inductors.

Test variable T_2 cannot be negative. If T_2 is negative design Q or end resistances, and possibly input capacitance, may have to be changed. Example conditions fit alright so we get:

$$R_5 = 22.0617\ 10^9$$
$$T_2 = 827.815\ 10^9 \qquad R_4 = 64.3356\ 10^6$$
$$X_c = -274.782 \qquad C_c = 81.01\ pF$$
$$X_e = 227.822 \qquad L_e = 5.071\ \mu H$$
$$X_f = 28.7208 \qquad L_f = 0.639\ \mu H$$

Non-prime inductor values were correct in this example. Note that C_c is the same as for Network 4.

network 6

Fig. 5 shows this to be the tuned circuit often found in receiver front ends. The test variable is T_3 and there is only one set of solutions. Our example condition results in:

$$R_5 = 2.57302\ 10^9$$
$$T_3 = 89.9416\ 10^9 \qquad R_4 = 21.2063\ 10^6$$
$$X_c = -250.921 \qquad C_c = 88.71\ pF$$
$$X_d = -32.9475 \qquad C_d = 675.60\ pF$$
$$X_e = 250.000 \qquad L_e = 5.565\ \mu H$$

The inductor is the same value as in Network 2.

wideband response

Fig. 6 shows the frequency response over a two-decade range for the examples given. Voltage response has been calculated with constant end resistances. As such, it will be the same in either direction. Frequency is normalized to 7.15 MHz.

The joker in the deck is Network 3. The extra peak on the high side of resonance will vary in frequency and relative amplitude depending on the design Q and end impedances. A saving grace is that Network 3 has the best low-frequency attenuation. All networks will vary in other applications, primarily with different design Q; the general shape of the response curve, however, will still be the same.

Choice of a network depends on the application. A frequency-multiplier chain should consider Networks 2, 5, or 6 because of their better low-frequency attenuation. The last stage could use Networks 1, 4, or 5 in the output to reduce unwanted harmonics. An amplifier chain such as an i-f strip could alternate Networks 5 and 6 between stages for best skirt attenuation.*

*There are better ways; this would only be for miniature construction or multiple stages with degeneration of gain.

table 1. Design Q determination with fixed-value capacitors

Network 1, Fixed C_c: $Q_1 = \frac{R_4 - X_c - R_i N}{X_i N}$

Network 2, Fixed C_d: $Q_2 = (R_4/R_x) - \left[\frac{R_o(X_o + X_d)}{X_o X_d}\right]$

Network 3, Fixed C_c: $Q_3 = \frac{R_o R_x - X_o(X_c + R_i N)}{X_i X_o N}$

Networks 4 and 5, Fixed C_c: $Q_{45} = -\frac{R_o R_i (X_o + X_c) + X_c X_o R_i}{X_c X_o X_i}$

impedances presented to the source

The impedance presented to the source can be easily calculated.[2,3] Networks 1 through 4 have variations far from band center that might cause stability problems. Network 1 is inductive from the band center to about 20.75 MHz from the example. Parallel resistance climbs abruptly to about a megohm while parallel reactance changes swiftly to capacitive reactance. Network 2 has a similar, less abrupt change at about 9.76 MHz. Network 3 is also similar with the changeover coinciding with the voltage response peak at about 24.3 MHz.

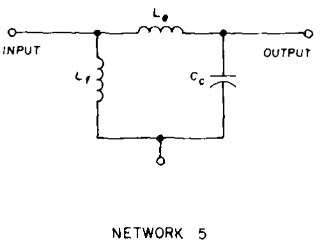
Network 4 becomes inductive below 4.9 MHz and capacitive above 28 MHz, resistance peaking to 4 megohms at that frequency. Networks 5 and 6 were much less susceptible to changes, and showed only slight variations at passband edges. All components were assumed lossless so a practical circuit would exhibit much less variation due to finite component Q.

The swamping-resistor method with transistor collectors works well from 6 meters and down with f_t of 150 MHz or greater. This increases a normally low-collector conductance but does have some power loss. A swamping resistor can also be used at the load since load changes reflect to the source. A

$R_5 = X_i^2 + Q_i^2$

$T_2 = R_i R_5 - R_o X_i^2$ (SEE TEXT IF NEGATIVE)

$R_4 = \sqrt{R_o T_2}$



$X_c = -\left[\frac{R_o X_o X_i}{R_o X_i + X_o Q_i}\right]$

$X_e = (X_i/R_5)(R_o Q_i + R_4)$ ✓or✓

$X'_e = (X_i/R_5)(R_o Q_i - R_4)$

$X_f = -\frac{R_o R_i X_i}{R_4 + R_o R_i}$ ✓or✓

$X'_f = \left[\frac{R_o R_i X_i}{R_4 - R_o R_i}\right]$

fig. 4. Network 5.

rule of thumb is to add 2 pF for every resistor, using only carbon composition types.

Mismatch loss will add out-of-band attenuation when used with receiver front ends. An exception is where the input impedance of the network matches the antenna impedance out of band. This is rare, but it happens. A similar condition occurs with different equipment connected with coax cable: the line length may cause a match out of band.

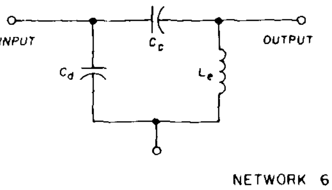
non-standard component values

A cure for non-standard component values is to change design Q slightly to accommodate at least one standard, fixed value. The other two compon-

$R_5 = X_o^2 + Q_o^2$

$T_3 = R_i R_5 - R_o X_o^2$ (SEE TEXT IF NEGATIVE)

$R_4 = \sqrt{R_o T_3}$



$X_c = -(X_o/R_5)(R_4 + R_o Q_o)$

$X_d = \frac{R_o R_i X_o X_i}{R_4 X_i - R_o R_i X_o}$

$X_e = R_o/Q$

fig. 5. Network 6.

ents can be trimmable. **Table 1** is a tabulation of design Q for Networks 1 through 5 based on one specified reactance.

Our example for Network 2 gives a C_d value of 36.74 pF. A 39 pF mica is a standard value with a reactance of −570.755 ohms at 7.15 MHz. This X_d value is used and gives a design Q of 20.5074. The X_c of Network 2 is unaffected by Q but X_e changes to 243.814 ohms or 5.427 μH.

Network 6 is a bit difficult to solve for Q but it can be done by programming an HP-25 or similar calculator to solve either X_c or X_d with manual Q input. An approximation that works in some cases is:

$$Q_6 \cong (R_o + R_i)/X_c$$

The Network 6 example had C_c at 88.71 pF. A fixed 82 pF capacitor will have −271.456 ohms so the approximate Q is 18.6034. Recalculating with Q gives:

$X_c = -272.564$	$C_c = 81.67\,pF$	*(close!)*
$X_d = -37.0998$	$C_d = 599.99\,pF$	
$X_e = 268.768$	$L_e = 5.983\,\mu H$	

C_c is within 0.4 per cent of desired value so it should work well.

Usual tolerances of fixed components are only 5 per cent. This will have little effect on matching when the other two components are trimmable. Lower design Q will show less sensitivity to tolerances.

applications and variations

Some situations cannot be met with given end resistances. This can sometimes be cured by using

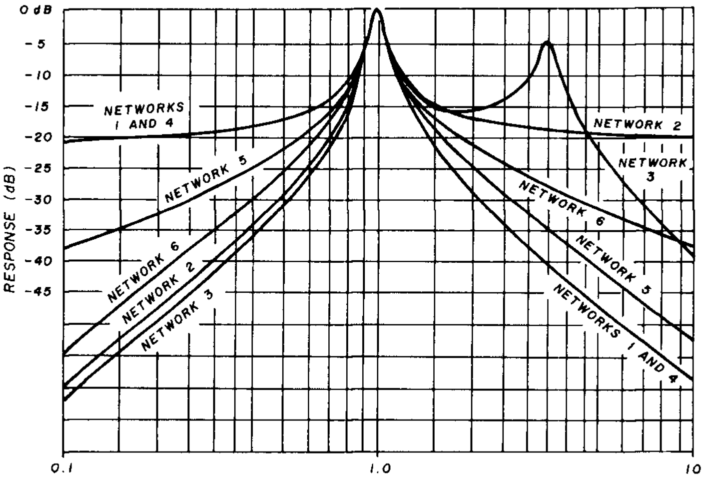


fig. 6. Frequency response of the six networks with the band center normalized to 7.15 MHz.

a broadband toroidal transformer.[9] Impedance changes of up to 16:1 are possible. Remember that parallel reactance is also changed by the same ratio in this application.

Good bypassing in active-device applications cannot be over-emphasized. A poor bypass and decoupling not only cause trouble between stages but also become networks. Resonance of bypass capacitors with lead length is common at higher frequencies; one cure is to double up several lower-capacitance bypasses. Short lead lengths and a large ground plane should always be used.

An interesting application is the replacement of the preselector tuning used in 1960-era receivers such as the Heath SB-300 series. These all have one band per bandswitch position so the front end can be stagger-tuned over the desired portion of the band. Fixed capacitors replace the variable and Networks 5 or 6 can be used at the antenna input. The Heath design has enough room to add a couple of bandswitch wafers to allow selection of other matching networks for other antennas such as a longwire. The variable capacitor can be retained, insulated, and used as part of a Wien-bridge audio notch filter.

Antenna networks used in receivers should have a dc path to ground such as in Network 5. This avoids static build-up during electrical storms and potential arcing.

other networks

End impedance frequency sensitivity may require simpler networks. An excellent treatment of L-networks is found in the first reference along with proper use of the Smith chart. Tabulated values are available[4,5] and theoretical aspects can be found in texts.[6]

Application of transistor amplifiers and matching is well covered in reference 7. Access to a computer that speaks FORTRAN can use the program of reference 8 to calculate other networks and also determine amplifier stability.

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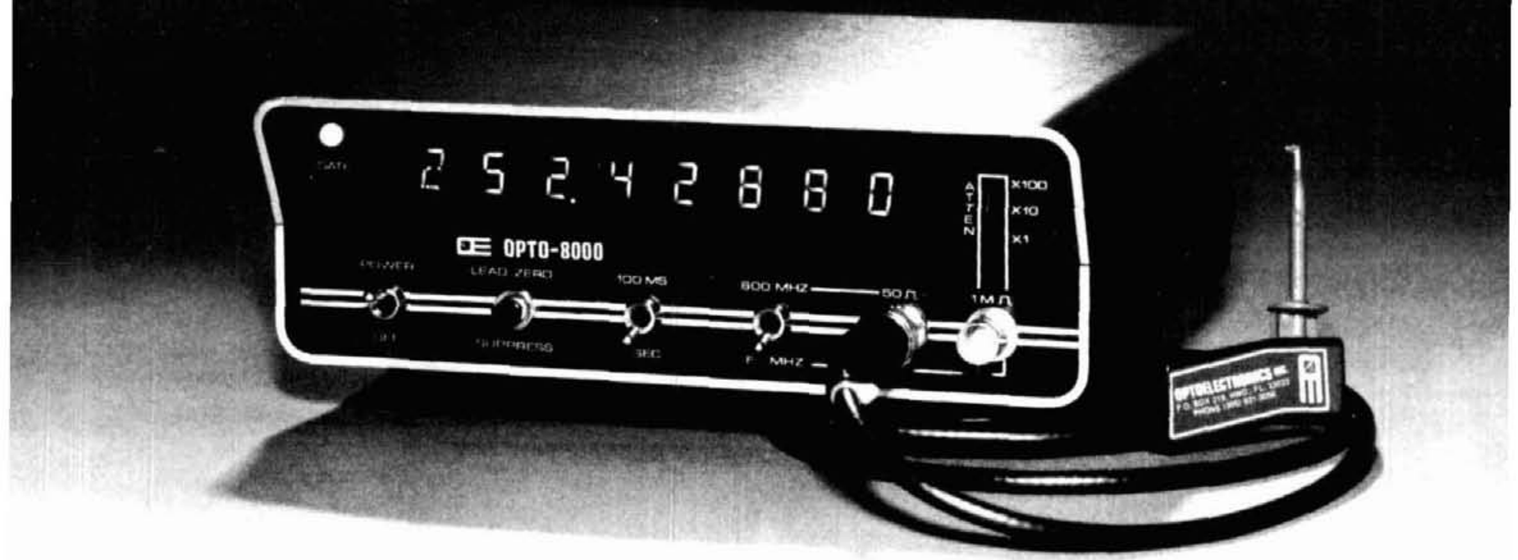

This new *instrument* has taken a *giant step* in front of the multitude of counters now available. The Opto-8000.1 boasts a combination of features and specifications not found in units costing several times its price. Accuracy of ±0.1 PPM or better — *Guaranteed* — with a factory-adjusted, sealed TCXO (Temperature Compensated Xtal Oscillator). **Even kits require no adjustment for guaranteed accuracy!** Built-in, selectable-step attenuator, rugged and attractive, black anodized aluminum case (.090″ thick aluminum) with tilt bail. 50 Ohm and 1 Megohm inputs, both with amplifier circuits for super sensitivity and both diode/overload protected. Front panel includes "Lead Zero Blanking Control" and a gate period indicator LED. AC and DC power cords with plugs included.

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Display—8 digit LED
Gate Times—1 second and 1/10 second
Selectable Input Attenuation—X1, X10, X100
Input Connectors Type —BNC
Approximate Size—3″h x 7½″w x 6½″d
Approximate Weight—2½ pounds
Cabinet—black anodized aluminum (.090″ thickness)
Input Power—9-15 VDC, 115 VAC 50/60 Hz or internal batteries

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overtone crystal oscillators without inductors

A discussion of overtone crystal oscillator circuits which don't require inductors

Until recently, all of the circuits for overtone crystal oscillators I've seen have included tuned LC circuits. It seemed necessary to have an LC resonator, tuned to the desired overtone frequency, to be sure the oscillator would operate at the proper overtone frequency and prevent operation at the crystal's fundamental frequency or some undesired overtone. It would be nice if the LC tuned circuit could be eliminated, of course, because it would simplify bandswitching of the crystal oscillator.

International Crystal has introduced a crystal oscillator circuit called the OF-1. Although the OF-1 circuit has no inductor, and thus no LC tuned circuit, it can be used with crystals operating in the third-overtone mode. I found this quite interesting and did some relevant experimenting to satisfy my curiosity. My efforts are documented here for others who may share this interest.

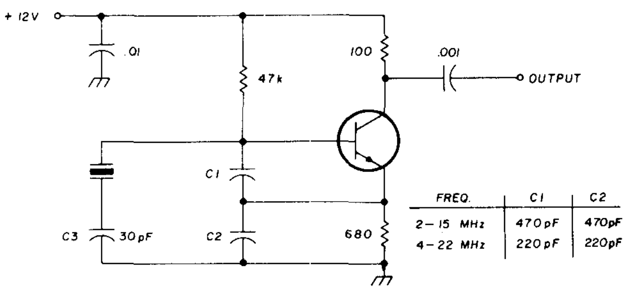

fig. 1. International Crystal OF-1 LO oscillator circuit for fundamental-mode crystals.

the circuit

International Crystal supplies two different kits of the OF-1 type. The OF-1 LO uses crystals operating in the fundamental mode from 2 to 22 MHz; **fig. 1** is a schematic of the circuit. The OF-1 HI uses crystals operating in the third-overtone mode from 18 to 60 MHz; **fig. 2** is a schematic of the OF-1 HI. Notice that in the latter circuit capacitor C3 has been omitted; in the overtone mode, the crystal operates near series resonance.

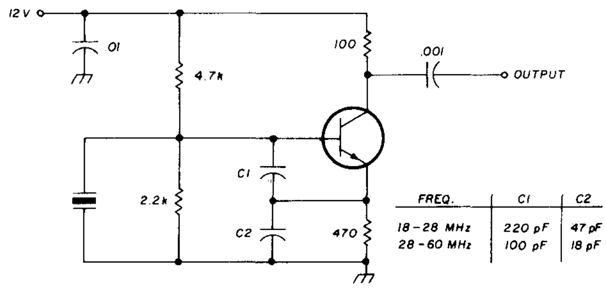

fig. 2. International Crystal OF-1 HI oscillator circuit for third-overtone crystals; note that no inductors are required.

I breadboarded and tested both of these circuits using a 2N4996 transistor. The crystal I used in all of my tests was a 28.3 MHz third-overtone type originally purchased for use in International Crystal's older OX oscillator circuit (which has an inductor). Using this crystal, the circuit of **fig. 1** had an output frequency at the crystal's fundamental, or about 9.43 MHz. The circuit in **fig. 2** produced the third-overtone frequency of 28.3 MHz when the smaller values shown for C1 and C2 were used; using the larger values for C1 and C2 given in **fig. 2** produced oscillation at the fundamental frequency of 9.43 MHz.

By Courtney Hall, WA5SNZ, 7716 La Verdura Drive, Dallas, Texas 75248

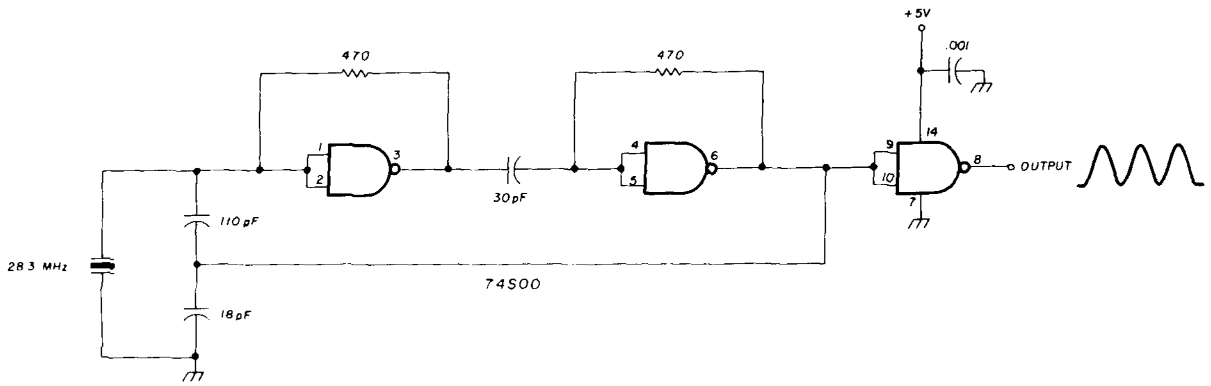

fig. 3. Third-overtone crystal oscillator which uses a 74S00 Schottky TTL gate; no inductors are required.

TTL version

To obtain increased output amplitude I developed the circuit shown in **fig. 3**. The 74S00 quad NAND gate acts as both oscillator and output buffer. One of the gates in the IC is unused. A 74S00 is required because of the relatively high frequency; I don't believe the 7400, 74LS00, or 74H00 will work as well, if at all, at 28.3 MHz.

All wiring should be as short as possible, and the circuit should be shielded. The peak-to-peak output amplitude swings from about 0.4 to 3.5 volts, which are acceptable TTL levels. If the capacitor which couples the first two gates together (30 pF in **fig. 3**) is too large (1000 pF), the circuit's output frequency will drop down to the crystal's fundamental. Some experimentation with the value of this capacitor may be required for different crystals.

conclusion

The simplicity of these overtone crystal oscillators could make them desirable in many applications. I have not had the opportunity to investigate them to the degree I would like; therefore, I would welcome comments from interested readers on this type of overtone oscillator circuit.

ham radio

simple method for making printed-circuit boards

I have developed a method of making printed-circuit boards that, while not professional in appearance, work well and beat the mess that results from using wire-wrap sockets and soldering them in place.

I use paint (enamel) in a K&E *Leroy* pen, inserted in a hole of a test probe. When using paint, do not use the insert for the pen, which is used to prevent the ink from running through. The paint is not thinned but just as it comes from the can; the result is a slow writing paint pen. I have found that a no. 2 or no. 3 *Leroy* pen is just about right. Cleaning is easy with pipe cleaners and paint thinner. I use an inking pen with paint for the long straight lines, and a small brush for the large areas.

To make a PC board, place the copper board under the layout, secure it so it won't slip, then use a sharp instrument such as a scribe and punch a prick hole through each place where there is a dot, or where there will be a hole. This gives the spacing you'll need to draw (free hand) the circuit on the board. After you have completed the hole punching (not through the board), remove the layout. Note that this does not destroy your layout. Now you can fill your pen and draw the circuit on the copper board.

I recommend you practice with simple circuits before you jump into a complex layout. After the paint has dried naturally or in an oven heated to 150 to 160 degrees, check to see that your layout is as you want it. If two lines are touching, the scribe will allow you to make a fine line between them.

I have tried two types of etching solutions: ferric chloride and an etch solution from Vector Electronics. The paint stood up well in both. First I tried two types of paint, Sears acrylic enamel, and Ace Hardware quick-drying black enamel. There was no apparent difference.

Drilling the board can be done before painting the circuit or after etching. The little prick marks show you exactly where the hole belongs. Removing the paint after etching is done with trichloroethane or other solvent that will not contaminate the board.

Robert H. Kernen, W4MTD

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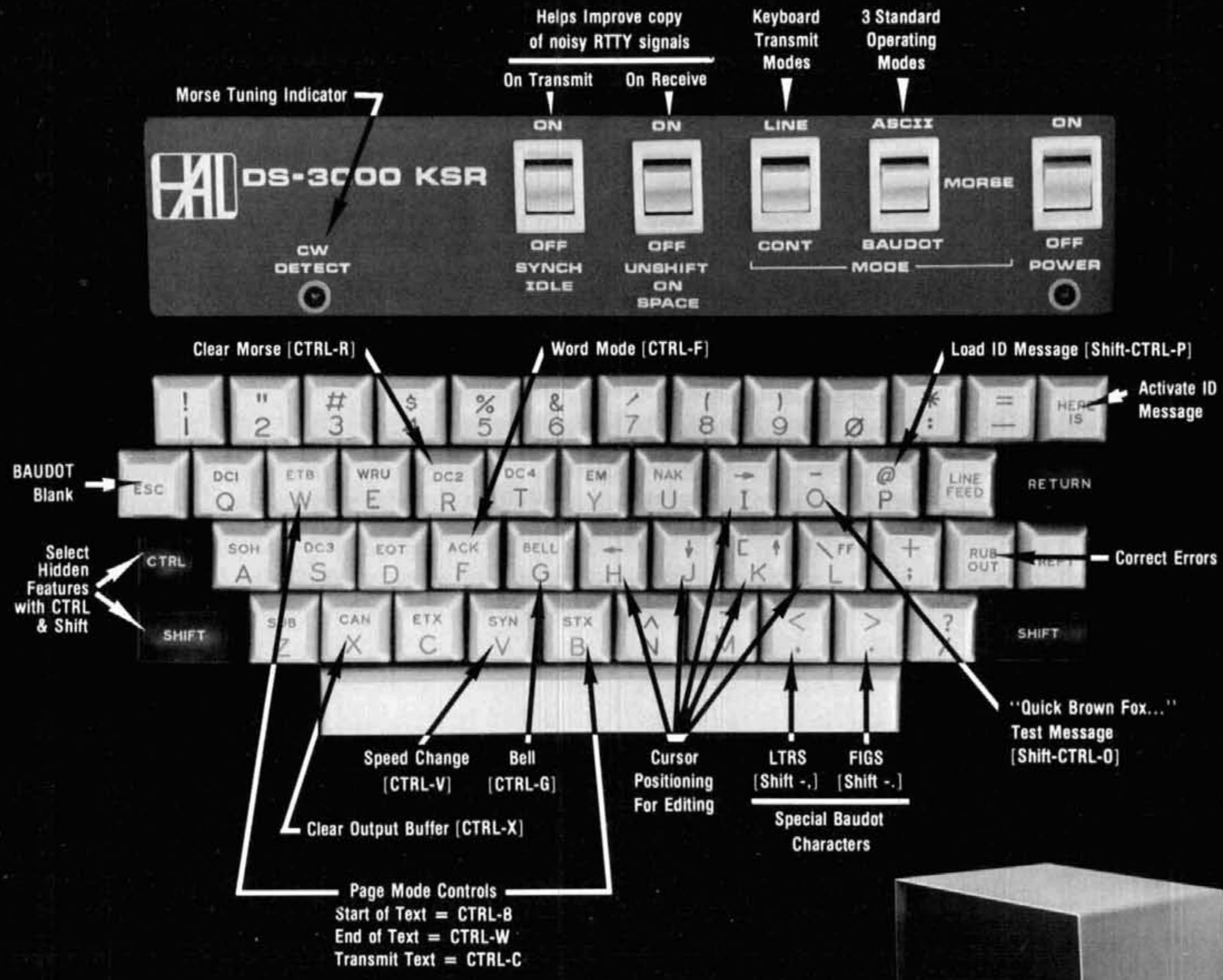

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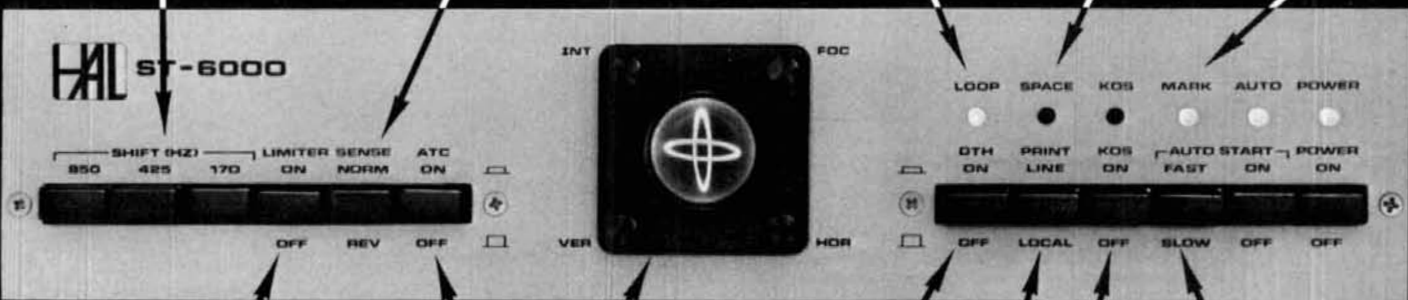

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repeater interference: some corrective actions

Suggestions for repeater associations to minimize the ever-increasing interference problem

Our local repeater, WR4ADC, operating on the 146.34/146.94 pair, has been experiencing interference on both the input and output frequencies. This article covers the technical corrective actions which have been considered for this interference. Some of these are now being used or are being installed for future use.

input channel anti-interference measures — lockout

Since repeaters operate on a fixed-frequency plan, one method of avoiding unnecessary call-up is to use a repeater lockout signal on transmissions not intended for repeaters. Technically, as in **fig. 1**, this would be simple. An NE-555 subaudible oscillator producing a small deviation would be required at the non-using transmitters. The repeater input receiver would need a tone detector and a relay contact in series with the carrier-operated relay (COR) to block operation when the tone is received.

While technically simple, this method of input interference prevention requires cooperation by the interfering stations. Lacking this, it is completely ineffective. It does seem to be a useful approach and could be adopted as a national standard. If done, the standard should include designation of the lockout tone frequency and of the deviation it produces.

While the lockout tone is not currently a useful technique, there are several technical anti-interference approaches available based on the characteristics of interfering signals. One of the most common of these is the frequency window, **fig. 2**, which locks out transmission if the incoming signal is appreciably off frequency. While this is usually done to avoid excessive distortion, it is a powerful antiinterference method. In its simplest form, it may be a Schmidt trigger on the dc voltage at the discriminator, set to the value corresponding to the selected value of allowable frequency error, with the trigger interrupting the COR. Some filtering, on the order of a tenth-second time constant, is needed.

signal characteristic measures

Other anti-interference measures based on the characteristics of interfering signals include:

1. a-m rejection
2. Wideband fm rejection
3. Non-voice modulation rejection

The purpose of a-m rejection, **fig. 3**, is to prevent operation of the repeater COR by an a-m carrier. In simple form this requires a pickoff from the receiver ahead of the limiters, feeding to an agc-controlled stage and an a-m detector. The lockout circuit would require presence of carrier in the fm section and in the a-m section, plus presence of a-m exceeding some percentage, say around 25 per cent. Some protection against noise pulses or unmodulated pauses would probably be required, at least several seconds time constant of filtering.

Rejection of wideband fm, **fig. 4**, would require a wideband receiver, and a form of detection of the out-of-band energy component. A simple form would be to compare audio levels of the narrow and wideband discriminator outputs: if they approach equality, the incoming signal would be wideband. Another wideband modulation detector would beat the limiter output against a local oscillator, followed by a highpass filter. The filter cutoff frequency would be set to the desired peak deviation.

A lockout system based on the absence of voice modulation, **fig. 5**, would be a powerful anti-interference technique. In principle, this is relatively simple — just measure the ratio of peak-to-average power in the audio. The ratio varies from 3 dB for pure sine waves, to 13 dB for noise, and to about 16 dB for voice. Noise-operated squelch makes the distinction between the last two easy.

However, in current fm practice, the approach is not this simple, since most fm transmitters use a

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limiter to keep the deviation high, and this reduces the peak-to-average ratio. Because of this, a peak-to-average detector would have to be set to a lower ratio. About 6 dB, or 2:1 in voltage, with a time constant of about one-tenth second should be good and would provide appreciable protection against non-voice signals.

wanted signal anti-interference measures

All of the above anti-interference measures are based on some characteristic of the unwanted signal. The other family of input frequency measures is based on characteristics of wanted signal. The two standard forms of this, **fig. 6**, are the subaudible tone, or *private line*, required continuously for access; and whistle-on, a single tone or a Touch-Tone signal giving access. Access may be for a definite time period or until the COR is dropped for a time period. Various other forms have been used, including carrier-formed Morse and two-band interrogation.

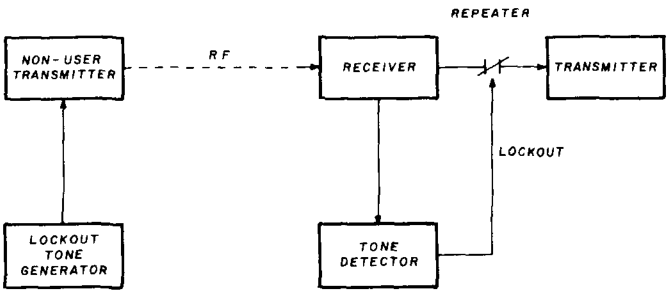


fig. 1. Elements of the repeater lockout system.

Most of these methods have the disadvantage of requiring additions to the using transmitter. This is not a severe problem for closed repeaters, where the number of users is small, but represents a serious drawback for open repeaters, especially those on the common frequencies or close to holiday areas or heavily traveled routes.

One method of partially overcoming this objection is to operate the repeater with carrier access only during no interference periods, and to switch to one of the other access methods when interference is experienced. This switch could be coupled with voice announcement on the identifier, giving instructions as to access method. A Touch-Tone access is probably best since such pads are common and since the signal should never be used in normal communication, except for signaling.

It would seem that this method would give good protection against incidental interference. Willful interference is another matter, since operation in ac-

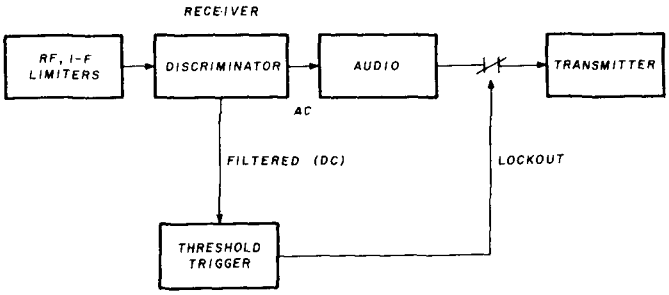


fig. 2. Elements of the frequency window, which locks out transmission if the incoming signal is appreciably off frequency.

cordance with the announcement instructions would turn the repeater on. However, malicious intent would seem to be proven if interference continues.

output channel anti-interference measures — channel guards

Repeaters also cause interference, and some technique of guarding against the interference they cause may be desirable. Probably the best method of doing this is to guard the output channel, **fig. 7**, holding repeater operation in abeyance if the channel signal exceeds some predetermined value. To account for emergency operation, this guard could be combined with a timer identification announcement, giving instructions on the procedure to override the lockout.

The setting of the guard receiver would need to be based on a value of "signal to be protected," and a margin, or "protection ratio." These are common concepts in other radio services, but they have not been used in amateur operations. As a result, there are no accepted values for these quantities.

The closest other service is the Mobile Service, which includes land mobile. Since this was the source of many 2-meter repeater concepts and equipment, it should provide good guidance.

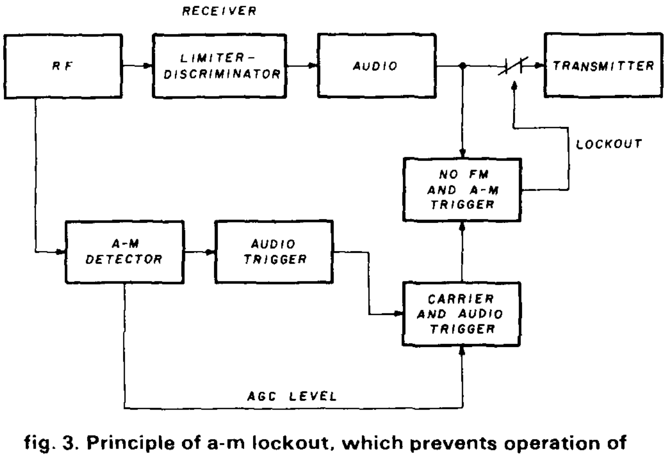


fig. 3. Principle of a-m lockout, which prevents operation of the repeater COR by an a-m carrier.

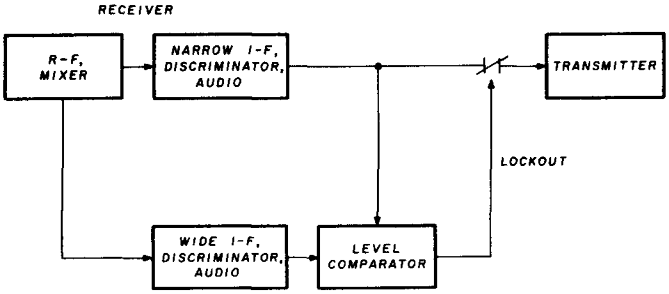


fig. 4. Principle of wideband fm lockout, which would require a wideband receiver and a form of detecting the out-of-band energy component.

The International Consultative Committee (CCIR) data which relates to this matter is in their Report 358. This report states that, not considering the effects of natural or man-made noise, the median signal levels to be protected for an average grade of mobile service are:

band (MHz)	signal (dB above 1 μV/m)
30-50	8
50-100	14
100-200	20
400-470	28

The report also states that the protection ratio, based on degradation from 20 dB S/N to 14 dB S/I, should be 8 dB for fm with fm interference, or 7 dB for fm with a-m interference. The report lists other ratios, but it has not been necessary to consider these, since the 2-meter National plan is based on narrow-band fm.

The 1971 Special Joint Meeting of the CCIR gives the following noise values:

typical urban noise, 150 MHz, +35 dB (KTB)
typical rural noise, 150 MHz, +3 dB (KTB)
maximum cosmic noise, 150 MHz, +8 dB (KTB)

These figures are based on omnidirectional antennas.

Many repeaters are located in urban areas and will see full urban noise levels; others are remotely located and will see less noise. Assume that the cosmic value, 8 dB, represents typical noise. Then the value of signal to be protected becomes 20+8 or 28 dB above 1 μV/m — about 25 μV/m. (This signal would give 20 dB output signal/noise in the average mobile installation and in a typical location.) The allowable value of interference that would give protection for this signal is 28-8, or 20 dB above 1 μV/m — just 10 μV/m. (Presence of the interference would reduce the 20 dB S/N to 14 dB S/I.)

It is easy to see the effect of this protection level if the values are translated to power and distance relationships. For average terrain, the effective radiated transmitter power to just reach the 10 μV/m signal, as a function of distance from the repeater, is approximately:

distance km	distance (miles)	transmitter antenna height: 100 ft (30m) or more transmitter power (W)	30 ft (9m) transmitter power (W)	10 ft (3m) or less transmitter power (W)
10	6.2	0.02	0.3	3.0
15	9.3	0.5	3.0	31.0
20	12.4	1.0	6.0	60.0
30	18.6	8.0	50.0	500.0
50	31.0	100.0	500.0	5000.0
100	62.1	3000.0	8000.0	—

The ERP is the product of the transmitter output and the antenna gain multiplied by 2.5 (to include the effect of ground reflection).

Stated another way, the 10-microvolt signal would be produced by a typical mobile at about 8 miles (13km) distance, by a typical base station at about 18 miles (29km), or by a DX station about 90 miles (144km) away and beaming toward the repeater. The signal would be the same as that of another typical repeater if it were some 50 miles (80km) away.

These values indicate that output channel guarding to the 10 μV level would not eliminate *all* interference. However, it would prevent interference to simplex operation within reasonable distances, to the point that any resulting interference would hardly be considered harmful. Accordingly it would appear that repeater associations wishing to minimize interference problems should consider this approach.

Construction of an automatic protection system should not be difficult. The output-channel monitor-receiver antenna could be mounted on the repeater tower at a height of 30 feet (9m). (This level is used because the height gain is zero.) Assuming use of a

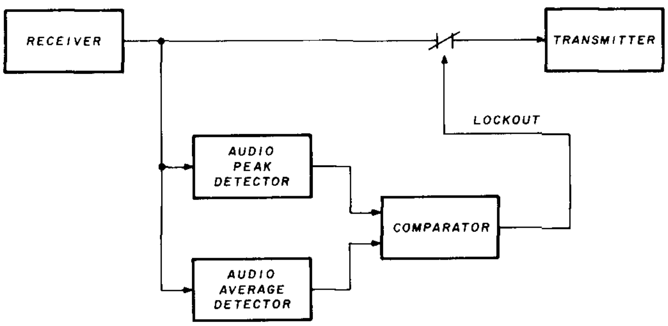


fig. 5. The non-voice lockout principle. Comparator measures peak-to-average power ratio in the audio, which varies from 3 dB for pure sine waves to about 16 dB for voice.

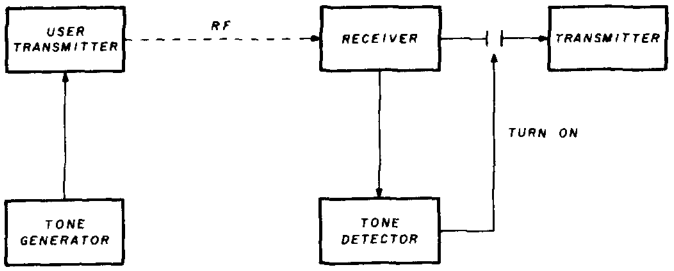


fig. 6. Elements of the tone-access system, which is based on characteristics of the wanted, rather than the interfering, signal.

dipole, its captive area is 0.54 m^2 (free space), or 1.35 m^2 over average ground. A signal of 10 $\mu V/m + 10$ dB above 1V corresponds to a flux level of -136 dB W/m^2, so the antenna captures about -133 dB relative to 1 watt. Assuming 3 dB line loss, this becomes -136 dBW at the receiver input, or about 0.2 microvolt.

Setting the squelch to operate at this signal level would allow use of the squelch circuit to produce the protect control signal. Suitable timers and desensitize circuits would have to be included. A five-minute delay after detection of the channel-occupied signal might be used. A voice announcement could be included to the effect that the output channel is occupied. Emergency over-ride instructions could also be included.

auto-QSY

Probably the most important HF method of interference control is to QSY. This is not really practical in repeater operation. However, there is a partial step which seems possible, **fig. 8**, and one which would make QSY for the station on the repeater output channel easier. This is to shift the repeater output frequency by a small amount; 5 kHz would seem to be suitable.

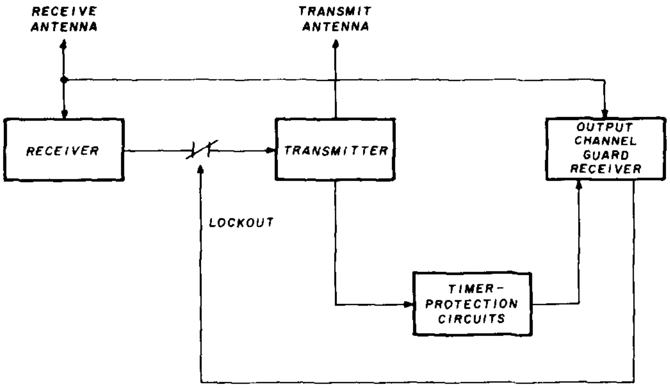


fig. 7. The output-frequency guard receiver principle, in which repeater operation is locked out if the channel signal exceeds some predetermined value.

A limited number of tests with this method indicate that the effect on normal repeater operation would be negligible — a small increase in distortion and a small decrease in range. With no other action, the effect on the other station on the channel would also be small. However, if the other station would also QSY by the same amount, but in the opposite direction, he would escape the interference completely. At the same time his operation would be affected only by a small amount. For the repeater users and the other stations involved, nearly full operation could be restored by retuning.

The signal for this scheme can be generated by the circuitry used for transmit lockout. (If desired, the channel-occupied receiver could be widened out a bit to give better detection of these slightly offset signals.)

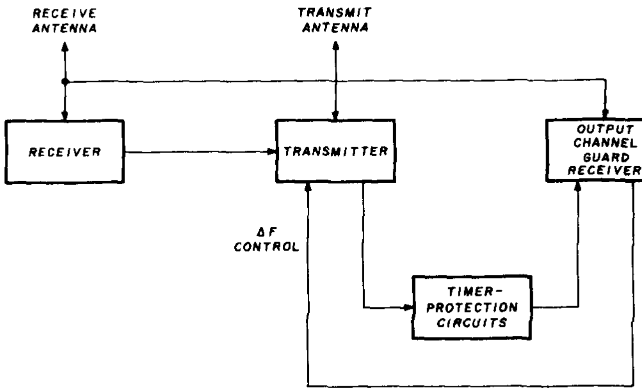


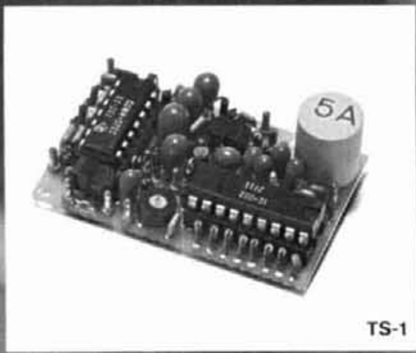
fig. 8. Principle of channel-occupied QSY method, which would make it worthwhile for the station on the repeater output to change frequency slightly.

It may be noted that this offset method is used in vhf television to reduce co-channel interference and to allow closer station spacings. It is a powerful anti-interference technique.

A number of other anti-interference measures have been considered with respect to WR4ADC interference problems. The other technical ones considered did not seem attractive and so are not reported. There were several operational ones of value, and these have been discussed along with the technical solutions above.

With increasing numbers of amateurs on the air, and especially with the continued expansion of repeater operation, interference is going to increase. It seems to be time for the repeater associations to consider the steps that should be taken to minimize those interference problems. This should include recommendation of standard techniques and of protection ratios.

ham radio



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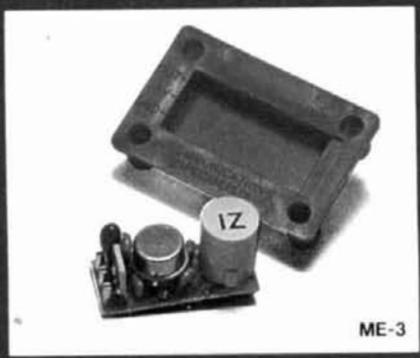

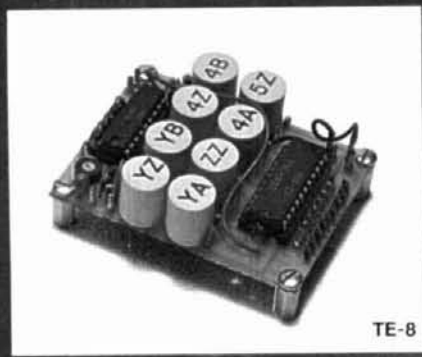

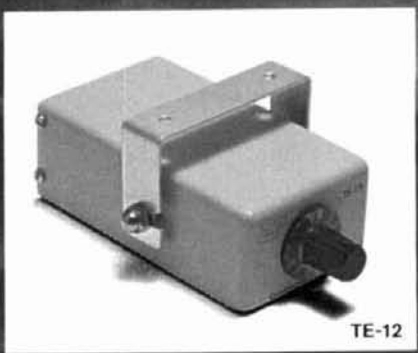

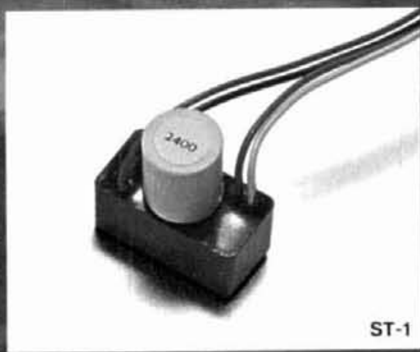

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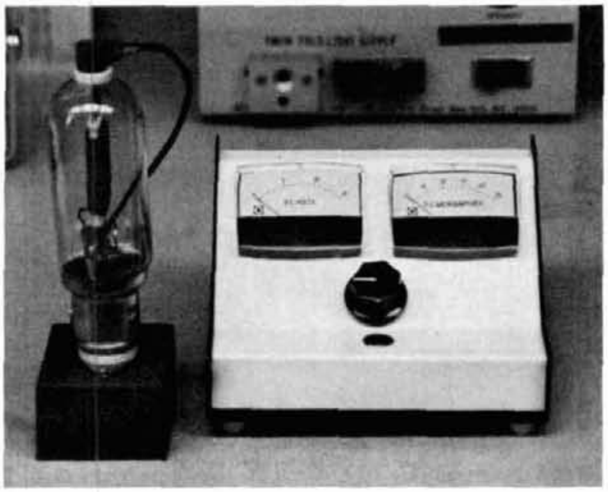
testing power tubes

Several months ago a reader of this column wrote me asking the question, "How do you test rf power tubes?" Not long afterward a close friend of mine, about to build a linear amplifier for 80 through 10 meters, bought a box full of used 4X150 power tetrodes at a hamfest. After building the amplifier he found that none of the tubes were good and that he would have been better off using another type of tube from his collection that was known to be good. He naturally wanted to know if there was a better way available to amateurs than testing by trying.

In my work, medical electronic servicing, we repair electrosurgical generators, which are rf power oscillators that produce up to several hundred watts in the 500-2500-kHz range. Some old but still perfectly good electrosurgical generators use the type UXCV-11 power triodes in a push-pull pair. These tubes cost about $120 each ($240 per pair!) unless you are clever enough to know about United Electronics in Newark, New Jersey. The cost makes stocking a number large enough to permit routine testing by substitution too costly — especially in a shop with a limited budget. Again, there "has to be a better way!"



Final project built by the author for testing high-power vacuum tubes.

Three times within the past six months people and situations have forced the issue of testing transmitting tubes, so perhaps it is time that we covered that point in this column. But before we answer the question posed originally, let's recap a little about elementary vacuum tubes.

A vacuum tube consists of an electron emitter called a cathode, several grids, and an electron collector called either a plate or an anode. The cathode can be either an incandescent filament, or an indirectly heated metal cylinder that has a heated filament at its center. The plate surrounds the cathode but is insulated from it. The various grids are placed in the space between the cathode and the plate.

The plate will have an electrical potential that is positive with respect to the cathode. Grid no. 1, the *control grid,* is given a negative potential with respect to the cathode so it can control the flow of electrons between cathode and plate. The second grid is called the screen grid or accelerator grid and is given a positive potential with respect to the cathode. The third grid, that nearest the plate (G3), is called the *suppressor grid* and may be either biased negative with respect to the cathode or (more commonly) tied directly to the cathode either internally or externally, as shown in **fig. 1**.

Variations in the grid voltage will produce variations in the plate current, which by Ohm's law, become variations in the voltage drop across the load resistor, R_L.

Several vacuum tube parameters are of interest when trying to ascertain overall quality. These are: amplification factor (mu or μ), plate resistance (R_p from the spec sheet or data book), and the transconductance (gm).

The amplification factor is defined as the change in output voltage caused by a given change in grid no. 1 (input) voltage. In fact, the amplification factor of any amplifying device can be given by:

$$amplification = \frac{E_{out}}{E_{in}} \quad (1)$$

But for vacuum tubes specifically we can use the notation:

$$\mu = \frac{\Delta E_b}{\Delta E_c} \quad (2)$$

where:

ΔE_b is the change in plate voltage
ΔE_c is the change in plate current

The transconductance rating relates a change in plate *current* (I_p) for a *small change* in grid voltage, with the plate *voltage* held constant. In other words:

$$gm = \frac{\Delta I_p}{\Delta E_c} \Bigg|_{E_b = constant} \qquad (3)$$

where:

ΔI_p is the change in plate current expressed in amperes

ΔE_e is the small change in grid bias voltage expressed in volts

gm is the transconductance expressed in mhos

This relationship will lead us to a method for testing tubes easily and quickly.

tube tester configurations

Fig. 2 *shows the basic circuit for a short-circuit* tester. This will give only limited information but is useful when screening a large number of hamfest specials. Why perform a more time-consuming test on a tube that has a high resistance short between the filament and cathode, for example?

The tester circuit is nothing more than a series of several continuity testers arranged to ascertain the existence of any resistance paths between adjacent elements. Each continuity tester consists of a low-current filament transformer (*i.e.*, rated at less than 1 ampere) and a compatible lamp. I was tempted to specify the use of light emitting diodes instead of lamps but was quickly persuaded to use lamps because the LEDs made the device too sensitive. This problem causes apparent shorts, when all we are reading is electrons flowing from the heated filament to the electrode on half cycles when the ac is positive-going.

It is best to test for shorts after allowing the tube to warm up for a few minutes. Some shorts do not become apparent until after the tube has reached operating temperature. This is also the reason why the use of an ohmmeter is not recommended in this case.

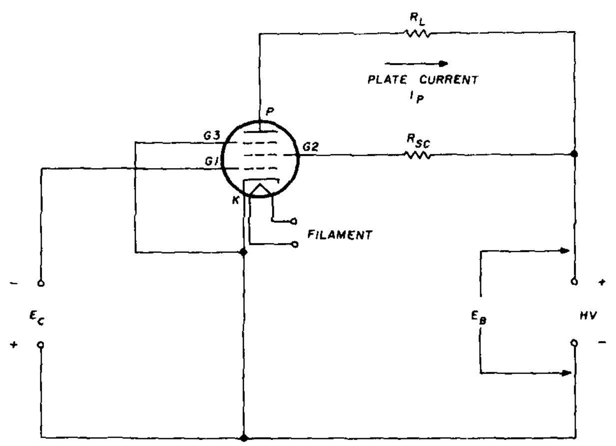


fig. 1. Fundamental vacuum-tube circuit.

The simplest tester circuit that gives us a qualitative insight into the worth of any given tube is the *emission tester* of **fig. 3**. This circuit tests the tube for the emission of electrons from the cathode. Notice that the tube is connected in a diode configuration in which all elements except the cathode and filaments are connected to the plate.

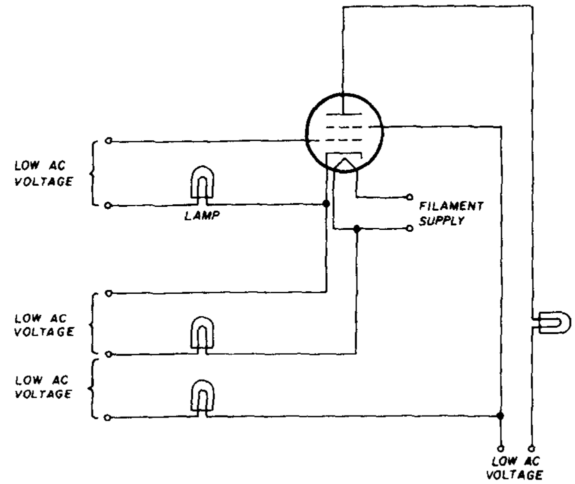


fig. 2. Short-circuit tester.

The tube in a circuit such as **fig. 3** acts very much like the classic diode. The *emission current* is defined as the saturation current of the tube. If the plate voltage is increased from near zero, we find that the plate current will also increase in a nearly linear manner (except at very low plate potentials). But once a certain critical plate voltage is reached, we find that the plate will attract all the electrons that the cathode produces. Any further increase in plate voltage will produce very little, if any, increase in plate current. The current level at which this occurs is the *saturation* or *emission* current.*

The emission-type tube tester checks for this current, and if the current is low, it will indicate "reject." Some tubes cannot be tested at the actual saturation current either because it is inconvenient to do so, or because the current is too high and may damage the tube. These tubes are tested at a specific plate potential, at which a given current is expected. If the tube will not produce at least a certain predetermined percentage of that current (usually 80 to 90 per cent), then it is rejected. The emission tester can spot a grossly bad tube, but there are certain problems with

*An interesting variation of the emission tester, which does not require high voltage but uses a low-voltage bias source, is described in reference 1. editor.

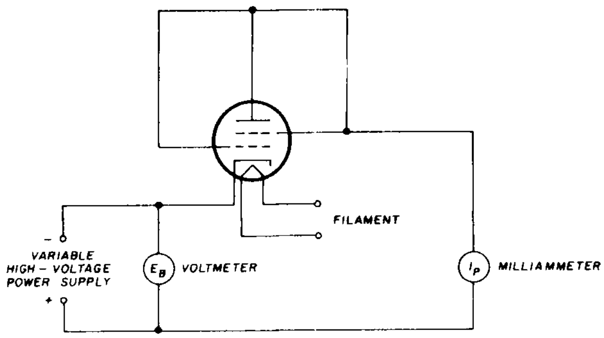


fig. 3. Circuit for simple emission-type tube tester.

this type. Note that most drug-store tube testers are emission types.

Defects that will prevent the tube from operating normally will not always show up on a simple emission tester, so some better means is necessary. The type of tester preferred by most professional servicers is the *mutual conductance* or *transconductance* tester. Examples of this type of circuit are given in **fig. 4**. The circuit shown in **fig. 4A** is a static transconductance tester using the *grid-shift* method, also called grid-level shift.

Switch S1 iis in position 1 at the beginning of the test, making grid voltage E_c equal to $E2$ alone. Both $E1$ and $E2$ are adjusted to produce a convenient (safe!) plate current. It is usually wiser to begin with the plate voltage ($E1$) at some level well within the range that can be tolerated by the particular tube being tested, and have the grid voltage at some value in excess of cutoff for that tube. This may be unnecessary much of the time but could save you some grief often enough to make it a good standard practice. You may then adjust the grid voltage downward until the plate current is at a convenient level. When this adjustment is completed, make a note of the values of E_b, I_p, and E_c.

To perform the test, place switch S1 in position 2. This operation makes grid voltage E_c equal to $(E2 + E3)$. The plate voltage is then measured and the supply readjusted if a change has been noted. The plate voltage at this point must be equal to the plate voltage that existed initially.

Now read the plate current, and find the *difference* between this reading and the initial reading. Plug the difference current into the formula:

$$gm = \frac{\Delta I_p}{1.5} \quad (4)$$

where:

ΔI_p is in amperes

gm is in mhos

This answer is given in *mhos*, but most vacuum tube spec sheets list the transconductance in micromhos. 1 mho = 1-million μmhos, so multiply the answer by 10^6.

A *dynamic* transconductance tube tester is shown in **fig. 4B**. This circuit is the basis for most commercial tube testers. A low voltage ac transformer is connected in series with the grid bias power supply, so that the transformer secondary voltage forms the "*delta-E_c*." Actual transconductance is measured on the plate ac milliammeter, which is calibrated in units of conductance.

Fig. 5 and the photo show a test jig I built to test power tubes at work. It will serve equally well for amateurs. The circuit is the simple grid-shift method for finding the transconductance of the tube being tested. Note that it is not a real tube tester construction project because it lacks power supplies. Almost all amateurs can jury-rig adequate power supplies to make this test or can borrow bench supplies. Note also that it is not strictly necessary to use the full operating voltage of the tube to obtain meaningful results.

Please, be very careful when using this jig; high voltages will be exposed! If there is any doubt, place the tube socket subassembly inside an insulated or *grounded* metal enclosure. I may seem to harp on

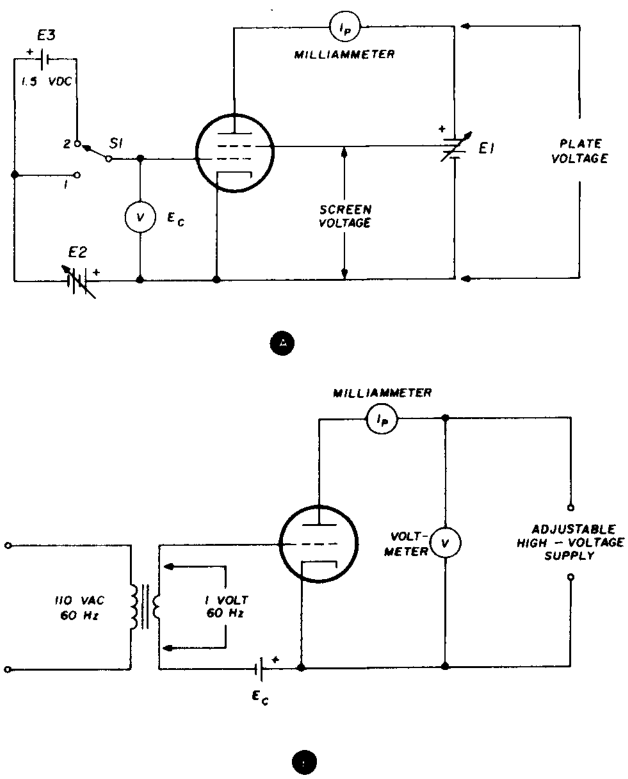


fig. 4. Static transconductance tester, A, and dynamic transconductance tester, B.

safety an awful lot, but it is so very important. Recall the note on one of my past articles which pointed out that an editor almost canned himself while working on a 4000-volt final amplifier power supply.

The test jig is built in two main parts, a main assembly and a tube-socket subassembly. A multiconductor cable connects the two parts of the jig. This design was selected so that the same mainframe can be used to accommodate a larger number of different tube types.

Jacks J1-J8 are heavy duty banana-jack binding posts, while J9/P1 are a mating pair of multipin connectors such as the circular MS or AN series. I used a high-voltage power supply that delivered 500 Vdc, but if greater potentials are needed (I doubt that they will) it would be wise to change J1-J4 to high-voltage chassis connectors.

Meter M1 is an appropriate high-voltage meter, although in my case a voltmeter was made from a suitable multiplier resistor (a pot and a fixed resistor) and an available 0-50-μA meter movement. Meter M2 is a dc voltmeter with a range suitable for the range of grid voltages expected.

Resistors R1-R3 were selected so that varying R2 produced approximately 1 volt of grid voltage change. R1 is used to allow precision trimming of that change. In my case, E_c' was 25 Vdc, but if your voltage is different, use the normal voltage divider equation to find appropriate resistor values.

Resistor R4 is made approximately equal to the plate resistance of the tube being tested, which in this case was about 3000 ohms.

I used an external plate current milliammeter

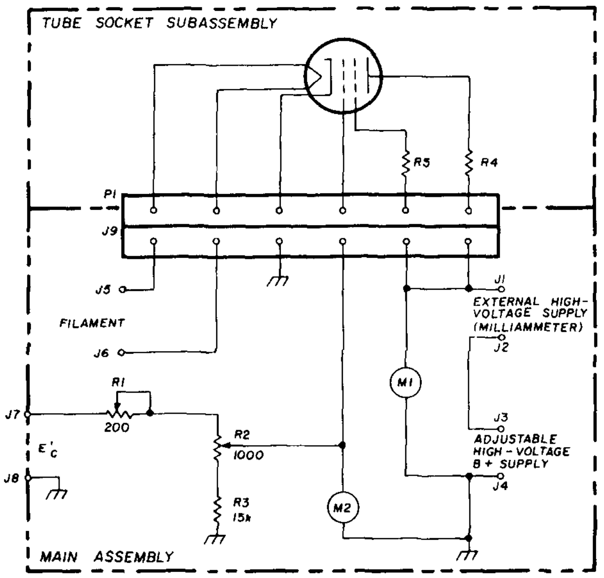


fig. 5. Schematic for a practical static (grid-shift) transconductance tube tester.

table 1. Transconductance measurements of type UXCV-11 tubes made by the author on the test jig shown in fig. 5.

bad tubes	ΔE_c (volts)	ΔI_p (amps)	gm (micromhos)
1	1	0.0013	1300
2	1	0.0015	1500
3	1	0.0014	1400
good tubes			
1	1	0.0045	4500
2	1	0.0039	3900
3	1	0.0043	4300

because there was a digital multimeter available that would measure current very accurately.

Table 1 lists the values of transconductance actually measured for both known good and known bad tubes. The UXCV-11 has a *mu* of 14 and a plate resistance of 3220 ohms. The spec sheet does not give the transconductance, but we may compute it from:

$$gm = \frac{\mu}{R_p} \quad (5)$$

$$gm = \frac{14}{3220}$$

$$gm \approx 0.00435 \; mhos$$

$$gm \approx 4350 \; \mu mhos$$

(Note that all the bad tubes had grossly lowered gm readings.)

This tester is not "scientifically" designed but is intended to allow amateurs to test power tubes on an occasional basis. Its saving grace is that it can be built inexpensively! Note that I defined "bad tube" by saving from my own work those tubes that had been found to produce customer complaints similar to, "It seems to work, but has low output." These observations were confirmed by an rf ammeter in series with my 500-ohm (not 50-ohm; these were medical rf generators) dummy load. The real value of this tester is that it will allow you to test hamfest specials or perform preventive maintenance on your equipment. Of course, if you have an rf power meter or rf ammeter in your feedline, then a low-power output coupled with seemingly normal drive will point the finger to the final amplifier tubes. But the thought of that poor guy building an entire linear amplifier around a whole box full of bad tubes seems to justify doing a little testing — at five minutes per tube, how could he have gone wrong?

references

1. Neil Johnson, W2OLU, "Testing High-Power Tubes," *ham radio* (*The Ham Notebook*), March, 1972, page 64.

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microcomputer interfacing: interfacing a 10-bit DAC

A Digital-to-Analog Converter or DAC is an electronic device that converts digital signals into analog signals. A typical converter consists of an arrangement of "weighted" resistors, each controlled by a single bit of input data, that develops varying output analog voltages or currents in accordance with the digital input code.[1] You could use a DAC to provide a small analog error signal from a microcomputer used in a feedback circuit, to convert a sequence of bytes in memory into analog-*vs*-time data and thus simulate the output from an analog instrument such as a rotator control box for tracking OSCAR; to provide analog data for the two channels of an x-y recorder, or in general, to operate any device that requires an analog voltage or current and is interfaced to a digital device, such as a microcomputer.

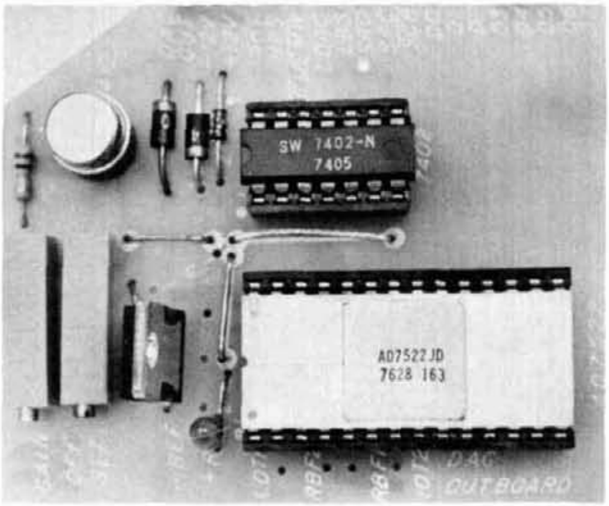
For a general discussion of the principles of analog/digital conversion, you should read the excellent Analog Devices conversion handbook[1] or the series of small pamphlets distributed by National Semiconductor Corporation.[2] Important terms and concepts associated with DACs include resolution, accuracy, scale error, gain error, offset error, linearity, differential linearity, settling time, slew rate, overshoot and glitches, temperature coefficient, supply rejection, conversion rate, and output drive capability. A few of the terms have been summarized in **table 1**.

To help understand how you'd interface a DAC to an 8-bit microcomputer, **fig. 1** shows the connections between an Analog Device AD7522 and an 8080A-based microcomputer. An important feature of this specific DAC is the fact that it is double buffered; this means that there exist within the device

By Peter R. Rony, Jonathan A. Titus, Christopher Titus, and David G. Larsen, WB4HYJ

Mr. Larsen, Department of Chemistry, and Dr. Rony, Department of Chemical Engineering are with the Virginia Polytechnic Institute and State University, Blacksburg, Virginia. Mr. Jonathan Titus, and Dr. Christopher Titus are with Tychon, Inc., Blacksburg, Virginia.

The DAC *Outboard* circuit board contains all the circuitry shown in *fig. 1*. A copy of the board layout is available from the authors.



two independent 10-bit registers, the DAC register and the two-bit and eight-bit shift registers (**fig. 2**).

A DAC is an output device for a microcomputer, and thus data is strobed from the microcomputer data bus into the internal registers or latches, of the DAC. In **fig. 1**, are shown the connections to the 8-bit bidirectional data bus, D0 through D7, the 8080A control signals OUT or MEMW, which are used with accumulator I/O or memory I/O data transfers,[3] and the channel select outputs 003 through 005 that are generated by a decoder tied to the microcomputer address bus.[4]

Since the AD7522 is a 10-bit DAC, it is not possible to simultaneously load all ten bits from an 8-bit microcomputer. The sequence that actually occurs can be summarized as:

1. The DAC input bits DB0 through DB7 are first strobed into the 8-bit shift register/latch using a positive device select pulse applied at pin 24, (LBS or Low Byte Strobe).

2. The most significant two bits, DB8 and DB9, are then strobed into the 2-bit shift register via the use of a device select pulse applied at pin 25 (HBS or High Byte Strobe).

3. Finally, a device select pulse applied at pin 22 (LDAC or Load DAC) transfers the ten bits of input data, DB0 through DB9, into the second buffer within the DAC chip, the DAC register.

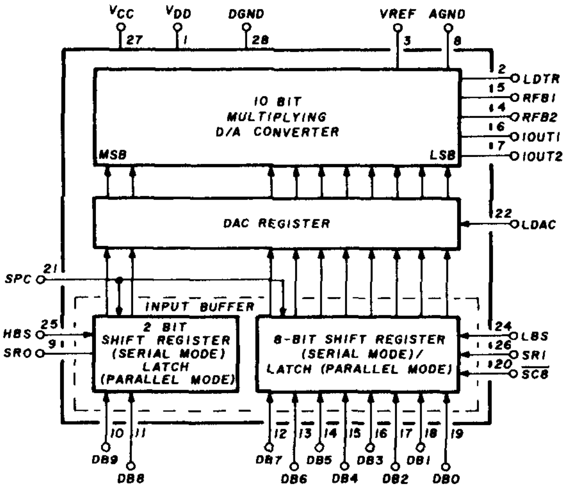


fig. 2. Functional diagram of the AD7522 IC. This figure is courtesy of Analog Devices.

The output current appears at IOUT1 and IOUT2 and is converted into a voltage with the aid of a 741 operational amplifier. The two most significant bits are loaded from the eight-bit microcomputer bus using any two bits. Generally bits D0 and D1 are chosen since it makes data formatting easier. Thus, the ten bits are transferred as eight bits D0 to D7 and as two additional bits, D0 and D1.

A simple program that exercises the DAC over its full operating range is provided in **table 2**. The program generates a slow linear ramp at the analog out-

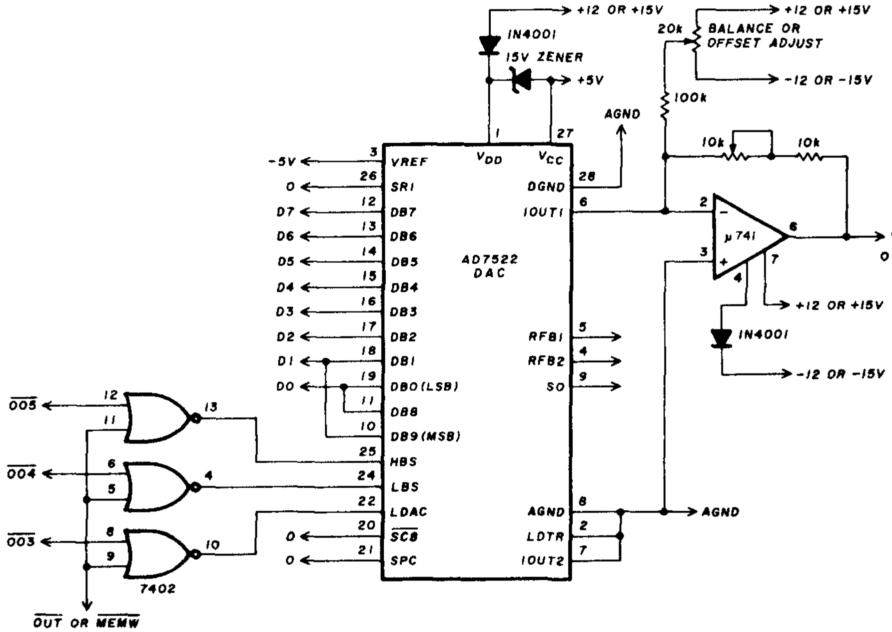


fig. 1. Schematic diagram of an interface circuit between an 8080A-based microcomputer and an Analog Devices AD7522 digital-to-analog converter.

table 1. Important concepts and terms associated with digital-to-analog converters.

Resolution	The smallest standard incremental change in output voltage of a DAC. A converter with *n* input bits can resolve one part in 2^n.
Accuracy	Describes the worst case deviation of the DAC output voltage from a straight line drawn between zero and full scale; it includes all errors.
Settling time	The elapsed time after a code transition for a DAC output to reach a final value within specified limits.
Conversion rate	The speed at which a DAC can make repetitive data conversions.
Nonlinearity	Error contributed by a deviation of the DAC transfer function from a best straight line function. Normally expressed as a percentage of full scale range.
Monolithic chip	An integrated circuit chip in which both active and passive elements are simultaneously formed in a single small silicon wafer via the use of diffusion and epitaxial processes. Metallic stripes are evaporated onto the oxidized surface of the silicon to interconnect the elements.
Multiplying DAC	A digital-to-analog converter in which the output analog signal is the product of the number represented by the digital input code and the input analog reference voltage, which may vary from scale to zero, and in some cases, even to negative values.

table 2. Memory I/O program that generates a slow linear ramp. Execution starts at HI = 003 and LO = 000.

LO address byte	instruction byte	mnemonic	comments
START: 000	042	SHLD	Strobe ten bits of digital data into the AD7522 DAC shift registers. The ten input data bits are contained in register pair H. The address select code for the LBS input is HI = 200 and LO = 004; the address select code for the HBS input is HI = 100 and LO = 005.
001	004	004	
002	200	200	
003	062	STA	Strobe ten bits of digital data from the input buffer into the DAC register within the AD7522 DAC. The address select code for the LDAC input is HI = 200 and LO = 003.
004	003	003	
005	200	200	
006	043	INX H	Increment register pair H
007	315	CALL*	Call 10 ms time delay routine, DELAY
010	277	277	LO address byte of DELAY
011	000	000	HI address byte of DELAY
012	303	JMP	Unconditional jump to START, where the input of new data into the DAC occurs
013	000	000	LO address byte of START
014	003	003	HI address byte of START

*On the 8080-based microcomputer that we use in our courses, a 10 millisecond time delay subroutine is located in EPROM starting at HI = 000 and LO = 277. Such a routine can be located anywhere in memory.

put of the AD7522. This can be observed on a Vom, digital multimeter, or oscilloscope. The ramp is subdivided into 1024 small steps, each step being approximately 5 mV in magnitude. The total time required to change from 0.0 volts to +5.12 volts is 10.24 seconds. The SHLD <B2> <B3> instruction outputs two data bytes in succession, from register pair H, into the input buffer registers of the DAC. The contents of register L are transferred into the 8-bit shift register, while the least significant two bits in register H go into the 2-bit shift register. Note that the address is automatically incremented, and a second MEMW control pulse generated by the 8080A when the SHLD instruction is executed. The STA <B2> <B3> instruction provides only a strobe pulse at the LDAC input to the DAC; no data transfer occurs between the accumulator and the DAC.

Other small monolithic and hybrid DAC systems are available from different manufacturers. The Analog Devices converter was chosen because of the on-the-chip latches and double buffering registers. The use of a reference potential is common to many DAC modules. Perhaps in the future it, too, will be included in the module.

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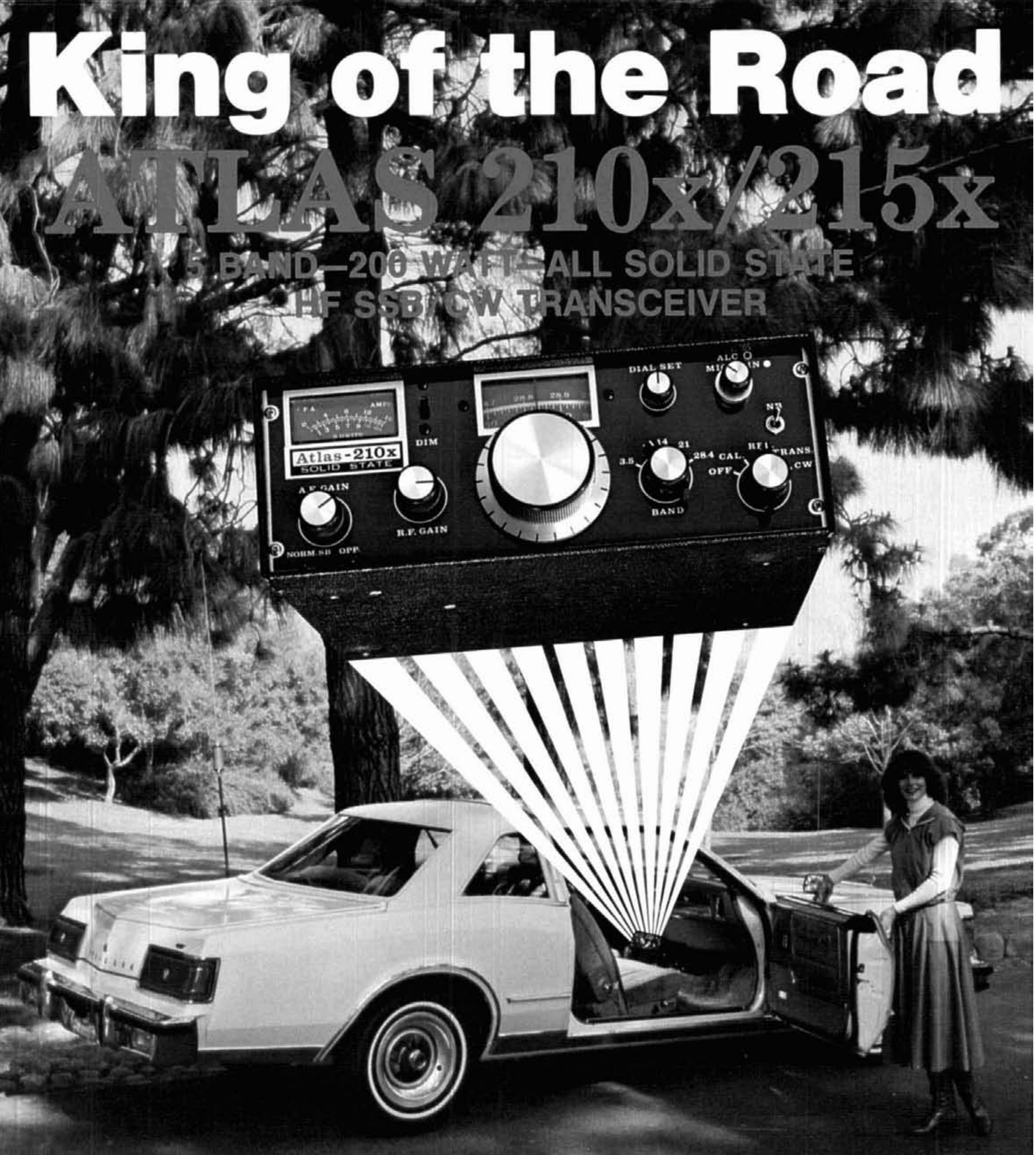

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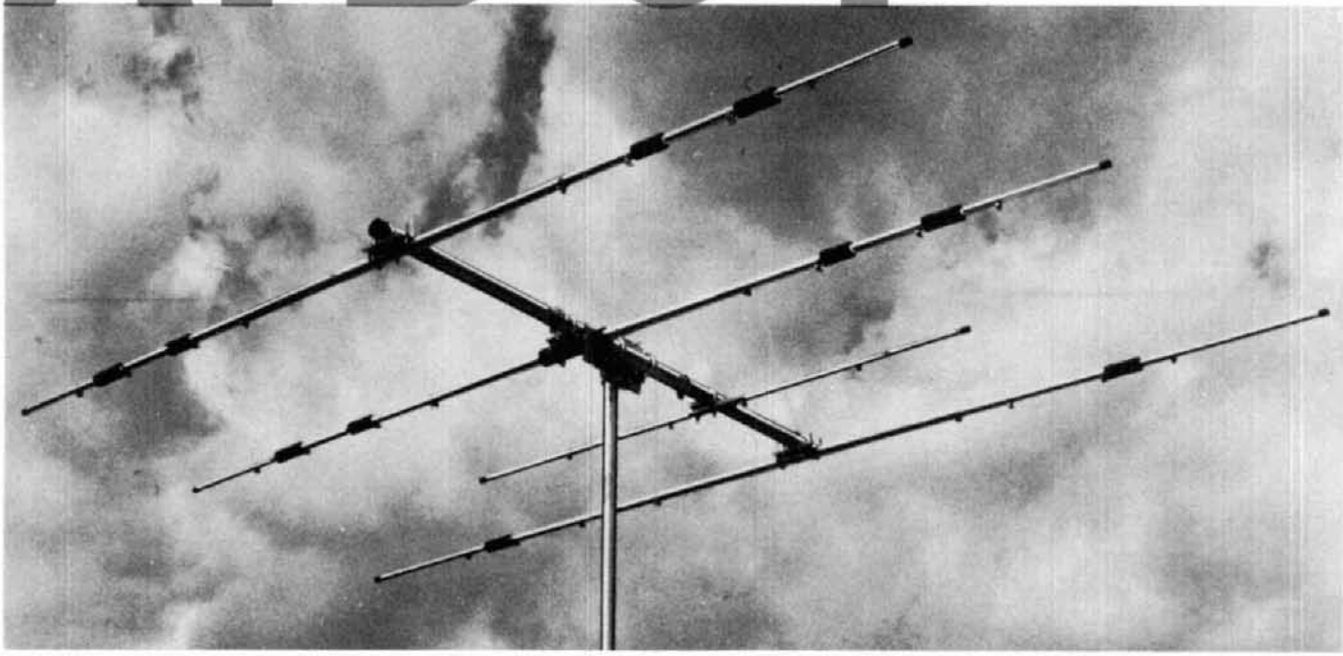

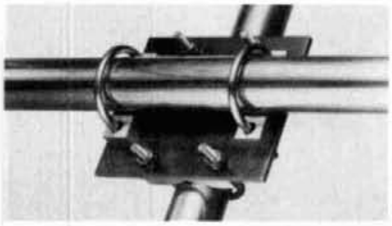

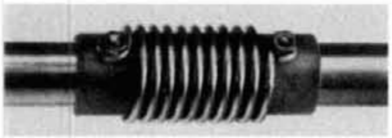

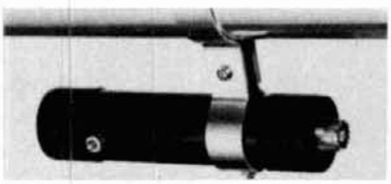

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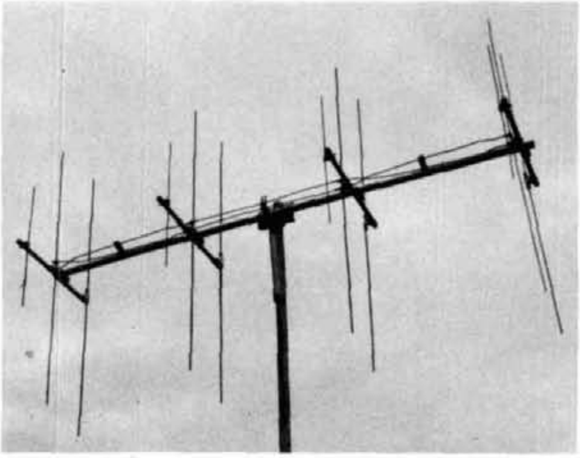

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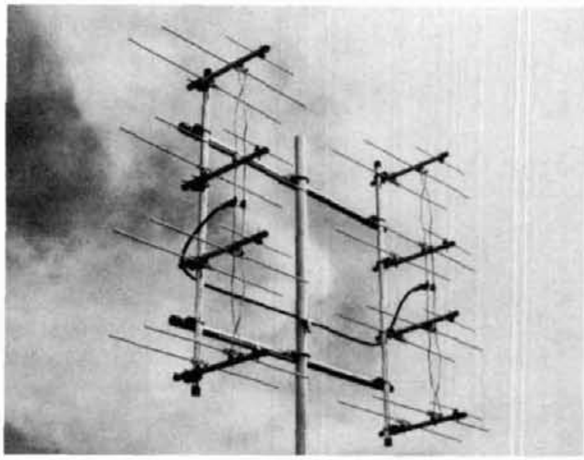

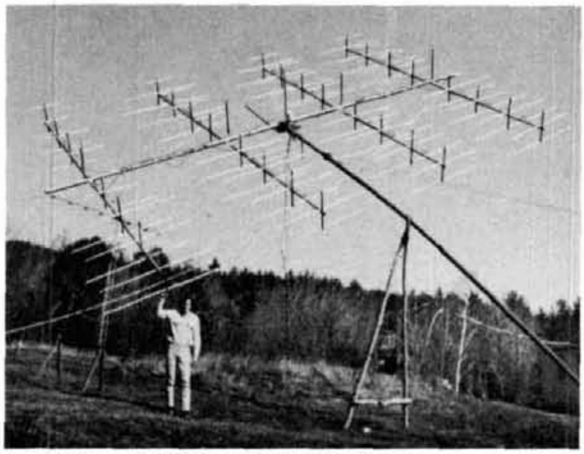

Dave Olean, K1WHS, with his 160 Element DX-Array and Polar Mount EME System

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More Details? CHECK—OFF Page 126

let's reduce audio pollution in the vhf bands

How's your audio quality over the repeater? Here are some ideas for improvements

Much has been written about the control and power output of vhf fm equipment. But little has been written, that I can discover, about the input part — audio signal generation — particularly the how and why of obtaining audio quality throughout the entire chain of repeater operation.

As a one-time audiophile I have paid close attention, the past several months, to the audio quality of signals being retransmitted by amateur repeaters. It's been a pretty sad experience, but there's a glimmer of hope. I've also been hearing others commenting on the situation, and steps are being taken to remedy it.

Other than from equipment electrical malfunctions, bad audio is caused mostly by *close talking* into the dynamic microphone that usually comes with the radio. A couple of years ago I wrote an article on the subject of proximity effect for my club paper. The article was subsequently published in *ham radio*.[1] It pointed out that all conventional moving-coil microphones (including speakers used as microphones) accentuated the bass tones by a factor of 3 to 4 dB at 200 Hz (**fig. 1**). Such response can easily cause the first audio amplifier stage to overload and feed this signal into the transmitter.

I've heard many reasons why amateurs don't remain at least 5 cm (2 inches) from their microphones, none of which are valid. Many old wives' tales have sprung up regarding the use of microphones. The most ridiculous is that foreign equipment manufacturers have "set the audio stages of their radios for their countrymen." Sheer nonsense! It's usually not necessary to touch any internal controls in any radio, regardless of its origin, foreign or domestic. This statement is qualified by reminding you that, in most transceivers, audio is set using one audio frequency at a time from an audio oscillator, and that the radio and microphone are packaged in a carton as the equipment goes out the door.

talking across the microphone

The next fallacious theory to be laid to rest is the "talk-across-the-microphone" nonsense. *All* micro-

By **Budd Meyer, K2PMA**, 6505 Yellowstone Blvd., Forest Hills, New York 11375

phones are designed for speaking directly into the active element (**fig. 2**). The sound waves should impinge onto the element perpendicularly; the microphone element should be in the same plane as your mouth. The microphone element is designed to react to, or transduce, all frequencies pretty much equally.

Unfortunately, microphone designers have a problem in that the higher audio frequencies are more directive. The further from a straight line access to the microphone element, the greater will be the loss of the higher audio frequencies reaching the microphone element. **Fig. 3** gives an idea of what happens when an audio signal (your voice) impinges indirectly onto the microphone element. Audio frequencies above about 800 Hz travel almost directly toward the element, whereas lower audio frequencies have much less directivity and much more power per unit bandwidth.

So if you talk *across* the microphone, the bass voice frequencies can overload the audio chain in the radio; and the high frequencies, which are essential to intelligibility, are attenuated. I'd like to emphasize that the information above applies to *all* microphones whether crystal, dynamic, ceramic, or whatever. If you talk across the microphone, you're going to sacrifice the higher audio frequencies.

over-deviation

Another situation that should be remedied is the idea that increasing the fm deviation will increase power output. More nonsense! The rf output of an fm transmitter is constant, and no amount of "dinking" with the deviation control will increase power

Test setup used by K2PMA to check VU meter readings with a calibrated audio power meter.

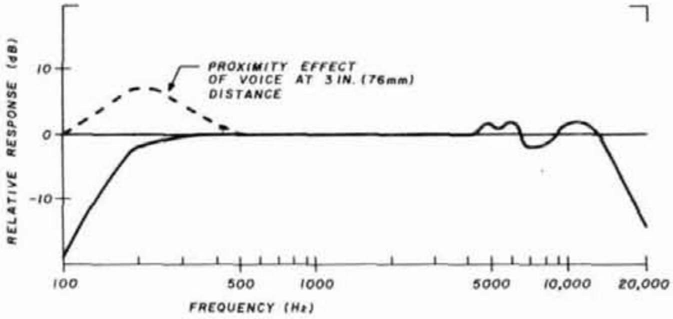

fig. 1. Generalized response curve of the dynamic microphone. Note the accentuated bass tones at 200 Hz (reference 1).

output. If you adjust the deviation control so that the radio is over-deviating, all you will do is increase distortion, since the transmitted signal bandwidth will be increased.

Most amateur repeaters have some form of deviation limiting control to prevent their signals from spreading out. If your radio's deviation setting is above this predetermined limit, you'll probably experience the phenomenon of "popping out the repeater." You won't gain anything but shrugs from other repeater users who know better. You may even get an admonishment to clean up your signal.

some answers to the problem

How do you obtain good audio quality through the repeater? Much of the answer lies in the desire to do so. A standard answer you'll get, as I have, when you mention to someone who is over-deviating that perhaps he might try backing away from the microphone, is "You're the first to complain." Don't let this answer put you off. Most amateurs will back away from the microphone if asked to do so. Of course, if you tell him he sounds better at *that* distance from the microphone, and he realizes he's a half-meter (2 feet) away, he just *might* consider adjusting his microphone gain control!

simple tests

Some time ago I came across a good buy in volume unit (VU) meters for my stereo system. I purchased an additional meter and connected it to the output of my vhf transceiver, which uses an 8-ohm speaker. I now had a reference of sorts — the fact that it was qualitative rather than quantitative served my purpose. At a comfortable listening level, the VU meter indicated average and peak audio output. To get quantitative values, I connected a calibrated audio-power meter to the output, which proved the validity of my VU-meter measurements.

Being aware that most amateurs don't have VU or audio-power meters, I connected a Simpson 260 Vom to the speaker output. Lo and behold — the VU-meter and Simpson 260 Vom readings matched! Such instrumentation can be used to advise the fel-

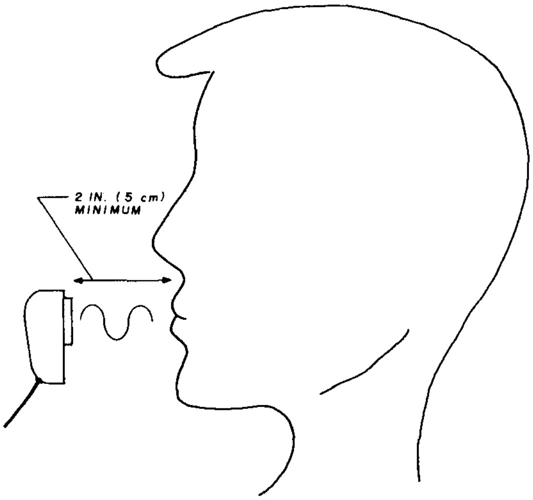

fig. 2. Most fm problems with poor audio can be remedied by speaking at least 5 cm (2 inches) from the microphone. Loud speech doesn't improve power output.

low on the other end that he's over-deviating, and you can prove it by comparison.

As a rough approximation, I suggest to the other fellow that he adjust his level to cause my VU meter, which has 300 ohms rather than the standard 600-ohm impedance, to average −7 VU and peak at −4 VU. (A correction factor of 15.75 dB must be added to the reading.) This comes out to about 7.5 mW average and 15 mW peak. These numbers are only guides; you can set your own standards based on your own equipment.

measurement problems

Audio measurements are subjective, which means that all such measurements obtained by instrumentation are subject to individual taste and hearing characteristics. I've listened to rigs set up by deviation meters and oscilloscopes and note that this method just doesn't accomplish the job. There are resolution problems with oscilloscopes and frequency problems with deviation meters. Personal habits and voice characteristics color the tests. Obviously an oscilloscope presentation will show clipping; however, by the time the clipping is discernible the distortion will probably be very high.

How can we obtain good audio? Probably 90 per cent of the distortion heard on repeaters can be eliminated if we all talked at least 5 cm (2 inches) away from the microphone. Also, accepting the fact that talking loudly does not equate to more output power will eliminate the other major culprit causing distortion, as noted previously.

I've no recommendation for setting up metered standards since an infinite number of variables must be considered. And no matter what one could come up with, we still must contend with the subjective aspect of audio measurements. I've heard many amateurs with unbearable distortion access a repeater and ask for an audio check, only to be told that they sound good. The smartest thing to do would be to listen to your favorite repeater for awhile and learn which people give accurate reports. If one fellow consistently *improves* the audio of others with honest reports, then he's your man.

level setting

Here are some ideas for setting the level of your radio. As an example of the typical imported fm transceiver, I've provided a couple of sketches showing where to find the deviation and microphone gain control (**fig. 4**).

First, you must have someone listen to your signal over a period of time, preferably on a simplex frequency, so that the repeater doesn't affect your audio. Start by practicing the technique of talking at *least 5 cm (2 inches) from the microphone*. Once you get the hang of it, this procedure will become second

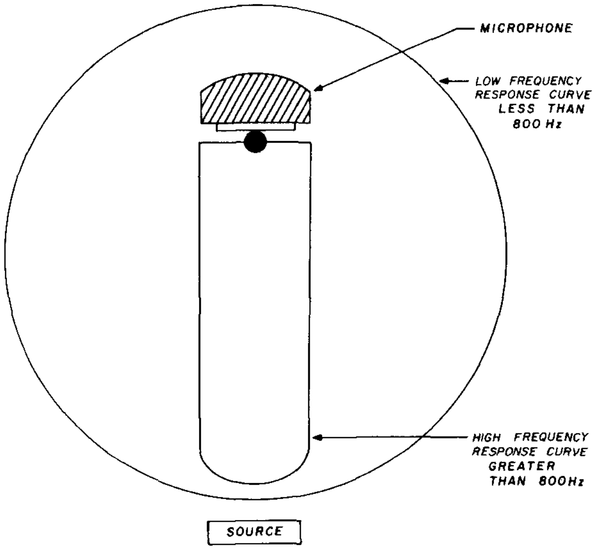

fig. 3. Response patterns of dynamic microphones. Higher audio frequencies, which convey intelligence, are more directional than the lower frequencies — a case supporting the theory for talking directly into the microphone rather than across it.

nature. Chances are you won't have to touch the radio's innards.

If you have to reduce the audio, locate the microphone gain control (not the deviation control), and turn it in small increments until your listener agrees

the microphone

As a final attempt to clean up your signal, you can always buy a new microphone. Perhaps yours is out of spec, or you dropped it once too often. The thing to look for here is not so much impedance match

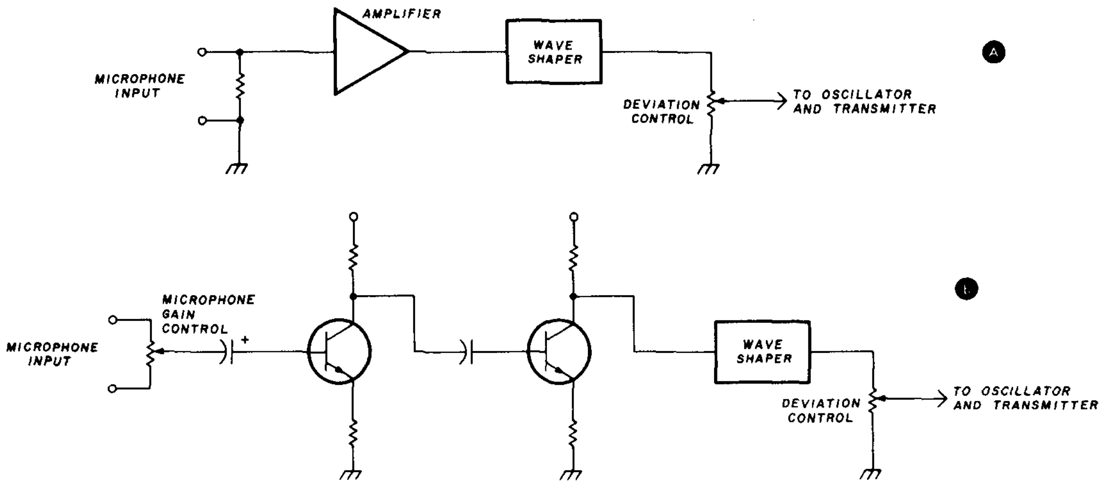

fig. 4. Generalized sketch of most fm vhf radios showing where to find the deviation control, A, and the microphone gain control, B.

that your audio is sufficient and clean — no rough edges. I don't believe that these tests can be done in committee fashion — stay with one man!

Only rarely is it necessary to touch the deviation control. What procedure do you use if there's only one control in your radio? Not much, really. The situation then becomes a tradeoff between microphone talking distance and level. Make every attempt to keep your audio clean. If your radio has only one control, not eating the microphone is bound to improve your signal quality. Some hand-held radios use the speaker as a microphone; not much can be done here.

If your radio has a deviation control but no microphone gain control, it's easy to adjust the deviation control so that your signal isn't two barn doors wide. As I've mentioned previously some repeaters are deviation limited, and if your fm signal swing is beyond the preset standard set by the repeater, the machine will pop out when you try to talk.

Locate the deviation control, which is generally near the end of the speech amplifier and audio chain. Ask your friend to listen to your signal while you adjust the deviation control for minimum swing with clean audio. Keep in mind that excessive deviation *does not* increase talk power or rf output. You can adjust the deviation control so that your signal will deliver all the talk power you can use without objectionable distortion.

(since most dynamic microphones will work with most transistorized amplifiers) but the microphone output level — somewhere in the vicinity of 10-50 mV across 2000 ohms. How do you ascertain this without meters? Buy it with return privileges.

closing remarks

Let us not forget that your transmitter is only one-third of the communications chain. Repeater operators should also get into the loop. Someone in the group should listen to the average quality of the repeater audio. Sad to say, many repeaters have all kinds of added exotic features, but repeated audio is poor because no one has taken the time to check it.

We have autopatch, beeps, timers, tone-encoded outputs, frequency checks, automatic signal reports, welcoming speeches, and the like, but bad audio in many cases. I earnestly believe that control stations should absolutely and positively comment on bad audio coming into the repeater.

I hope this article will help improve a long-neglected aspect of amateur radio. I've found most hams to be cooperative in helping to rid the airwaves of audio pollution.

reference

1. "Dynamic Microphones," *ham radio*, (Circuits & Techniques), June, 1976, page 49.

ham radio

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OSCAR amateur radio satellites

For most amateurs, obtaining information about any of the Oscar satellites has been somewhat of a piecemeal process. Almost every area of interest was covered in a separate publication or source, and some information just did not seem to be available at all.

Author and engineer Stratis Caramanolis has written a very useful book, OSCAR Amateur Radio Satellites, that ties it all together in one place. The original was published in German, and now an English-language version is being distributed by the Radio Society of Great Britain, and by Ham Radio's Communications Bookstore.

Almost any question that you may have about amateur radio satellites can be answered by consulting *OSCAR Amateur Radio Satellites.* It shows a high degree of competence in presenting the technical and theoretical facts of satellite operation, but at the same time these facts are covered in language that the non-engineer can understand and work with. The author uses just enough mathematics to prove whatever theory or point is being discussed, but not so much as to scare away any neophyte to amateur radio or satellite communications.

The book starts out with a discussion of the solar system from the first misconceptions that placed the earth at the center of the universe, and brings you right through the changes in scientific thinking into the current period of knowledge, which was started by Johannes Kepler. The first three laws formulated by Kepler are essential in understanding the mechanics of orbiting bodies, no matter what size they are. The book uses this Keplerian foundation to build a subsequent discussion of satellite orbital mechanics starting with a simple explanation of the trajectory of a projectile and then developing this into a trajectory that has no terminus, thus being an orbit.

Once the essential mechanics of orbits and satellite theory are taken care of, the author gets into a good explanation of satellite types, systems, control, power requirements, and structures. This is followed by a chapter on the fundamentals of telecommunications by means of satellites, which includes such areas as free-space loss (of signal strength) and receiver sensitivity — both important considerations if you plan to use a satellite system.

All of this groundwork material takes up a good portion of the book, but it is interesting and essential reading if you are to really understand what makes any satellite behave the way it does. Chapter seven is the first one that is devoted entirely to amateur radio satellites, and it starts with the background of amateur work in planning, building, and launching the first OSCAR. It continues with a history of each amateur satellite right through to the current OSCARs 6 and 7.

This chapter is followed by one that, logically, devotes many pages to detailed descriptions of all the systems that go to make up an OSCAR satellite package — transponders, telemetry, antennas, receivers, modes, decoding, and much more. It explains how you can calculate orbital information, how you can interpret the telemetry data that indicates the condition of the spacecraft systems, and what you need at your home (ground) station to work with satellites.

The last chapter is devoted to the subject of learning with satellites. It is a section designed to acquaint educators with the possibilities of classroom use of amateur satellite signals and experiments. The author points out some of the uses that have been made of this material, and provides names and addresses where those interested can obtain more information. Some samples of techniques useful in classrooms are given, such as measurement and explanation of Doppler effects, orbital data, and transmission of amateur television and weather satellite pictures through the OSCAR spacecraft. Future satellites are discussed, but because of the many variables in timing of launch vehicles, satellite construction, and the like, this coverage can be only tentative at best.

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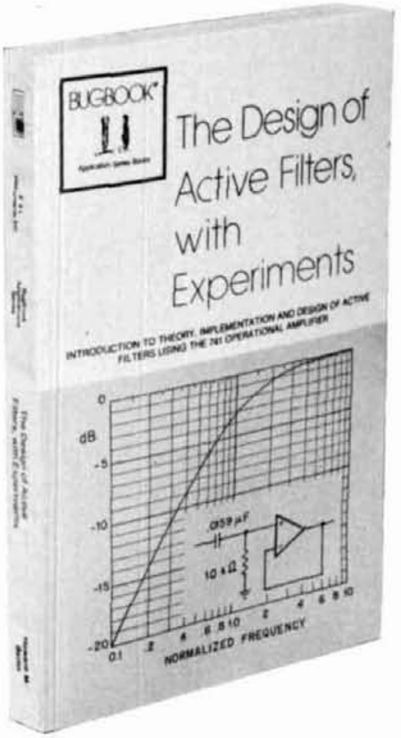
All-in-all, the book is recommended as an excellent reference for the amateur (or anyone else) who is curious about any phase of satellites, or who is seriously contemplating the use of OSCAR for communication or educational purposes. *OSCAR Amateur Radio Satellites*, by Stratis Caramanolis, is 194 pages, paperback, and is $8.50 postpaid from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048; order RS-O.

active filter design



The Design of Active Filters and Experiments by Howard M. Berlin, published by E&L Instruments, is an introduction to theory, implementation, and design of active filters. The text includes descriptions and schematic diagrams of designs for the various lowpass, highpass, bandpass, and notch filters based on the 741 operational amplifier IC. Recommended for the experimenter and hobbyist who want to learn the basics of active filters, the book is complete enough to serve as a college-level laboratory manual. It is also an excellent reference for the practicing circuit designer.

Mathematical equations and techniques are held to a minimum; the filter design characteristics are conveyed primarily by means of tables, graphs, and illustrative examples. The subjects covered include the basics of operational amplifiers and filters, first order lowpass and highpass active filters, and second and higher order active filters. Other topics described are bandpass and notch filters as well as the state-variable filter. Twenty-six experiments demonstrate active-filter characteristics.

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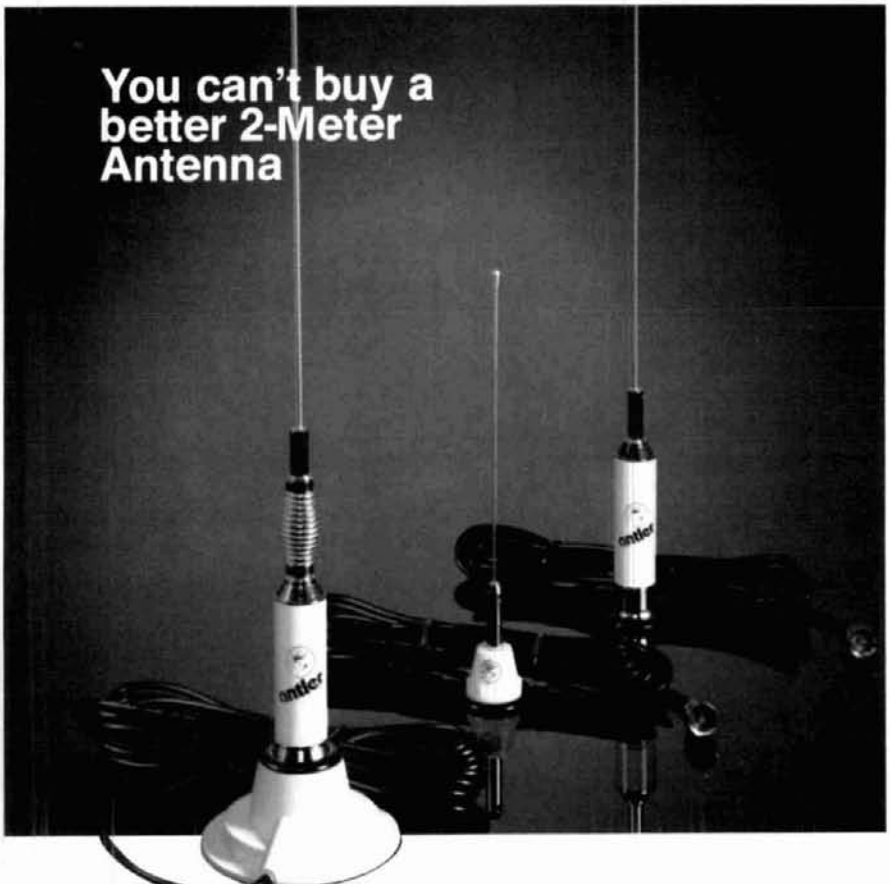

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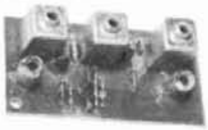

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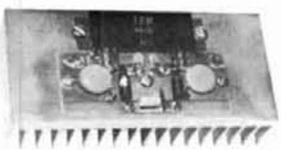

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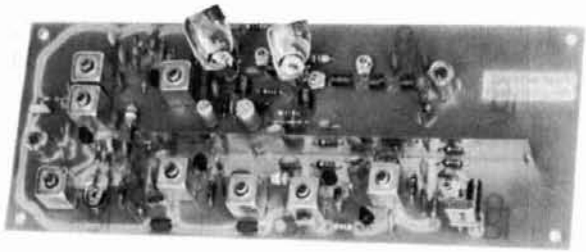

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C144	144-146	28-30
C145	145-147 (OSCAR)	28-30
C146	146-148	28-30
C110	Aircraft	28-30
C220	220 band	28-30
Special	Other i-f & rf ranges available	

MODEL	RF RANGE (MHZ)	I-F RANGE
C432-2	432-434	28-30
C432-5	435-437 (OSCAR)	28-30
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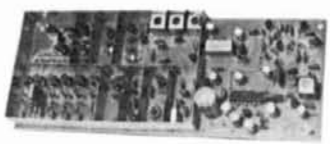

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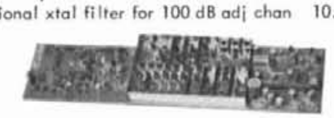

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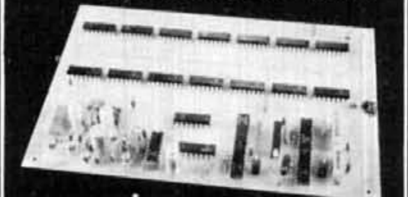

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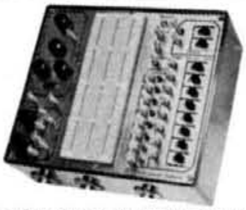

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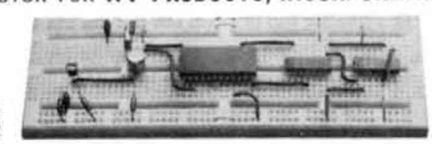

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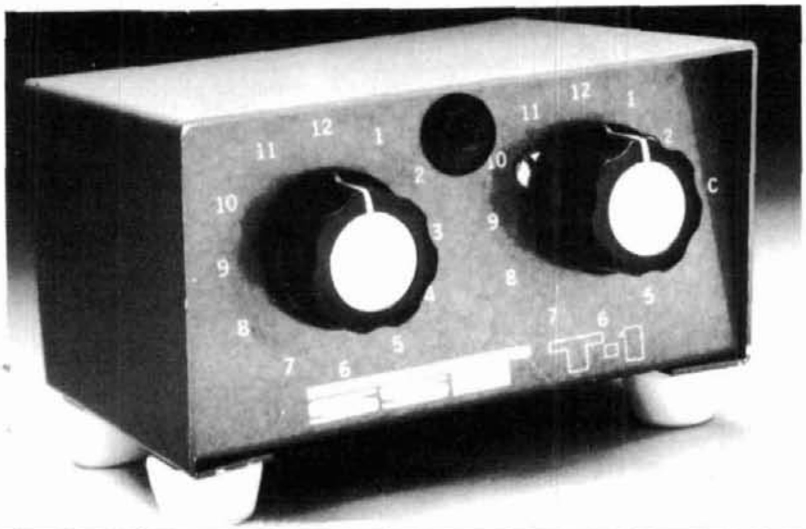

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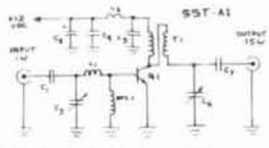

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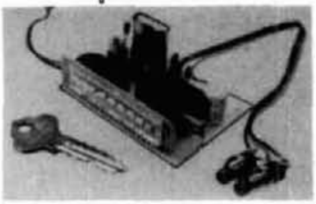

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AUTHORIZED DEALER for DenTron, KLM, Larsen, Bearcat, etc., Big Catalog 201-962-4695 Narwid Electronics, 61 Bellot Road, Ringwood, N.J. 07456.

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WANTED: Hallicrafters HA-20 VFO. Write or call W1QUT, 22 Woodridge Road, Wayland, MA 01778. 617-358-4953.

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QSL FORWARDING SERVICE — 30 cards per dollar. Write: QSL Express, 30 Lockwood Lane, West Chester, PA. 19380.

Coming Events

MASSACHUSETTS: The Wellesley Amateur Radio Society is conducting its annual auction on Saturday, April 15, 1978, beginning at 11:00AM at the Wellesley High School Cafeteria on Rice Street, Wellesley, Massachusetts. Talk-in on –96:36, 04-64 and 52. Doors open at 10:00AM. Contact Kevin P. Kelly, WA1YHV, 7 Lawnwood Place, Charlestown, Massachusetts 02129.

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SN7425N	.29	SN74100N	1.25	SN74186N	15.00
SN7426N	.29	SN74107N	.39	SN74187N	6.00
SN7427N	.37	SN74109N	.95	SN74188N	3.95
SN7429N	.42	SN74116N	1.95	SN74190N	1.79
SN7430N	.25	SN74121N	.39	SN74191N	1.25
SN7432N	.25	SN74122N	.39	SN74192N	.89
SN7437N	.35	SN74123N	.50	SN74193N	.89
SN7438N	.35	SN74125N	.60	SN74194N	1.25
SN7439N	.25	SN74126N	.60	SN74195N	.75
SN7440N	.21	SN74132N	1.25	SN74196N	1.00
SN7441N	.89	SN74136N	.95	SN74197N	1.00
SN7442N	.69	SN74141N	1.15	SN74198N	1.75
SN7443N	.75	SN74142N	2.95	SN74199N	1.75
SN7444N	.75	SN74143N	3.25	SN74200N	5.59
SN7445N	.75	SN74144N	3.00	SN74251N	1.79
SN7446N	.89	SN74145N	1.15	SN74279N	.90
SN7447N	.69	SN74147N	2.35	SN74283N	2.25
SN7448N	.89	SN74148N	2.00	SN74284N	6.00
SN7450N	.25	SN74150N	1.25	SN74285N	6.00
SN7451N	.25	SN74151N	.79	SN74365N	.75
SN7453N	.25	SN74153N	.89	SN74366N	.75
SN7454N	.25	SN74154N	1.25	SN74367N	.75
SN7459A	.25	SN74155N	.89	SN74368N	.75
SN7460N	.25	SN74156N	.89	SN74390N	2.25
SN7470N	.45	SN74157N	.89	SN74393N	2.25

20% Discount for 100 Combined 7400's

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CD4000	.23	CD4044	.89	CD4520	1.29
CD4001	.23	CD4046	1.79	MC14562	14.50
CD4002	.23	CD4047	2.50	CD4566	2.25
CD4006	1.19	CD4048	1.35	MC14583	3.50
CD4007	.25	CD4049	.49	**74C00 Series**	
CD4009	.49	CD4050	.49	74C00	.39
CD4010	.49	CD4051	1.19	74C02	.55
CD4011	.23	CD4053	1.19	74C04	.75
CD4012	.25	CD4056	1.49	74C10	.65
CD4013	.39	CD4059	9.95	74C14	3.00
CD4014	1.39	CD4060	1.49	74C20	.65
CD4015	1.19	CD4066	.79	74C30	.65
CD4016	.49	CD4068	.39	74C42	2.15
CD4017	1.19	CD4069	.45	74C73	1.50
CD4018	.99	CD4070	.55	74C74	1.15
CD4019	.49	CD4071	.23	74C89	4.00
CD4020	1.19	CD4072	.49	74C90	3.00
CD4021	1.39	CD4076	1.39	74C93	2.00
CD4022	1.19	CD4081	.23	74C95	2.00
CD4023	.23	CD4082	.23	74C107	1.25
CD4024	.79	CD4098	2.49	74C151	2.90
CD4025	.23	MC14409	14.95	74C154	3.00
CD4026	2.25	MC14410	14.95	74C157	2.15
CD4027	.69	MC14411	14.95	74C160	3.25
CD4028	.89	MC14419	4.95	74C161	3.25
CD4029	1.19	MC14506	.75	74C163	3.00
CD4030	.49	MC14507	.99	74C164	3.25
CD4035	.99	CD4508	3.95	74C173	2.60
CD4040	1.19	CD4510	1.39	74C193	2.75
CD4041	1.25	CD4511	1.29	74C195	2.75
CD4042	.99	CD4515	2.95	80C95	1.50
		CD4518	1.29	80C97	1.50

LINEAR

LM300H	.80			LM739N	1.19
LM301H	.35	LM340T-8	1.25	LM741CH	.35
LM301CN	.35	LM340T-12	1.25	LM741CN	.35
LM302H	.75	LM340T-15	1.25	LM741-14N	.39
LM304H	1.00	LM340T-18	1.25	LM747H	.79
LM305H	.60	LM340T-24	1.25	LM747N	.79
LM307CN	.35	LM350N	1.00	LM748H	.39
LM308H	1.00	LM351CN	.65	LM748N	.39
LM308CN	1.00	78MG	1.75	LM1303N	.90
LM309H	1.10	LM370N	1.15	LM1304N	1.19
LM309K	1.25	LM373N	3.25	LM1305N	1.40
LM310CN	1.15	LM377N	4.00	LM1307N	.85
LM311H	.90	LM380N	1.25	LM1310N	2.95
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LM318CN	1.50	LM382N	1.79	LM1458CN	.59
LM319N	1.30	NE501K	8.00	LM1496N	.95
LM320K-5	1.35	NE510A	6.00	LM1556V	1.75
LM320K-5.2	1.35	NE529A	4.95	LM2111N	1.95
LM320K-12	1.35	NE531H	3.00	LM2901N	2.95
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LM320T-15	1.25	NE561B	5.00	LM5556N	1.85
LM320T-18	1.25	NE562B	5.00	MC5558V	1.00
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LM340K-12	1.35	LM709N	.29	75454CN	.39
LM340K-15	1.35	LM710N	.79	75491CN	.79
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74LS00	.29			74LS155	1.25
74LS02	.29			74LS157	1.50
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74LS05	.35	74LS83	1.75	74LS162	1.95
74LS08	.29	74LS85	2.49	74LS163	1.95
74LS10	.29	74LS86	.49	74LS164	1.95
74LS13	.69	74LS90	.89	74LS175	1.95
74LS14	1.75	74LS92	.89	74LS181	3.69
74LS20	.29	74LS93	.89	74LS190	2.49
74LS26	.39	74LS95	1.50	74LS191	2.49
74LS27	.39	74LS96	1.89	74LS192	2.49
74LS28	.39	74LS107	.59	74LS193	2.49
74LS30	.29	74LS109	.59	74LS194	1.89
74LS32	.39	74LS112	.59	74LS195	1.89
74LS40	.39	74LS123	1.25	74LS253	1.75
74LS42	1.25	74LS132	1.25	74LS257	1.75
74LS47	1.25	74LS136	.59	74LS260	.55
74LS51	.29	74LS138	1.25	74LS279	.79
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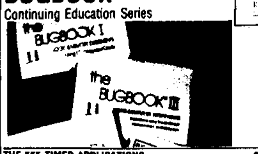

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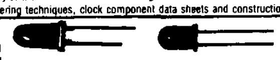

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XC209	Green	4/$1	XC111	Green	4/$1
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XC22	Yellow	4/$1	XC526	Green	4/$1	XC556	Green	4/$1	INFRA-RED LED
XC22	Orange	4/$1	XC526	Yellow	4/$1	XC556	Yellow	4/$1	1/4" x 1/4" x 1/16"
SSL-22	RT	4/$1	XC526	Clear	4/$1	XC556	Orange	4/$1	Flat 5/$1.00
						XC556	Clear	7/$1	

SALE DISPLAY LEDS SALE

TYPE	POLARITY	HT	PRICE	TYPE	POLARITY	HT	PRICE
MAN 1	Common Anode-red	.270	2.95	MAN 6680	Common Cathode-orange	.560	.79
MAN 2	5 x 7 Dot Matrix-red	.300	4.95	MAN 6710	Common Anode-red-D.D.	.560	.79
MAN 3	Common Cathode-red	.125	4/$1	MAN 6730	Common Anode-red ±1	.560	.79
MAN 4	Common Cathode-red	.187	1.95	MAN 6740	Common Cathode-red-D.D.	.560	.79
MAN 52	Common Anode-green	.300	1.00	MAN 6750	Common Cathode-red ±1	.560	.79
MAN 71	Common Anode-red	.300	1.25	MAN 6760	Common Anode-red	.560	.79
MAN 72	Common Anode-red	.300	.79	MAN 6780	Common Cathode-red	.560	.79
MAN 74	Common Cathode-red	.300	1.50	DL701	Common Anode-red ±1	.300	1.00
MAN 81	Common Anode-yellow	.300	.79	DL702	Common Cathode-red	.300	1.25
MAN 82	Common Anode-yellow	.300	.79	DL704	Common Cathode-red	.300	.99
MAN 84	Common Cathode-yellow	.300	.79	DL707	Common Anode-red	.300	.99
MAN 3620	Common Anode-orange	.300	.79	DL741	Common Anode-red	.600	1.49
MAN 3630	Common Anode-orange ±1	.300	1.35	DL746	Common Anode-red ±1	.630	1.95
MAN 3640	Common Cathode-orange	.300	.79	DL747	Common Anode-red	.600	1.49
MAN 4610	Common Anode-orange	.300	.79	DL749	Common Cathode-red ±1	.630	1.95
MAN 4640	Common Cathode-orange	.400	.79	DL750	Common Cathode-red	.600	1.49
MAN 4710	Common Anode-red ±1	.400	.79	DL338	Common Cathode-red	.110	3/1.00
MAN 4730	Common Anode-red	.400	1.00	FND70	Common Cathode (FND359)	.250	.69
MAN 4740	Common Cathode-red	.400	.79	FND503	Common Cathode (FND500)	.500	.99
MAN 4810	Common Anode-yellow	.400	1.00	FND507	Common Anode (FND510)	.500	.99
MAN 6610	Common Anode-orange-D.D.	.560	.79	5082-7300	4 x 7 Sgl. Digit-RHDP	.600	19.95
MAN 6630	Common Anode-orange	.560	.79	5082-7302	4 x 7 Sgl. Digit-LHDP	.600	19.95
MAN 6640	Common Cathode-orange-D.D.	.560	.79	5082-7304	Overrange character (±1)	.600	15.00
MAN 6650	Common Cathode-orange ±1	.560	.79	5082-7340	4 x 7 Sgl. Digit-Hexadecimal	.600	22.50
MAN 6660	Common Anode-orange	.560	.79				

RCA LINEAR

CA3013	2.15	CA3082	2.00
CA3023	2.56	CA3083	1.60
CA3035	2.48	CA3086	.85
CA3039	1.35	CA3089	3.75
CA3046	1.30	CA3091	3.50
CA3053	1.50	CA3102	2.95
CA3059	3.25	CA3123	2.15
CA3060	3.25	CA3130	1.39
CA3080	.85	CA3140	1.25
CA3081	2.00	CA3401	.49
		CA3600	1.75

EXAR

XR-2206KB Kit $19.95 — XR-2206KA Kit $14.95

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XR-2207CP	3.85	XR-1468	3.85	XR-556CP	1.85
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XR-1310CP	$3.20	XR-1489	4.80	XR-2240CP	4.80
XR-1310EP	3.20	XR-2208	5.20	**PHASE LOCKED LOOPS**	
XR-1800P	3.20			XR-210	5.20
XR-2567	2.99			XR-215	6.60
				XR-567CP	1.95
				XR-567CT	1.70

IC SOLDERTAIL — LOW PROFILE (TIN) SOCKETS

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.17	.16	.15	24 pin	$.38	.37	.36
14 pin	.20	.19	.18	28 pin	.45	.44	.43
16 pin	.22	.21	.20	36 pin	.60	.59	.58
18 pin	.29	.28	.27	40 pin	.63	.62	.61
22 pin	.37	.36	.35				

SOLDERTAIL STANDARD (TIN)

	1-24	25-49	50-100		1-24	25-49	50-100
14 pin	$.27	.25	.24	28 pin	$.99	.90	.81
16 pin	.30	.27	.25	36 pin	1.39	1.26	1.15
18 pin	.35	.32	.30	40 pin	1.59	1.45	1.30
24 pin	.49	.45	.42				

SOLDERTAIL STANDARD (GOLD)

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.30	.27	.24	24 pin	$.70	.63	.57
14 pin	.35	.32	.29	28 pin	1.10	1.00	.90
16 pin	.38	.35	.32	36 pin	1.75	1.40	1.26
18 pin	.52	.47	.43	40 pin	1.75	1.59	1.45

WIRE WRAP SOCKETS (GOLD) LEVEL #3

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.40	.38	.35	22 pin	.95	.85	.75
10 pin	.45	.41	.37	24 pin	$1.05	.95	.85
14 pin	.39	.38	.37	28 pin	1.40	1.25	1.10
16 pin	.43	.42	.41	36 pin	1.59	1.45	1.30
18 pin	.75	.68	.62	40 pin	1.75	1.55	1.40

50 PCS. RESISTOR ASSORTMENTS $1.75 PER ASST.

ASST. 1	5 ea.	10 OHM / 27 OHM	12 OHM / 33 OHM	15 OHM / 39 OHM	18 OHM / 47 OHM	22 OHM / 56 OHM	1/4 WATT 5%	50 PCS.
ASST. 2	5 ea.	68 OHM / 180 OHM	82 OHM / 220 OHM	100 OHM / 270 OHM	120 OHM / 330 OHM	150 OHM / 390 OHM	1/4 WATT 5%	50 PCS.
ASST. 3	5 ea.	470 OHM / 1.2K	560 OHM / 1.5K	680 OHM / 1.8K	820 OHM / 2.2K	1K / 2.7K	1/4 WATT 5%	50 PCS.
ASST. 4	5 ea.	3.3K / 8.2K	3.9K / 10K	4.7K / 12K	5.6K / 15K	6.8K / 18K	1/4 WATT 5%	50 PCS.
ASST. 5	5 ea.	22K / 56K	27K / 68K	33K / 82K	39K / 100K	47K / 120K	1/4 WATT 5%	50 PCS.
ASST. 6	5 ea.	150K / 390K	180K / 470K	220K / 560K	270K / 680K	330K / 820K	1/4 WATT 5%	50 PCS.
ASST. 7	5 ea.	1M / 2.7M	1.2M / 3.3M	1.5M / 3.9M	1.8M / 4.7M	2.2M / 5.6M	1/4 WATT 5%	50 PCS.

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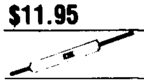

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SW7417	2.20	19.00	180.00	SW7491	3.50	31.00	300.00
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SW7425	1.70	14.00	130.00	SW7496	3.50	31.00	300.00
SW7427	1.50	12.00	110.00	SW74100	6.90	66.00	650.00
SW7430	1.50	12.00	110.00	SW74104	2.20	19.00	180.00
SW7432	1.50	12.00	110.00	SW74105	2.20	19.00	180.00
SW7437	1.50	12.00	110.00	SW74107	2.20	19.00	180.00
SW7438	1.50	12.00	110.00	SW74121	2.20	19.00	180.00
SW7440	1.20	9.00	80.00	SW74123	2.50	22.00	200.00
SW7443	3.50	31.00	300.00	SW74145	5.50	51.00	500.00
SW7444	3.50	31.00	300.00	SW74150	6.90	66.00	650.00
SW7445	3.50	31.00	300.00	SW74151	3.50	31.00	300.00
SW7446	4.00	36.00	350.00	SW74153	4.00	36.00	350.00
SW7450	1.20	9.00	80.00	SW74156	4.00	36.00	350.00
SW7453	1.20	9.00	80.00	SW74180	4.50	41.00	400.00
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Pre-tubed • No mixing or combining prices

TV GAME CHIP SET — $7.95

Includes AY-3-8500-1 Chip and 2.010 mhz crystal
(2.010 crystal — $.99 ea/AY-3-8500-1 Chip — $7.50 ea.)

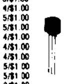
ZENERS — DIODES — RECTIFIERS

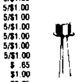
TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
1N746	3.3	400m	4/1.00	1N4005	600 PIV	1 AMP	10/1.00
1N751A	5.1	400m	4/1.00	1N4006	800 PIV	1 AMP	10/1.00
1N752	5.6	400m	4/1.00	1N4007	1000 PIV	1 AMP	10/1.00
1N753	6.2	400m	4/1.00	1N3600	50	200m	6/1.00
1N754	6.8	400m	4/1.00	1N4148	75	10m	15/1.00
1N959	8.2	400m	8/1.00	1N4154	35	10m	12/1.00
1N965B	15	400m	4/1.00	1N4305	75	25m	20/1.00
1N5232	5.6	500m	.28	1N4734	5.6	1w	.28
1N5234	6.2	500m	.28	1N4735	6.2	1w	.28
1N5235	6.8	500m	.28	1N4736	6.8	1w	.28
1N5236	7.5	500m	.28	1N4738	8.2	1w	.28
1N456	25	40m	6/1.00	1N4742	12	1w	.28
1N458	150	7m	6/1.00	1N4744	15	1w	.28
1N485A	180	10m	6/1.00	1N1183	50 PIV	35 AMP	1.60
1N4001	50 PIV	1 AMP	12/1.00	1N1184	100 PIV	35 AMP	1.70
1N4002	100 PIV	1 AMP	12/1.00	1N1185	150 PIV	35 AMP	1.50
1N4003	200 PIV	1 AMP	12/1.00	1N1186	200 PIV	35 AMP	1.80
1N4004	400 PIV	1 AMP	12/1.00	1N1188	400 PIV	35 AMP	3.00

SCR AND FW BRIDGE RECTIFIERS

C36D	15A @ 400V	SCR	$1.95
C38M	35A @ 200V	SCR	1.95
2N2328	1.6A @ 200V	SCR	.50
MDA 980-1	12A @ 50V	FW BRIDGE REC.	1.95
MDA 980-3	12A @ 200V	FW BRIDGE REC.	1.95

TRANSISTORS





MPS A05	.30	2N3638	5/$1.00	PN4250	4/$1.00
MPS A06	5/$1.00	2N3702	5/$1.00	2N4400	4/$1.00
2N918	4/$1.00	2N3704	5/$1.00	2N4401	4/$1.00
2N2219A	3/$1.00	2N3705	5/$1.00	2N4402	4/$1.00
2N2221	4/$1.00	2N3706	5/$1.00	2N4403	4/$1.00
2N2222A	5/$1.00	2N3707	5/$1.00	2N4409	5/$1.00
2N2369	5/$1.00	2N3711	5/$1.00	2N5086	4/$1.00
2N2369A	4/$1.00	2N3724	$.65	2N5087	4/$1.00
2N2484	4/$1.00	2N3725	$1.00	2N5088	4/$1.00
2N2906A	4/$1.00	2N3772	$2.25	2N5089	4/$1.00
2N2907A	5/$1.00	2N3903	5/$1.00	2N5129	5/$1.00
2N2925	5/$1.00	2N3904	4/$1.00	2N5138	5/$1.00
2N3035	2/$1.00	2N3905	4/$1.00	2N5139	5/$1.00
2N3055	$.89	2N3906	4/$1.00	2N5209	5/$1.00
MJE2955	$1.25	2N4013	3/$1.00	2N5210	5/$1.00
MJE3055	$1.00	2N4014	3/$1.00	2N5432	$2.00
2N3392	5/$1.00	2N4123	6/$1.00	2N5449	3/$1.00
2N3398	5/$1.00	PN4249	4/$1.00	2N5951	5/$1.00
PN3567	3/$1.00			C106B1SCR	2/$1.00
PN3568	4/$1.00			40409	$1.75
PN3569	4/$1.00			40410	$1.75

CAPACITOR CORNER

50 VOLT CERAMIC DISC CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
10 pf	.05	.04	.03	.001μF	.05	.04	.035
22 pf	.05	.04	.03	.0047μF	.05	.04	.035
47 pf	.05	.04	.03	.01μF	.05	.04	.035
100 pf	.05	.04	.03	.022μF	.06	.05	.04
220 pf	.05	.04	.03	.047μF	.06	.05	.04
470 pf	.05	.04	.035	.1μF	.12	.09	.075

100 VOLT MYLAR FILM CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
.001mf	.12	.10	.07	.022mf	.13	.11	.08
.0022	.12	.10	.07	.047mf	.21	.17	.13
.0047mf	.12	.10	.07	.1mf	.27	.23	.17
.01mf	.12	.10	.07	.22mf	.33	.27	.22

+20% DIPPED TANTALUMS (SOLID) CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
.1/35V	.28	.23	.17	1.5/35V	.30	.26	.21
.15/35V	.28	.23	.17	2.2/25V	.31	.27	.22
.22/35V	.28	.23	.17	3.3/25V	.31	.27	.22
.33/35V	.28	.23	.17	4.7/25V	.32	.28	.23
.47/35V	.28	.23	.17	6.8/25V	.36	.31	.25
.68/35V	.28	.23	.17	10/25V	.40	.35	.29
1.0/35V	.28	.23	.17	15/25V	.63	.50	.40

MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS

Axial Lead	1-9	10-49	50-100	Radial Lead	1-9	10-49	50-100
.47/50V	.15	.13	.10	.47/25V	.15	.13	.10
1.0/50V	.16	.14	.11	.47/50V	.16	.14	.11
3.3/50V	.14	.13	.10	1.0/16V	.15	.13	.10
4.7/25V	.16	.14	.12	1.0/25V	.16	.14	.11
10/25V	.15	.13	.10	1.0/50V	.16	.14	.11
10/50V	.16	.14	.12	4.7/16V	.15	.13	.10
22/25V	.17	.15	.12	4.7/25V	.15	.13	.10
22/50V	.24	.20	.18	4.7/50V	.16	.14	.11
47/25V	.19	.17	.15	10/16V	.14	.12	.09
47/50V	.25	.21	.19	10/25V	.15	.13	.10
100/25V	.24	.20	.18	10/50V	.16	.14	.12
100/50V	.35	.30	.28	47/50V	.24	.21	.19
220/25V	.32	.28	.25	100/16V	.19	.15	.14
220/50V	.45	.41	.38	100/25V	.24	.20	.18
470/25V	.33	.29	.27	100/50V	.35	.30	.28
1000/16V	.55	.50	.45	220/16V	.23	.17	.16
2200/16V	.70	.62	.55	470/25V	.31	.28	.26

More Details? CHECK—OFF Page 126

flea market

RARS SIXTH ANNUAL HAMFEST, April 23, Crabtree Valley Mall, US 70 West, Raleigh, NC. Big, big flea market — all under cover. Fantastic prizes. Ladies activities, meetings. Walk to nearby motels, restaurants, shopping. More info? Write RARS, Box 17124, Raleigh, NC 27609.

ZERO DISTRICT QSO PARTY, organized by the Mississippi Valley Radio Club; 2000Z, Saturday, April 22nd to 0200Z, Monday, April 24th. Stations outside of zero district work zero stations *only*, while zero stations may work any station. Same station once each band, each mode, except for special mobile stations which may be worked each time they change counties. All stations exchange QSO number and ARRL section; all zero stations must send county. Suggested frequencies are, Phone: 3900, 7270, 14300, 21370, 28570; CW: 3560, 7060, 14060, 21060, 28060; Novice: 3725, 7125, 21125, 28125. Awards to high scorer in each ARRL section and DX country. Also top Novice/Technicians and special mobile stations. For information/log forms (S.A.S.E.) WBØUUA, 3518 W. Columbia, Davenport, Iowa 52804.

HOLIDAY — IN — DIXIE FESTIVAL sponsored by ARCOS (Amateur Radio Club of Shreveport, LA) from 1800Z, April 21st, through 2400Z, April 30th, 1978. Contact a station within 75 miles of Shreveport, exchange RS(T), power, ARRL section or country. ARCOS members will be on air as a group from 1800Z, April 22nd, to 2400Z, April 23rd, but QSOs can be made anytime during the ten-day festival. Suggested CW: 3555, 3710, 7055, 7110, 14055, 21130, and 28130; Phone: 3935, 3975, 7235, 14280, 21380, and 28575; VHF: 50.110 +/− SSB, and 52.525 +/− simplex FM; 145.100 +/− SSB, and 145.015 +/− simplex FM. Local contacts for demos on 07/67 and 16/76. Locals work each other. For extra points ask members for their Holiday-in-Dixie number. Info and logs to Holiday-in-Dixie QSO Party, P.O. Box 1485, Shreveport, LA 71164.

MOULTRIE AMATEUR RADIO KLUB, 17th annual hamfest, April 23, 1978 at Wyman Park, Sullivan, Illinois. Heated indoor, and large outdoor, fleamarket. No charge to vendors. Info from M.A.R.K., Box 327, Mattoon, IL 61938. Talk in on .94.

ALABAMA FORESTRY FESTIVAL — April 22, 1978. The Twin Base Amateur Radio Club station WA4PRY will be operating on site in conjunction with the Alabama Forestry Festival at the State Fair Grounds, Montgomery, Alabama. Any ARS completeing a two-way contact will receive a special certificate in exchange for a QSL card and SASE. Operations will be conducted from 1600 hours to 2300 hours UTC on frequencies 14.300 MHz and 3.950 MHz normal SSB; slow C.W. (5 to 10 WPM) on 7.125 MHz during even hours UTC and 21.150 MHz during odd hours UTC. QRM frequency adjustments will be up band.

MIDWEST SPRING CONVENTION — Saturday, April 1, 1978, Holiday Inn-Holidome, Kearney, Nebraska sponsored by the Midway Amateur Radio Club. Flea market, auction, technical symposiums, ARRL Forum, State MARS Meeting, State QCWA meeting, special "Ladies Day", and evening banquet. Over $2,500.00 worth of door prizes up for grabs with a Wilson Mark II with touchtone presented to a ham and, for the lucky "Lady" — a microwave oven. Over 200 other prizes to be given away. For registration and further details contact Chuck, WØCRK, Midwest Spring Convention, 3605 Third Avenue, Kearney, Nebraska 68847.

ARIZONA: Tucson Hamfest, April 28-30, 1978 Ramada Inn (just off North I-10) Technical session with demonstrations, Microprocessors, Solar Poser, QRP, Fast/Slow Scan, RTTY, Remote Base, etc. Prizes, Ladies Programs, Banquet, Exhibits, Swap Meet. Sponsored by Old Pueblo Radio Club. Write OPRC, 1361 E. Edlin, Tucson, AZ 85711.

ATLANTIC CITY AREA HAMFEST Sunday, April 16th, Rain or Shine. 200 Indoor, 400 Outdoor table spaces; Stockton State College Campus, Pomona NJ; Lots of dealers; $1 advance, $1.50 gate; Door prizes; More info SPARC, P.O. Box 142, Absecon, NJ 08201.

27TH DAYTON HAMVENTION at Hara Arena, April 28, 29, 30, 1978. More room this year! Technical forums, exhibits and huge flea market. Program brochure mailed March 6th to those registered within past three years. For accommodations or advance flyer, write Hamvention, P.O. Box 44, Dayton, OH 45401 or call 513-854-4126.

THE 4TH ANNUAL NORTHWESTERN PENNSYLVANIA HAMFEST, May 6th, Crawford County Fairgrounds, Meadville, PA. Gates open at 8:00 $2 prize ticket required for admission — $1 to display. Children FREE. Hourly door prizes, refreshments, commercial displays welcome. Indoors if rain. Talk-in 04/64 and 52. Details CARS, P.O. Box 653, Meadville, PA. 16335.

More Details? CHECK—OFF Page 126

NEW

Frequency Counter

$79.95 kit

UTILIZES NEW MOS-LSI CIRCUITRY

You've requested it, and now it's here! The CT-50 frequency counter kit has more features than counters selling for twice the price. Measuring frequency is now as easy as pushing a button, the CT-50 will automatically place the decimal point in all modes, giving you quick, reliable readings. Want to use the CT-50 mobile? No problem, it runs equally as well on 12 V dc as it does on 110 V ac. Want super accuracy? The CT-50 uses the popular TV color burst freq. of 3.579545 MHz for time base. Tap off a color TV with our adapter and get ultra accuracy — .001 ppm! The CT-50 offers professional quality at the unheard of price of $79.95. Order yours today!

CT-50, 60 MHz counter kit $79.95
CT-50 WT, 60 MHz counter, wired and tested 159.95
CT-600, 600 MHz prescaler option for CT-50, add 29.95

SPECIFICATIONS
Sensitivity: less than 25 mv.
Frequency range: 5 Hz to 60 MHz, typically 65 MHz
Gatetime: 1 second, 1/10 second, with automatic decimal point positioning on both direct and prescale
Display: 8 digit red LED .4" height
Accuracy: 10 ppm, .001 ppm with TV time base!
Input: BNC, 1 megohm direct, 50 Ohm with prescale option
Power: 110 V ac 5 Watts or 12 V dc @ 1 Amp
Size: Approx. 6" x 4" x 2", high quality aluminum case
Color burst adapter for .001 ppm accuracy available in 6 weeks.
CB-1, kit $14.95

CLOCK KIT
6 digit 12/24 hour

Want a clock that looks good enough for your living room? Forget the competitor's kludges and try one of ours! Features: jumbo .4" digits, Polaroid lens filter, extruded aluminum case available in 5 colors, quality PC boards and super instructions. All parts are included, no extras to buy. Fully guaranteed. One to two hour. assembly time. Colors: silver, gold, black, bronze, blue (specify).

Clock kit, DC-5 . . . . $22.95
Alarm clock, DC-8, 12 hr only 24.95
Mobile clock, DC-7 . . . . 25.95
Clock kit with 10 min ID timer, DC-10 . . . 25.95
Assembled and tested clocks available, add $10.00

VIDEO TERMINAL KIT $149.95

A compact 5 x 10 inch PC card that requires only an ASCII keyboard and a TV set to become a complete interactive terminal for connection to your microprocessor asynchronous interface. Its many features are single 5 volt supply, crystal controlled sync and baud rates (up to 9600 baud), 2 pages of 32 characters by 16 lines, read to and from memory, computer and keyboard operated cursor and page control, parity error display and control, power on initialization, full 64 character ASCII display, block type see thru cursor. Keyboard/computer control backspaces, forward spaces, line feeds, rev. line feeds, home, returns cursor. Also clears page, clears to end of line, selects page 1 or 2, reads from or to memory. The card requires 5 volts at approx. 900 ma and outputs standard 75 ohm composite video.

TH3216 Kit . . . $149.95
TH3216, Assembled and Tested . . . 239.95
VD 1, Video to RF Modulator Kit . . . 6.95

CAR CLOCK KIT $27.95

12/24 Hour 12 Volt AC or DC
- High Accuracy (1 minute/month)
- 6 jumbo 4" LED readouts
- Easy, no polarity hookup
- Display blanks with ignition
- Case, mounting bracket included
- Super instructions

Complete Kit, DC-11 $27.95

AUTO-DIMMER $2.50
Automatically adjusts display brightness according to ambient light level. For DC 11 Car Clock.

CHEAP CLOCK KIT $8.95

DC-4 Features:
- 6 digit .4" LED
- 12 or 24 format

Does not include board or transformer

PC Board $2.95
Transformer $1.49

MINI-KITS

TONE DECODER KIT

A complete tone decoder on a single PC Board. Features: 400-5000 Hz adjustable frequency range, voltage regulation, 567 IC. Useful for touch-tone decoding, tone burst detection, FSK demod, signaling, and many other uses. Use 7 for 12 button touchtone decoding. Runs on 5 to 12 volts.
Complete Kit, TD-1 . . . . $4.95

SUPER SLEUTH AMPLIFIER

A super-sensitive amplifier which will pick up a pin drop at 15 feet! Great for monitoring baby's room or as a general purpose test amplifier. Full 2 watts of output, runs on 6 to 12 volts, uses any type of mike. Requires 8-45 ohm speaker.
Complete Kit, BN-9 . . . . $4.95

FM WIRELESS MIKE KIT

Transmit up to 300' to any FM broadcast radio, uses any type of mike. Runs on 3 to 9 V. Type FM-2 has added super sensitive mike preamp.
FM-1 . . . . . . $2.95 FM-2 . . . . . . $4.95

COLOR ORGAN/MUSIC LIGHTS

See music come alive! 3 different lights flicker with music or voice. One light for lows, one for the mid-range and one for the highs. Each channel individually adjustable, and drives up to 300 watts. Great for parties, band music, nite clubs and more.
Complete Kit, ML-1 . . . . $7.95

LED BLINKY KIT

A great attention getter which alternately flashes 2 Jumbo LEDs. Use for name badges, buttons, or warning type panel lights. Runs on 3 to 9 volts.
Complete Kit $2.95

POWER SUPPLY KIT

Complete triple regulated power supply provides variable ±15 volts at 200 mA and +5 volts at 1 Amp. 50 mV load regulation good filtering and small size. Kit less transformers. Requires 6-8 V at 1 Amp and 18 to 30 VCT.
Complete Kit, PS-3LT $6.95

SIREN KIT

Produces upward and downward wail characteristic of police siren. 5 watts audio output, runs on 3-9 volts, uses 8-45 ohm speaker.
Complete Kit, SM-3 . . . . $2.95

DECADE COUNTER PARTS

Includes: 7490A, 7475, 7447, LED readout, current limit resistors, and instructions on an easy to build low cost frequency counter.
Kit of parts, DCU-1 . . . . $3.50

600 MHz PRESCALER

Extend the range of your counter to 600 MHz. Works with all counters. Less than 150 mv sensitivity. Specify ÷10 or ÷100
Wired, tested, PS-1B $59.95
Kit, PS-1B $44.95

30 watt 2 meter Power Amp

The famous RE class C power amp now available mail order! Four Watts in for 30 Watts out, 2 in for 15 out, 1 in for 8 out, incredible value, complete with all parts, instructions and details on T-R relay. Case not included.
Complete Kit, PA-1 . . . . $22.95

CALENDAR ALARM CLOCK

Has every feature one could ever ask for. Kit includes everything except case, build it into wall, station or even car!
FEATURES:
- 6 Digits, .5" High LED
- 12/24 Hour Format
- Calendar shows mo./day
- Snooze button
- True 24 Hour Alarm
- 7001 chip does all!!
- Battery back up with built-in on chip time base

Complete Kit, less case, DC-9 . . . . $34.95

		LINEAR		REGULATOR		TRANSISTORS	
5314 Clock	$2.95	555	$ .50	78MG	$1.49	MRF-238 30W VHF	$11.95
74S00	.35	556	.75	309k	.89	NPN 2N3904 type	10/$1.00
74S112	.75	566	1.49	309H	.99	PNP 2N3906 type	10/$1.00
7447	.79	567	1.49	340K-12	.99	NPN Power Tab 40W	3/$1.00
7473	.35	1458	.50	7805	.89	PNP Power Tab 40W	3/$1.00
7475	.50	LED DRIVER		7812	.89	FET MPF-102 type	3/$2.00
7490A	.55	75491	.50	7815	.89	UJT 2N2646 type	3/$2.00
74143	3.50	75492	.50	7818	.89	2N3055 NPN Power	.75

SPECIAL

DIODES: 1KV,2.5A . . . . 5/$1.00 100V,1A . . . . 10/$1.00 1N914A type . . . . 50/$2.00

LED DISPLAYS

FND 359	.75
FND 510	1.25
DL 707	1.25
HP 7730	1.25

Red Polaroid Filter . . . 4.25" X 1.125" 59

741 OP-AMP SPECIAL
Factory prime mini dip with both Xerox and 741 part numbers
10 for $2.00

SOCKETS

14 PIN	5/$1.00
16 PIN	5/$1.00
24 PIN	2/$1.00
40 PIN	3/$2.00

FERRITE BEADS
with info and specs
15/$1.00
6 hole Balun Beads
5/$1.00

ramsey electronics
P.O. Box 4072 Rochester NY 14610
(716) 271-6487

TELEPHONE ORDERS WELCOME

master charge
BANKAMERICARD

Minimum Order $6.00

flea market

20TH ANNUAL S.E.M.A.R.A. HAMFEST April 9th, 1978, 8:00 A.M. until 3:00 P.M. at South Lake High School, 21900 E. Nine Mile Road at Mack Ave., St. Clair Shores, Michigan (in the Detroit Michigan metropolitan area).

TRI-COUNTY ARC ANNUAL HAMFEST May 7. 9:00 A.M. to 4:00 P.M. Rain or Shine — Indoors at Stirling, NJ Youth Center, just off Valley Road. Tables $5.00 Buyers $1.00. Door Prize — Bird Wattmeter. Talk-in on 146.52 and 147.855-147.255. Info from Tri-County ARC P.O. Box 412, Scotch Plains, NJ 07076 or call Herb W2CHA 201-647-3461.

HOSSTRADERS NET: Fifth annual Tailgate Swapfest Saturday, May 13th at Deerfield, New Hampshire Fairgrounds. (Covered buildings in case of rain.) Admission one dollar, no commission or percentage. Commercial dealers welcome at same rate. Excess revenues benefit Boston Burns Unit of Shriner's Hospital for Crippled Children. Last year we donated $430.80. Talk-in .52, 146.40-147.00, 3940 kHz. Questions: S.A.S.E. to Joe Demaso K1RQG, Star Rt. Box 56, Bucksport, ME 04416 or Norm Blake WA1IVB, P.O. Box 32, Cornish, ME 04020 or Check Hosstraders Net on Sundays 4 PM 3940 kHz.

KENTUCKY HAM-O-RAMA — Sunday, May 28 (Memorial Day Weekend). 7 minutes south of Cincinnati. Erlanger Lions Park, Erlanger, Kentucky. Donaldson Road exit, I-75 South. Prizes, exhibits, flea market. NKARC, Box 31, Ft. Mitchell, Kentucky 41017.

THE RADIO CLUB DE PUERTO RICO announces its 1978 Annual Convention and Hamfest at the famous Hotel El Conquistador in Fajardo, P.R., the weekend of April 28-30. For details, GPO Box 693, San Juan, PR 00936.

RADIO SOCIETY OF BERMUDA CONTEST, 0001 GMT April 22nd to 2400 GMT April 23rd, 1978; operate 36 hours of 48, single operator only; 3.5, 7.0, 14.0, 21.0 and 28.0 MHz bands. Top scorer in each state, county, province to receive printed award. Top scorer in each country to receive trophy presented *in Bermuda* at RSB Annual Dinner in October. For full information, send large S.A.S.E. plus 13¢ postage to Bernie Swandic, K3DH, 7417 Mill Run Drive, Derwood, MD 20855.

F.M. B*A*S*H, DAYTON, OHIO, April 28, 1978, Friday night of DAYTON HAMVENTION. Social evening for hams and friends, 8 P.M til midnight. *NEW LOCATION*: Downtown Dayton Convention Center, Main @ Fifth. Admission is free. Sandwiches, beverages, snacks and C.O.D. bar available. Live entertainment by TV personality Rob Reider (WA8GFF) and his group. Fabulous prize drawing featuring a complete Drake UV-3, including 144, 220 & 440 MHz synthesized modules, power supply, encoder mike and antenna plus other prizes. Winner of first prize need not be present. For further information contact: Miami Valley F.M. Assn. c/o Sue Hagedon, WB8GWQ, 1340 Brainard Woods Drive, Dayton, Ohio 45459.

ROCHESTER REPEATER SOCIETY HAMFEST, April 15th, from 9:00 A.M., at St. John's Grade School, 420 West Center Street, Rochester, Minnesota. Door prize donation $1; admission $1; Children under 12 FREE. Tables $2.50; plenty of parking. Talk-in 22/82 WRØAFT, and 52 simplex. Take I-90 to route 52 or 63 and go north. Advance ticket sales and information from KØTS, Joe Fishburn, 2514 4th Ave., N.W., Rochester, MN 55901; tel: (507) 288-2676; or Gary Sharp, WD8AMA, 1610 34th St., N.W., Rochester, MN 55901; tel: (507) 282-5119.

6TH ANNUAL MADISON SWAPFEST, Sunday, April 9th, Dane County Expo Center — Youth Building — Madison, Wisconsin. Doors open at 8:00 A.M., rain or shine; indoor facilities; food; movies; family fun; overnite camping accommodations. Tickets: $1.50 Advance; $2 at door; Tables: $2 Advance; $3 at door. Featuring electronic equipment and components for hams, hobbyists, computer & experimenters. Follow Coliseum highway signs from I-90/12-18 "beltline" interchange to "MC" 10 miles exit. Talk-in on 16/76. Write M.A.R.A., Box 3403, Madison, WI 53704. Reservations must be in by April 1st!

LAKE SHORE RADIO FEST, sponsored by Sheboygan County ARC; Saturday, April 8th, 9 A.M. to 5 P.M., Wilson Town Hall, Sheboygan, Wisconsin. Ham, marine, CB gear on display; door prizes; food; special YL-XYL section. Adults $2, children w/family FREE. Talk-in on 66/06 and 40 meters. Take I-43 to *business* exit — Route 141. For information, call (414) 452-7570.

FRAMINGHAM, MASSACHUSETTS ARC Flea Market, April 29th, doors open 9 A.M. for setup, public at 10 A.M. Location Framingham police drill shed; take Route 9 to Route 126, go south on 126 about 2.5 miles. Drill shed is behind Civic League building. Table $5. Half-tables also available. Contact F.A.R.C., Box 3005, Framingham, MA; or K1AZE, (617) 879-7456, or WA1UEH, (617) 653-6398.

More Details? CHECK—OFF Page 126

DIODES/ZENERS

1N914	100v	10mA	.05
1N4005	600v	1A	.08
1N4007	1000v	1A	.15
1N4148	75v	10mA	.05
1N753A	6.2v	z	.25
1N758A	10v	z	.25
1N759A	12v	z	.25
1N4733	5.1v	z	.25
1N5243	13v	z	.25
1N5244B	14v	z	.25
1N5245B	15v	z	.25

SOCKETS/BRIDGES

8-pin	pcb	.25	ww	.45
14-pin	pcb	.25	ww	.40
16-pin	pcb	.25	ww	.40
18-pin	pcb	.25	ww	.75
22-pin	pcb	.45	ww	1.25
24-pin	pcb	.35	ww	1.10
28-pin	pcb	.35	ww	1.45
40-pin	pcb	.50	ww	1.25

Molex pins	.01	To-3 Sockets	.45
2 Amp Bridge		100-prv	1.20
25 Amp Bridge		200-prv	1.95

TRANSISTORS, LEDS, etc.

2N2222A	NPN (2N2222 Plastic .10)	.15
2N2907A	PNP	.15
2N3906	PNP (Plastic)	.10
2N3904	NPN (Plastic)	.10
2N3054	NPN	.35
2N3055	NPN 15A 60v	.50
T1P125	PNP Darlington	.35
LED Green, Red, Clear, Yellow		.15
D.L.747	7 seg 5/8" High com-anode	1.95
XAN72	7 seg com-anode (Red)	1.25
MAN71	7 seg com-anode (Red)	1.25
MAN3610	7 seg com-anode (Orange)	1.25
MAN82A	7 seg com-anode (Yellow)	1.25
MAN74A	7 seg com-cathode (Red)	1.50
FND359	7 seg com-cathode (Red)	1.25

C MOS

4000	.15
4001	.15
4002	.20
4004	3.95
4006	.95
4007	.35
4008	.95
4009	.45
4010	.45
4011	.20
4012	.20
4013	.40
4014	.95
4015	.90
4016	.35
4017	1.10
4018	1.10
4019	.50
4020	.85
4021	1.00
4022	.85
4023	.25
4024	.75
4025	.30
4026	1.95
4027	.50
4028	.95
4030	.35
4033	1.50
4034	2.45
4035	1.25
4040	1.35
4041	.69
4042	.95
4043	.95
4044	.95
4046	1.75
4049	.45
4050	.45
4066	.95
4069	.40
4071	.35
4081	.70
4082	.45
MC14409	14.50
MC14419	4.85

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7400	.15
7401	.15
7402	.20
7403	.20
7404	.15
7405	.25
7406	.35
7407	.55
7408	.25
7409	.15
7410	.10
7411	.25
7412	.30
7413	.35
7414	1.10
7416	.25
7417	.40
7420	.15
7426	.30
7427	.45
7430	.15
7432	.30
7437	.30
7438	.35
7440	.25
7441	1.15
7442	.45
7443	.65
7444	.45
7445	.65
7446	.95
7447	.95
7448	.65
7450	.25
7451	.25
7453	.20
7454	.25
7460	.40
7470	.45
7472	.40
7473	.25
7474	.30
7475	.35
7476	.40
7480	.55
7481	.75
7483	.95
7485	.75
7486	.25
7489	1.35
7490	.55
7491	.95
7492	.95
7493	.35
7494	.75
7495	.60
7496	.80
74100	1.15
74107	.35
74121	.35
74122	.55
74123	.55
74125	.45
74126	.35
74132	1.35
74141	.90
74150	.85
74151	.65
74153	.75
74154	.95
74156	.95
74157	.65
74161	.85
74163	.85
74164	.60
74165	1.50
74166	1.35
74175	.80
74176	1.25
74180	.75
74181	2.25
74182	.95
74190	1.75
74191	1.05
74192	.75
74193	.85
74194	1.25
74195	.95
74196	1.25
74197	1.25
74198	2.35
74221	1.00
74367	.85
75108A	.35
75110	.35
75491	.50
75492	.50
74H00	.15
74H01	.25
74H04	.20
74H05	.20
74H08	.35
74H10	.35
74H11	.35
74H15	.45
74H20	.30
74H21	.25
74H22	.40
74H30	.20
74H40	.25
74H50	.25
74H51	.25
74H52	.15
74H53J	.25
74H55	.20
74H72	.45
74H101	.75
74H103	.75
74H106	.95
74L00	.25
74L02	.25
74L03	.30
74L04	.30
74L10	.30
74L20	.35
74L30	.45
74L47	1.95
74L51	.45
74L55	.65
74L72	.45
74L73	.40
74L74	.45
74L75	.55
74L93	.55
74L123	.85
74S00	.35
74S02	.35
74S03	.30
74S04	.30
74S05	.35
74S08	.35
74S10	.35
74S11	.35
74S20	.35
74S40	.20
74S50	.20
74S51	.25
74S64	.20
74S74	.35
74S112	.60
74S114	.65
74S133	.40
74S140	.55
74S151	.30
74S153	.35
74S157	.75
74S158	.30
74S194	1.05
74S257 (8123)	1.05
74LS00	.25
74LS01	.35
74LS02	.35
74LS04	.30
74LS05	.45
74LS08	.25
74LS09	.35
74LS10	.35
74LS11	.35
74LS20	.25
74LS21	.25
74LS22	.25
74LS32	.40
74LS37	.35
74LS40	.45
74LS42	1.10
74LS51	.50
74LS74	.65
74LS86	.65
74LS90	.95
74LS93	.95
74LS107	.85
74LS123	1.00
74LS151	.95
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74LS157	.85
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25 watts	—	25A	25C	25D	25E
50 watts	50H	50A	50C	50D	50E
100 watts	100H	100A	100C	100D	100E
250 watts	250H	250A	250C	250D	250E
500 watts	500H	500A	500C	500D	500E
1000 watts	1000H	1000A	1000C	1000D	1000E
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Headphone Impedance	3.2-20 ohms	2000 ohms	3.2-20 ohms	3.2-20 ohms	3.2-20 ohms	3.2-20 ohms	3.2-20 ohms	3.2-20 ohms
Microphone Frequency Response	–	–	–	–	50-8000 Hz	50-8000 Hz	50-8000 Hz	50-8000 Hz
Microphone Impedance	–	–	–	–	High	High	High	High
Microphone Sensitivity Below 1 volt/microbar at 1kHz	–	–	–	–	-51dB ±5dB	-51dB ±5dB	-51dB ±5dB	51dB ±5dB
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Model CM 610

Model C 1210

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Model C 610

Model CM 1210

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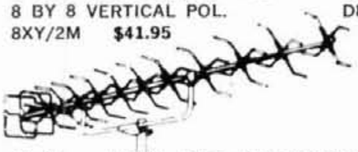

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- ". . . no-tune-up . . . high power operation requires no more adjustment than that involved in operating the exciter."
- ". . . there is no need to worry about whether the tubes or the power supply can handle a full kilowatt."
- "At no time . . . including two contest weekends when it was subject to constant use . . . did the writer feel that the amplifier capabilities were even close to being taxed." (QST, April 1975, p. 42-45.)

QST says it: ALPHA 76

- "Typically excellent ETO construction techniques . . . in the (ALPHA) 76."
- "The transformer . . . is a 1.5 KVA continuous-service unit."
- "The unit runs cool and quiet."
- "The ALPHA 76 more than exceeds harmonic attenuation (requirements.) . . . Third-order products are approximately 40 dB down . . . "
- ". . . 1 kW cw input provides excellent efficiency . . . "
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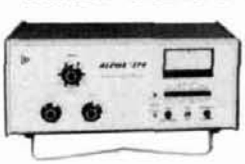

ALPHA 374
- No Tune Up, 80-10 meters!
- 2+ KW PEP, 1 KW avg., *NTL*
- RF output typically 1200+ watts PEP into 1.5:1 SWR
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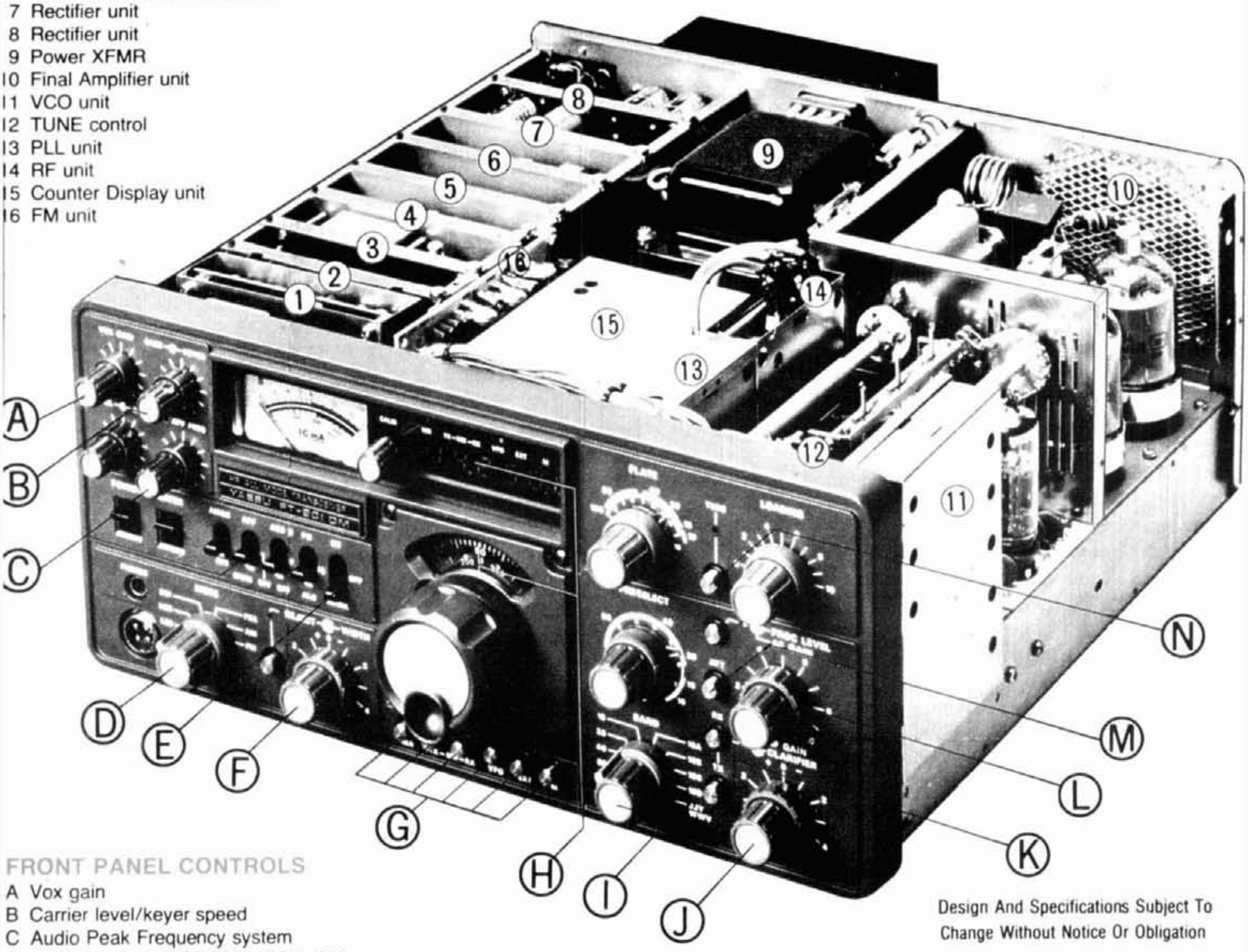

Design And Specifications Subject To Change Without Notice Or Obligation

FRONT PANEL CONTROLS

A Vox gain
B Carrier level/keyer speed
C Audio Peak Frequency system
D MODE switch (SSB, CW, FSK, AM, FM)
E Crystal calibrator/Noise blanker
F Rejection tuning/variable IF passband tuning
G Frequency memory system
H Digital plus analog frequency readout
I Band switch (160-10 meters + WWV/JJY receive)
J Clarifier control
K RX/TX Clarifier selector
L RF Processor level
M RF attenuator
N TUNE control (Places transmitter in "TUNE" condition for ten seconds, then returns to "receive" condition to protect final tubes from excessive key-down time)

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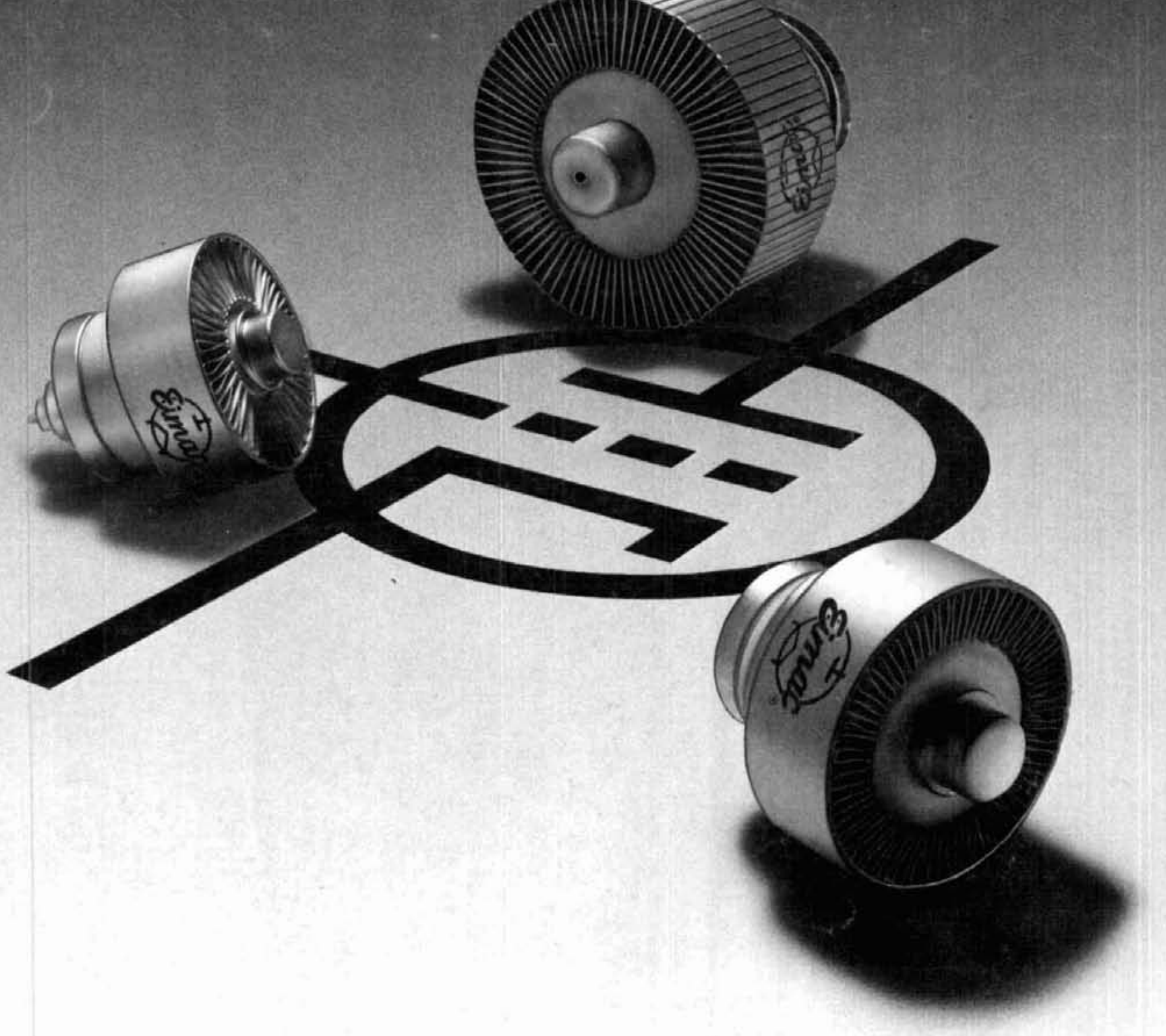

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