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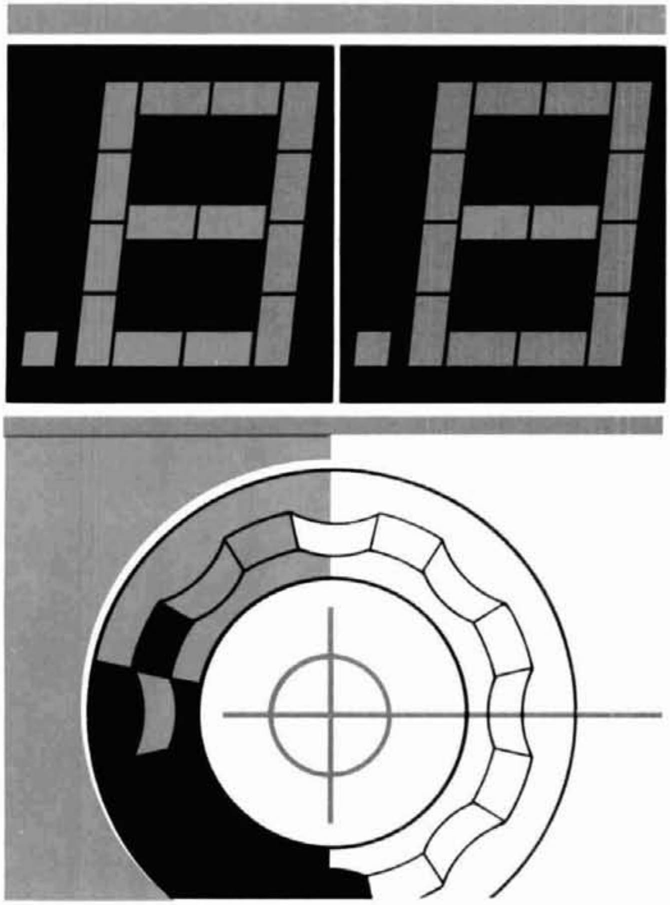
ham radio magazine

APRIL 1979

- propagation predictions 26
- memory keyer 32
- active filter for RTTY 46
- R-4C audio amplifier 48
- audio filter 62
- 432-MHz converter 74

40-meter receiver for home construction



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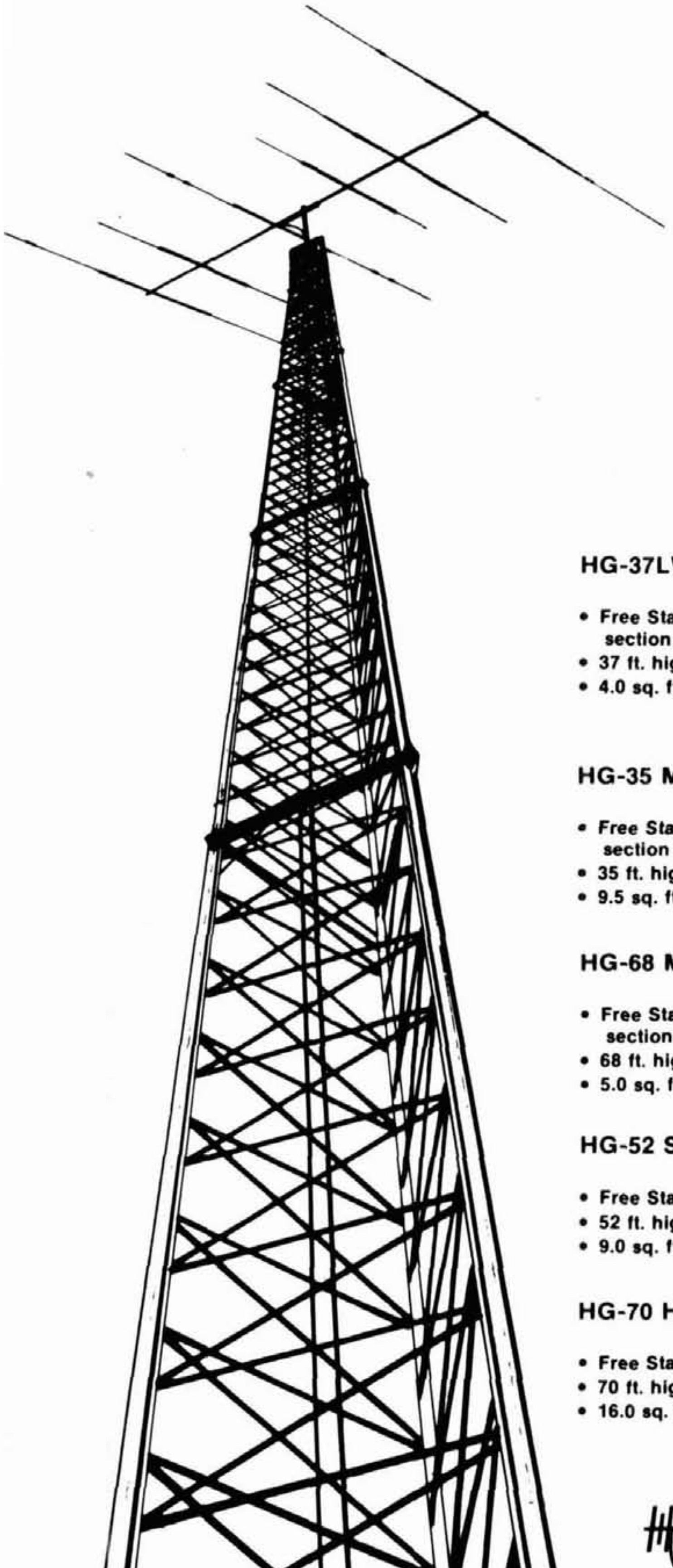

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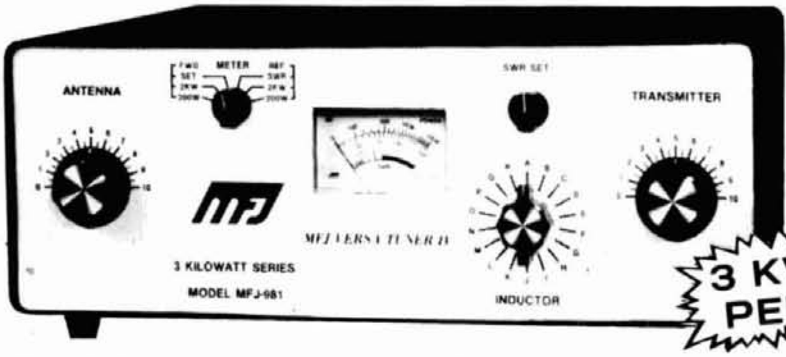

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More Details? CHECK—OFF Page 118

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volume 12, number 4

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contents

12 **40-meter receiver**
M.A. Chapman, K6SDX

23 **CW operator's PAL**
Carleton F. Maylott, W2YE

26 **calculator-aided propagation predictions**
Henry G. Elwell, N4UH

32 **deluxe memory keyer**
Robert C. Cheek, W3VT

46 **bandpass filter for RTTY**
Nathan H. Stinnette, W4AYV

48 **audio amplifier for the R-4C**
J. Robert Sherwood, WB0JGP
George B. Heidelman, K8RRH

50 **the verti-loop**
Herbert L. Bresnick, WB2IFV

56 **interesting solutions to the jammer problem**
Nuryev Sidelbandsk, UX3PU

62 **variable-frequency audio filter**
Lewis T. Fitch, W4VRV

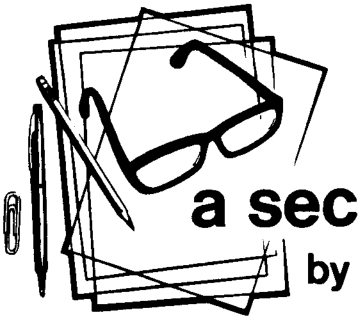
66 **tranceive and split operation with the TR-4/R-4B combination**
William P. Winter, Jr., WB8JCQ

74 **high-performance 432-MHz converter**
Charles H. Robinson, N9KD

80 **impedance measurements using an SWR bridge**
Benjamin J. Lowe, K4QF

86 **digital techniques: flip-flop internal structure**
Leonard H. Anderson

4 **a second look**
118 **advertisers index**
108 **coming events**
86 **digital techniques**
103 **flea market**
112 **ham mart**
92 **ham notebook**
6 **letters**
8 **presstop**
118 **reader service**
50 **weekender**



a second look

by Jim Fisk

It's spring and hamfest time again. It seems that no matter where you look, you'll see an announcement for yet another convention, radio auction, hamfest, or fm talk-in, and the Dayton Hamvention in Dayton, Ohio, which is billed as the *original* hamvention, is one of the biggest of the year. Scheduled for the last weekend of April, the Dayton Hamvention draws upon the large amateur population of the Midwest and has provided the model for many successful Amateur Radio conventions around the country. Several years ago, in fact, a contingent of Japanese amateurs came over to take a look at the Hamvention and its management, then returned to Japan where they staged a very popular and successful ham convention in the shadow of Mount Fujiyama.

Growing by leaps and bounds in recent years, more than 19,000 hams were in attendance last year, and even more are expected in 1979! But even with this huge influx of Amateur Radio enthusiasts, everything runs smoothly, and this year, as always, the Dayton Hamvention Committee has gone all out to ensure a lively, interesting weekend for all. Bright and early Friday, April 27th, the Amateur Radio manufacturers and distributors will start setting up their exhibits. At high noon the exhibition doors will be opened to the public — the exhibit hall will remain open until 8:30 PM, so you'll have plenty of time to browse through the many commercial exhibits. And, if this year's exhibit area is anything like those in the past, you can bet some major new Amateur Radio products will be unveiled by the manufacturers.

When the lights go out in the exhibit hall on Friday evening, they'll come on in other parts of town with banquets or gatherings scheduled for the QCWA and Old Old Timers as well as groups interested in DX, slow-scan TV, and fm repeaters (the well-known FM Bash). By early Saturday morning things will really be humming around Hara Arena, weekend home of the Hamvention, as vendors and traders from miles around start setting up shop for the famous Dayton Flea Market. Like other parts of the Hamvention, the flea market grows every year, and now takes up more than 10 acres of real estate behind the main convention hall. Whether you're interested in good used ham gear, replacement parts, vacuum tubes, new components, or antique radios, chances are good that the item you want will be for sale — your problem is to locate it among the thousands of items on display.

At 9:00 AM Saturday morning the first of many forums will be kicked off with sessions devoted to antennas, microwave techniques, space communications, microprocessors, and code proficiency. Between 1130 and 1300 there will be meetings for the DXers and Oscar enthusiasts. After lunch the emphasis will be on contests, QRP rigs, ATV, fm repeaters, vhf/uhf, and moonbounce. The traditional Saturday-night cocktail hour and banquet begins at 7:00 PM.

Early Sunday morning, just about sunrise, the Flea Market will open for another day's business; a few hours later the exhibition hall will open once again and the ARRL Forum will get under way. If you've never been to a major ham convention, the ARRL Forum offers you a chance to ask the ARRL's officers and directors questions about League affairs. The FCC Forum scheduled for Sunday afternoon gives you a similar opportunity to ask questions of FCC staff members from Washington. Both the ARRL and FCC Forums are among the most popular get-togethers at any convention, so be sure you arrive early to get a good seat.

In addition to the various sessions and forums, there will be technical and group meetings for ARPSC, OSSB, and MARS. Other special groups attending the Hamvention are the Buckeye Belles, Mid-Cars, Ten-Ten, Firebirds, Young Ladies Radio League, and others. If past performance and the 1979 schedule are any indication, this year's Dayton Hamvention will be another great show.

For amateurs who arrive in trailers and campers, parking will be permitted only in specially designated areas (no campers or travel trailers will be permitted to park in the Arena lot, including the flea market area). For those who stay in the downtown hotels, free bus service will be provided out to the Hamvention. A large allotment of rooms has been set aside for the Hamvention by the local hotels and motels; all room requests should be directed to the Accommodations Committee so that rooms can be alloted within the available supply. For more information, and a Hamvention brochure, write to the Dayton Hamvention, Post Office Box 44, Dayton, Ohio 45401.

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

Jim Fisk, W1HR
editor-in-chief

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***QST Magazine,
"Product Review",
Dec. 1978***

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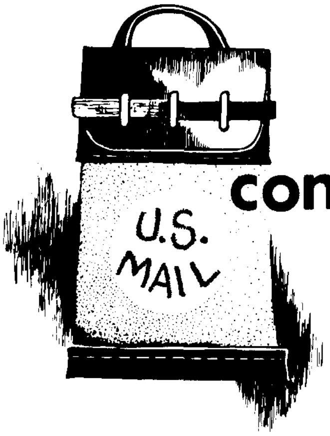

comments

Dear HR:

I've just finished reading the article on phase-locked loop demodulators in the September issue, and noticed a couple of minor errors which I'd like to comment on. In the discussion of capture and lock range, the PLL is referred to as a positive feedback system. That is not correct, as a PLL uses negative feedback (like any other closed-loop system) to reduce the error to a minimum. In this case, it is the phase error between the reference (the frequency you want to track) and the controlled variable (the VCO frequency) which is reduced by developing a signal proportional to their difference.

Another gremlin showed up in the caption for **Fig. 4** and the accompanying text, where the acquisition beat note is described as sinusoidal. It is not sinusoidal, as an examination of photographs of that signal will show. It can't be a sinusoid, because a sinusoid doesn't have a dc component. As the author points out, the output of the phase detector and low-pass filter has to have a dc component which will drive the VCO toward the reference. The most common phase detector in PLL systems is a multiplier which multiplies the input signal by a squarewave from the VCO. The multiplier produces a positive halfwave rectified output when the inputs are 90 degrees out of phase, and a negative half wave rectified output when the inputs are 180 degrees out of phase. The acquisition beat note consists of a series of these rectified sine waves which increase in period as the difference frequency decreases. The low-pass filter acts as an integrator which reduces the ripple and provides an average dc voltage which drives the VCO toward the reference. The PLL will lock when the signals are the same frequency and differ in phase by 90 degrees, which is the point where the phase detector and low-pass filter has zero dc output.

Academic nitpicking aside, I was pleased to see a nonmathematical discussion of amateur applications for the PLL. The PLL offers the possibility of some improvement in threshold sensitivity over conventional discriminators, due to its lower noise bandwidth. The noise bandwidth of the PLL is determined by the lowpass filter (approximately 3-5 kHz for an fm speech demodulator). A conventional discriminator must contend with the noise from the full i-f bandwidth. Since the i-f bandwidth for amateur fm is usually 14 kHz (and the noise bandwidth is even higher), the threshold sensitivity of a PLL demodulator should be several dB better than a conventional discriminator. I plan to run some experiments in the near future to determine the actual improvement in practice, and hope to see more discussion of this subject in *ham radio*.

John J. Murphy, K6JLF
Post Office Box 1875
Ridgecrest, California 93555

pi network design

Dear HR:

A small short circuit in my article on "Pi Network Design" (March, 1978 issue) has surfaced courtesy of correspondence with Robert F. White, W6PY. Three values at the top left-hand column on page 37 should be:

$$X_c' = -608.64 \quad C_c' = 36.572\,pF$$

In the second paragraph following the tabulation, total $C_m = 222.59$ *pF*. Total mid-point capacitance, due to improper calculation of the first (prime-valued) section, is off by exactly 5 pF.

Upon checking my original notes, I found there were two calculations for the dual-section pi network; I inadvertently used the wrong set. The equations are correct; only the calculated values are in error. W6PY found the error with his TI-59 calculator program.

Leonard Anderson
Sun Valley, California

data package for the programmable hf receiver

Dear HR:

Because of the large response I received to my programmable high-frequency receiver, which appeared in the October, 1978, issue of *ham radio,* I have put together a limited number of data packages as a help to prospective builders. These data packages include over 40 pages of circuit and wiring diagrams, mechanical diagrams, and other data; the cost is $9.75 per set, to cover costs of printing and mailing. The data package is intended as an aid in duplicating the receiver or using the same basic scheme to build a receiver to suit your readers' own needs. Printed-circuit layouts are not included, but the mechanical and wiring diagrams will be helpful in circuit-board development.

Norman J. Foot, WA9HUV
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LICENSE FEE REFUNDS came a step closer to reality with Commission adoption of a first Report and Order on its fee refund inquiry, General Docket 78-316. Only fees above $20 are to be refunded in this first phase, which will begin in April or May, after the necessary fee refund applications and instruction packets can be printed and distributed.

Commission Staff Work has already begun on Phase II of the refund program, which will cover fees of $20 or less. A Notice of Inquiry on Phase II is supposed to be ready for Commission action in 90 days.

AMATEURS WHOSE LICENSES EXPIRE now have five years instead of one in which to renew them without retaking the exam, the FCC decided at its February 27th meeting. The extended grace period applies only to the operator's license, however; station licenses can be renewed up to one year after they expire, so any Amateur who waits more than a year after expiration to renew will receive a new callsign appropriate for his license class from the group that is currently being issued.

Code Exam Credit, which could formerly be "redeemed" only at the Field Office at which the test had been taken, can now be used at any Commission Field Office.

Five Pending Petitions for Rule Making were also dismissed at the meeting. Four of them had asked for reestablishment of rules permitting Extras to select their own callsigns, while the other (RM-2360) had asked that stations operating on RTTY be permitted to identify by RTTY instead of CW as now required.

FCC's INVESTIGATION of licensing irregularities in the Indianapolis area, which began in mid-1976, is still in progress. A number of central Indiana Amateurs have received letters from the Commission asking them to explain questionable aspects of their licenses or calls.

AMATEUR TRANSMISSIONS of emergency information cannot be rebroadcast by commercial broadcast stations, the Commissioners decided on February 14th. Acting on a petition (RM-2830) filed by the National Association of Broadcasters, the Commissioners agreed to further consider whether CB emergency transmissions should be rebroadcast for public use.

ADDING A "PORTABLE" DESIGNATOR to a callsign has reportedly brought warnings to some Amateurs from FCC field engineers, but if so it's a misunderstanding of the rules. The rules change only eliminated the necessity for indicating portable or mobile status but did not prohibit the practice, which is required by some contest rules. Any Amateur cited or warned for signing portable should pursue the matter with the nearest FCC Field Office, or, if necessary, Washington.

220 MHZ OPERATORS now have a new certificate to shoot for, the "100 CCXX Award" just announced by 220 Notes. Required are 220-MHz contacts with 100 different stations. For an application for the award, which resembles similar 220 MHz incentives offered by the Northrup ARC and the 220 Club of San Diego, write WB9SNZ.

8-BAND WAS has been achieved by K5CM, with — oddly enough — the last holes being on 10 and 40 meters. After finishing up a 2-meter WAS last summer, and already having WAS on 6, Connie realized that he was within striking distance of becoming the first Amateur to earn the WAS award on eight bands. He began working on it in earnest in mid October, and within three months had all 50 cards for 160 in hand and only a few strays on other bands remaining to be cleaned up.

A NEW 2-METER DX RECORD was set February 13, when SV1DM made contact with ZS6DN in Pretoria at 1810Z over a 7117 km path. It didn't stand long, however, as SV1AB (on the north side of Athens) worked ZS6DN just three days later — February 16 — at about the same time to extend it another 10 km.

2-Meter DX is also heating up in the Americas — KP4ES, KP4Q, and KP4AAN all worked Argentina on 2-meter FM, February 19th. Prior KP4-LU contacts had all been SSB or CW.

AN "AMATEUR RADIO OPERATORS ONLY" chain letter seems to be spreading. Most copies seem to be going to DXers, and come from overseas Amateurs.

As Money Is Requested, this letter is strictly illegal under U.S. law. U.S. Amateurs are advised to ignore it.

KH0 CALLSIGNS WILL soon be showing up, as Northern Marianas Amateurs come under FCC jurisdiction. The Commission has decided to "grandfather" those who'd received their Amateur licenses from the former Trust Territory government into FCC licenses, though three KG6s who'd received their licenses as reciprocals must now reapply for an FCC reciprocal permit. Present Northern Marianas stations will be given their choice of keeping their present KG6 or stateside (if any) calls or receiving a new KH0.

Novice, QRP, 200 w, deluxe — good, better, best — $299, $369, $399, $699, $869, $899, $1069. TEN-TEC has them all. A choice of seven HF transceiver models — a choice of power levels — a choice of operating features (and accessories) for beginner or old timer. Best of all, there's a wide choice of prices to fit every amateur budget.

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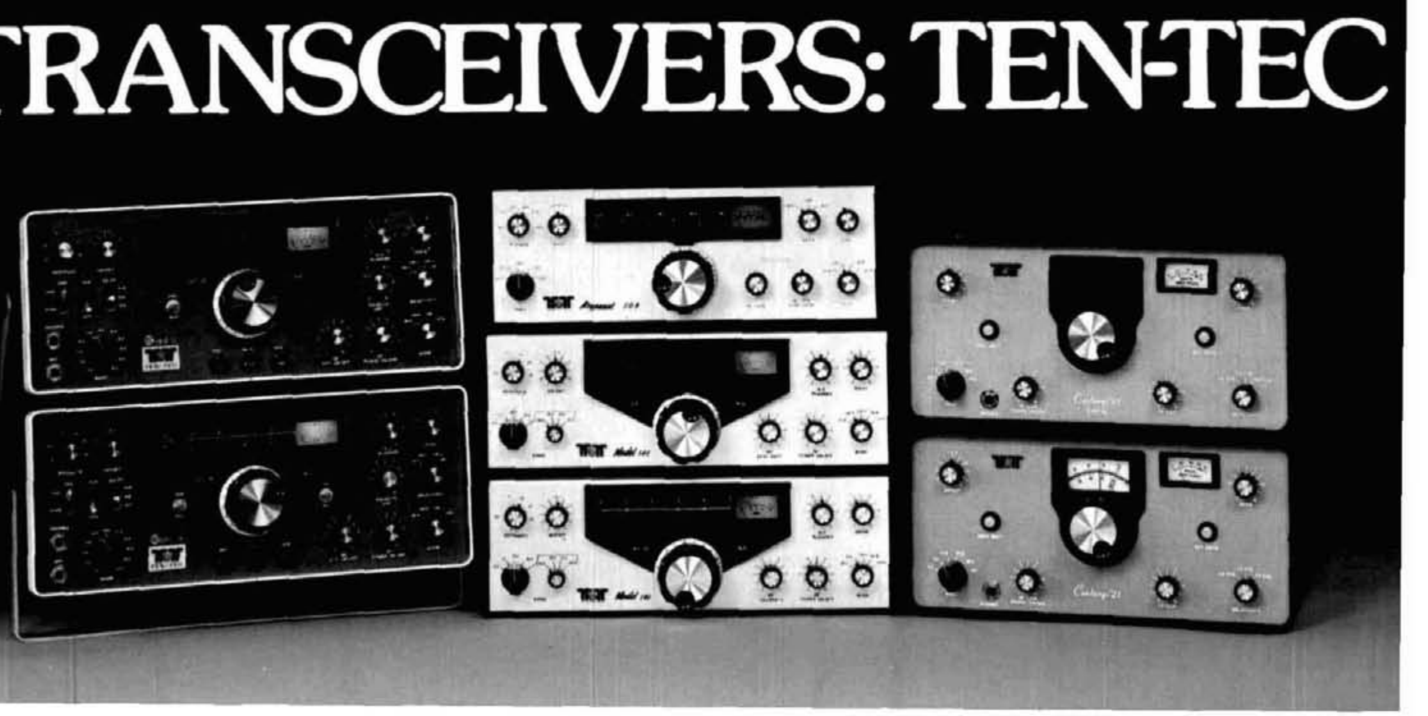
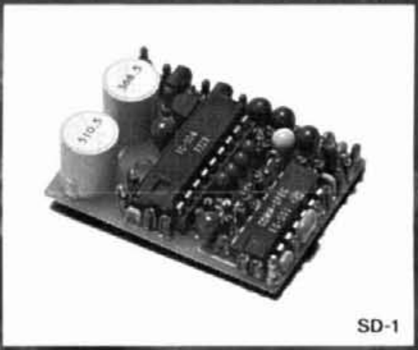



TS-1


TS-1JR


PE-2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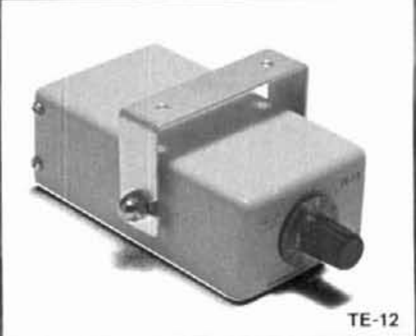
ME-3

TE-8


TE-12

ST-1

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ST-1 Burst-Tone Encoder • Measures .95" x .5" x .5" plus K-1 measurements • Frequency range is 1650 - 4200 Hz • **$29.95** with K-1 element.

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40-meter receiver

for home construction

A project for those wishing to build their own high-performance superhet — it features double conversion, all-variable-diode tuning, and digital frequency readout

This article describes an easily duplicated 40-meter receiver project. The design minimizes the use of "bells and whistles" and emphasizes performance. A variety of add-on features can be incorporated in the design.

Most Amateurs have at one time or another expressed interest in building their own receiver; however, most receiver projects fail. The reasons are usually ascribed to lack of parts, time, interest, or know-how. Many receiver projects fail simply because the builder tries for too much accessory performance before the basic receiver system is operating. That is, many hours may be spent attempting to incorporate certain special-purpose features such as a fast attack agc in a detector long before the i-f and oscillator sections are operating. The builder usually loses enthusiasm for the project by being engrossed too long in the details of getting some basic function to operate in some superhuman fashion. My recommendation is to concentrate on getting the basic system operating at least to the point where signals are heard; then real improvements and add-on features will keep the inertia of the project moving along to completion.

By M. A. Chapman, K6SDX, 935 Elmview Drive, Encinitas, California 92024

Some receiver construction projects can fail because the builder wants long-imagined, ultra-high performance and operating flexibility. To have "everything" in a receiver is impractical. A more rational approach is to think "reasonably." Compromise the first time, learn to get the basic functional blocks operating correctly, then add on the bells and incorporate your own whistles.

some ideas

Fig. 1 illustrates the functional blocks found in a superheterodyne ssb and CW receiver. One can argue the need for an rf stage; however, to reach out and snag the really tough signals and minimize spurious mixer products, the rf stage is preferable.

The biggest single problem with the superhet are image signals also available to the i-f stage. The use of moderate *Q* values in the rf and mixer tuned circuits can minimize these image effects. In some operating areas, where the image signal strength is many orders of magnitude greater than the principal tuned signal, tuned traps in the antenna feedline can reduce the image-signal amplitude to acceptable levels.

The conversion of the desired signal frequency to an intermediate frequency is the responsibility of the first mixer. Mixers are nonlinear devices, which approach a power-law transfer function; because of this nonlinearity they also send along other spurious products available from the antenna to the i-f amplifiers. These products are usually second- and third-order signals but represent other noise and image sources with which your receiver must contend.

The audio stage. Audio detection normally occurs in the second mixer. The beat oscillator provides an injected carrier for the amplified i-f signal to beat or heterodyne, resulting in a detected audio signal. The signal at the second mixer output contains, in addition to the desired audio, a wide spectrum of mixer-signal components. Most of these are in the rf range and are easily bypassed to dc ground. Some audio harmonic products exist and usually represent receiver background noise. Other noise sources are white noise from the antenna, which has an instantaneous frequency component equal to or near the audio signals of interest, and, active-device thermal noise from the receiver front end. When we speak of signal-to-noise ratio, we are usually referring to the amplitude ratio between a detected signal of say, 1 μV, with the receiver input grounded or shorted. Many commercial manufacturers also perform this test in an rf-free screened room, where discrete sources can't penetrate the chassis and only the receiver internal noise is actually measured.

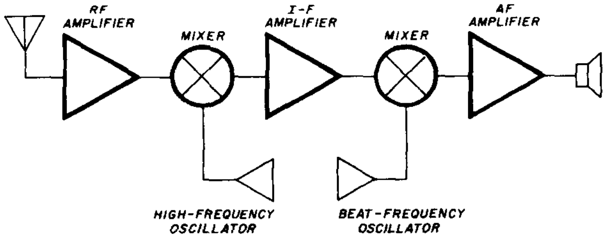


fig. 1. Block diagram of an ssb superhet receiver.

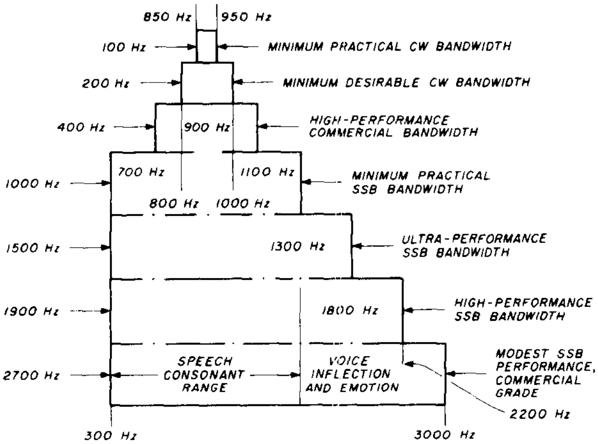


fig. 2. Summary of receiver audio bandwidth requirements for Amateur ssb and CW communications.

Restricting bandwidth. We may improve our system signal-to-noise ratio by restricting the bandwidth of detection to the smallest possible interval that will support communications. In the reception of a CW signal we're accustomed to a tone between 800 and 1000 Hz; one could visualize a system that has only a 200-Hz bandpass. Many superior receiving systems have a CW bandpass on the order of 100 Hz, centered near 900 Hz. The signal-to-noise ratios are extremely good, since only a random number of noise products fall within this bandwidth. For ssb communications, we need only a bandwidth wide enough to pass the voice consonants. Amateur receiving systems exist that employ filters limiting the audio bandwidth to approximately 1000 Hz. Various audio communications tests have shown that the average person requires audible communications bandwidths of 1500-2000 Hz. These ideas are illustrated in **fig. 2.**

If the useful audio bandwidth requirements for ssb communications are limited to 2 kHz the i-f bandpass need not exceed this interval; however, using LC-tuned circuits in the i-f or rf sections to achieve a sharp-skirted 2-kHz bandpass would require almost superhuman effort. The value of bandpass can be satisfied using crystal or mechanical filters that have extremely sharp passbands, so that infringing adjacent signals don't spill into the interval of interest.

High-frequency thermal noise in the receiver front-end active devices is normally outside the filter cutoff; however, noise originating in the detector and audio frequency sections should be considered. Most RC cutoff techniques are suitable for a-m and fm broadcast reception but don't provide sufficient attenuation of undesirable audio components for ssb and CW reception. A variety of active low, high, and bandpass circuits exist for audio processing, and most readers are at least tentatively aware of their implementation in a receiver. The *ARRL Handbook* illustrates several audio-processing designs that will provide good post-detector audio processing for optimizing ssb and CW reception.

design

We'll start with **fig. 3**, a block diagram of a 40-meter superheterodyne with all-variable-diode tuning and digital-frequency readout. There's no real difference in the diagrams of **figs. 1** and **3** — only some minor digital equipment, which allows 100-Hz tuned-frequency determination. The i-f has two separate frequencies, hence dual conversion — a desirable feature for high gain and narrow-bandpass tuning.

by a summation of signals from each of the principal oscillator sections and displayed using a five-decade LED arrangement.

Referring to **fig. 4**, signal preselection and initial receiver gain are achieved in the rf and first mixer stages, a major role in performance. We should have a good understanding of these circuits. The rf tuned circuit consisting of L1, C1, and VVC1 are parallel connected to Q1 gate 1. This is a parallel-tuned circuit. The 0.001 μF capacitor in parallel with the 620k resistor puts the bottom end of L1 at ac ground, as does the 0.05-μF capacitor on the VVC1 cathode. With a little arithmetic it's easy to see that L1 is also series resonant near 1.5 MHz, the i-f.

Parallel-tuned circuit L1, C1, VVC1 provides a high impedance to incoming signals at 7 MHz and a low impedance to signals outside the effective bandwidth.

The parallel-tuned circuit is frequency selective by providing maximum signal voltage to Q1 gate at the frequency of interest and a minimum voltage at frequencies outside the bandwidth. A similar tuned circuit L1, C2, VVC2, is on the input of Q2 gate 1. Variable-capacitance diodes VVC1 and VVC2 are ganged for peaking the incoming rf signal.

The series-tuned circuits in L1 and L2 act as image shunts for signals originating at the input. The shunting action occurs because of the low impedance to ground offered by the 1.5-MHz series-tuned circuit.

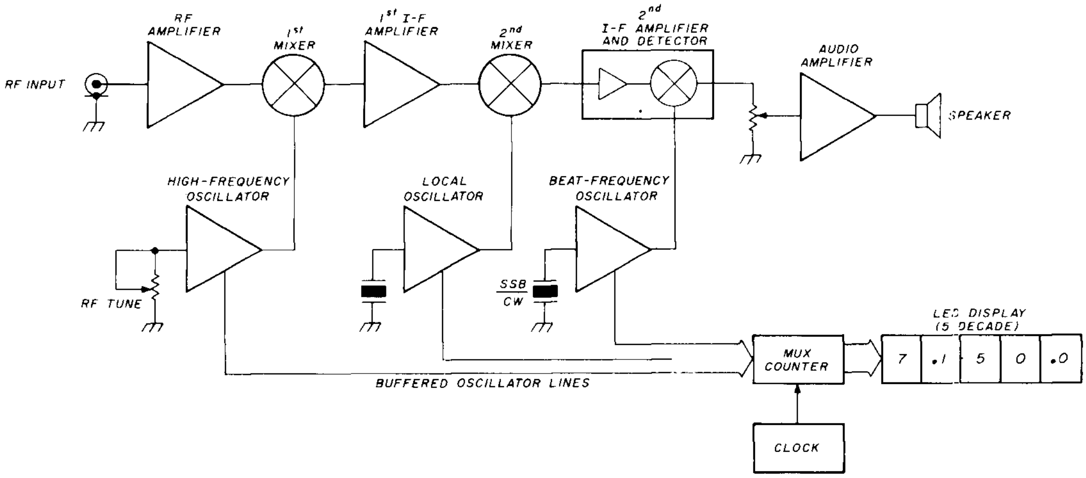


fig. 3. Block diagram of the 40-meter receiver.

Front end. The rf and first mixer tuned circuits are ganged independently from the HFO control voltage for improved signal peaking. Heterodyne oscillators are crystal controlled for stability. Over-all system bandpass shaping is achieved with a Collins mechanical filter in the second i-f. Tuning accuracy is derived

Q1 drain circuit is a simple inductive load ac coupled to Q2 input. Q2 drain load is the primary of the first i-f transformer, T1. In both cases, the drain circuits represent reasonably high ac impedances for good voltage gain. Care in the selection of the Q1 drain inductor is important, because stray winding capacitance can have the opposite effect, *i.e.*, shunting the output signal to ac ground.

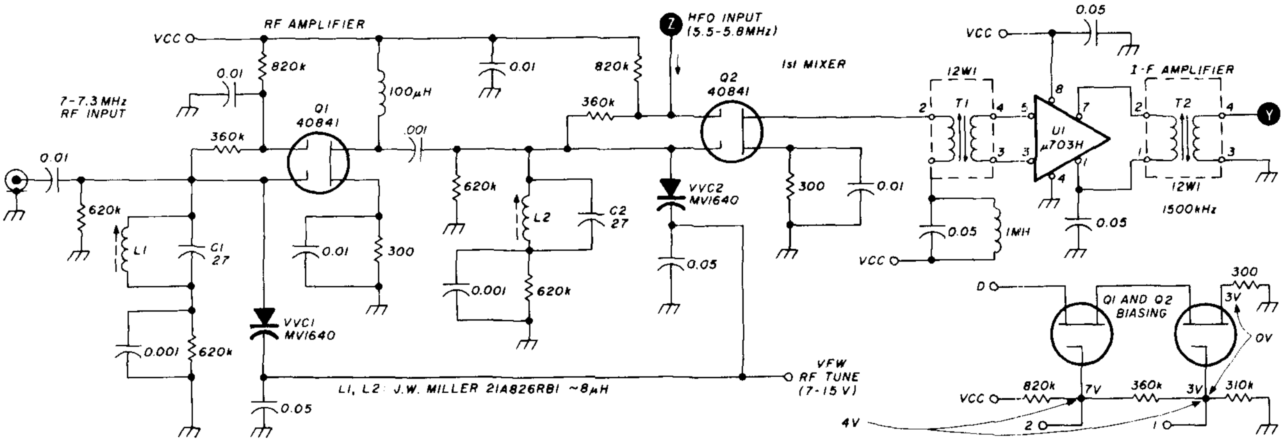


fig. 4. Rf amplifier and first-mixer schematic.

Fig. 4 illustrates the Q1 and Q1 mosfet gate bias voltages, which are set for near optimum gain even though Q2 is acting as a mixer amplifier. Because Q1 and Q2 gate currents are small, we can use high voltage-dividing resistance values for these circuits and minimize tuned circuit loading.

It may appear strange at first that the rf and first mixer are gate biased at the same point. Readers familiar with the principle of active device power-law mixing, where device biasing is set near the optimum nonlinear portion of the transfer curve, will realize the dual-gate mosfet acting as a cascode arrangement performs a modulation of the drain-circuit current for mixing, rather than a true nonlinear power-law mix. The mosfet can be biased for nonlinear mixing as well, and this is accomplished in the familiar circuit for Q6, the second mixer (see **fig. 6**).

Second i-f, detector, and BFO. Fig. 5 illustrates the i-f scheme for the second frequency conversion. The first i-f amplifier is a high-gain IC, U1, which provides about 28 dB signal gain. I-f transformers on the input and output preselect the desired signal and attenuate undesired heterodyne first mixer components. Q6, the second mixer, provides little actual system gain; the drain circuit load, T3, is the second i-f transformer. T3 includes a simple ceramic filter in the secondary for initial band shaping. This filter is inadequate for any serious ssb or CW use. The principal i-f amplification and detection occurs in U2, shown in **fig. 6**.

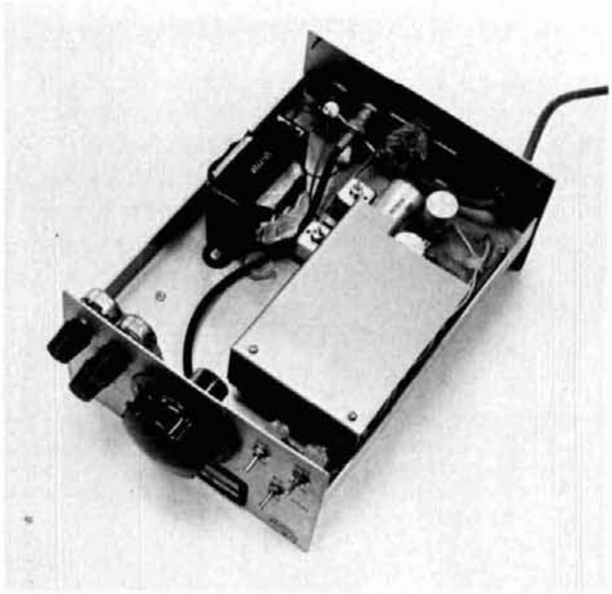
Preselected 455-kHz signals are ac coupled from T2 and internally amplified before presentation to a highly selective Collins mechanical filter for i-f bandpass shaping. I have used a Collins 1.9-kHz filter here; however, a variety of similar filters can be used, and adequate system gain is available for all but a few very exotic types whose attenuation characteristics exceed 30 dB. Nominal attenuation for the Collins unit is ≈10 dB. The selection of 1.9 kHz for bandpass shaping represents a compromise between CW and ssb signal reception. Superior ssb reception is accomplished with this filter; however, only good CW reception is possible.

Following the bandpass amplification section of U2 is a product detector agc feedback stage. Detected audio is RC filtered on the output to decouple rf components before audio amplification. The agc feedback circuit is not connected in this design but may be incorporated as an option.

Oscillator circuits. The oscillators are shown in **figs. 5, 6,** and **7**. Each of those oscillator circuits is a simple Colpitts arrangement followed by a fet bipolar isolation and shaper stage. The fet buffer minimizes oscillator loading and the effects of frequency distortion from the CE shaper stage. The high frequency oscillator (HFO) is tuned using a variable-capacitance diode, VVC3. The tuning voltage (V_{FV}) provides the range of reverse bias necessary for tuning the HFO for proper heterodyne-frequency conversion in the first mixer stage.

Audio amplifier and power supply. Detected audio amplification and dc power generation are shown in **figs. 8** and **9**. **Fig. 9** also includes the scheme for deriving the V_{FV} voltage and a method of band spreading the HFO frequency for ≈20 kHz/revolution of the front panel tuning control potentiometer.

Digital counter. Fig. 10 is the schematic of the digital counter. The clock from which the entire accuracy of our system depends is derived from the 10-MHz crystal oscillator, Q1, and multi-decade divided



Underchassis view of the 40-meter receiver.

by U1, U2, and U3. The combination of division in U1, U2, and U3 is 10^6, which provides a 10-Hz clock pulse. The *QB* and *QC* strobes are used for gate and transfer timing. U4 may be thought of as a ring counter, which further divides our 10-Hz clock into four equal time periods shown on the gate timing diagram, **fig. 11**. U5 is an AND gate to control the count periods. U6 is a large, multiple three-input AND device used to select a discrete period. The oscillator input counts and guides the signal in serial sequence through to the LED display.

We can best visualize how the digital counter works by remembering that the actual tuned frequency is a summation of the various rf stage oscillator outputs. That is, if we take a typical case, the oscillators can be summed to provide our actual tuned frequency, as in the following example:

HFO: 5.650.0 MHz
LO: 1.045.0 MHz
BFO: 0.456.0 MHz
SUM: 7.151.0 MHz (display)

Our counter works exactly as shown above. It takes each oscillator pulse train and looks at it for 100 ms, transferring the gated pulse train through the final gate, U7, in sequence. If we examine, for instance, the high-frequency oscillator at some nominal case, say 5.65 MHz, we see that there are 5,650,000 counts in one second. But since we have a gate window for the HFO of only 100 ms, then only 565,000 counts are seen by U7 in any one count cycle period. The same is true of the local oscillator and BFO counts. The total number of counts is proportional to the period established by the clock timing interval, and gated as shown in the timing diagram of **fig. 11**. The count cycle periods shown in **fig. 11** are derived from the ring counter AND circuits of U4 and U5. Each complete count cycle is 250 ms wide and has four separate parts:

1. Part 1 is count oscillator 1
2. Part 2 is count oscillator 2
3. Part 3 is count oscillator 3
4. Part 4 is transfer and display

Each of these periods is sequential so that only a serial stream of counts is seen and passed by U7. U8 serves only to delete the unwanted decade before passing the train into the LED display. U8 may be deleted for six-decade presentation if the LED selected has sufficient speed.

The transfer and clear signals are derived during the fourth time periods of our count cycle and are strobed for delayed timing by the *QB* and *QC* pulses so that there is adequate internal ripple and settling time before updating the display.

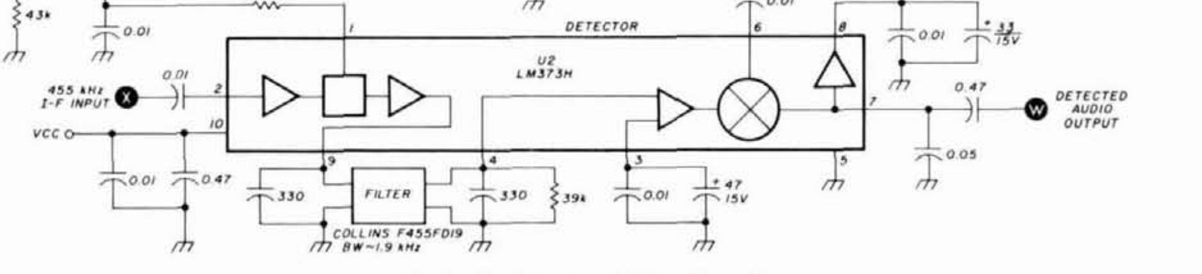


fig. 5. I-f detector and BFO schematic.

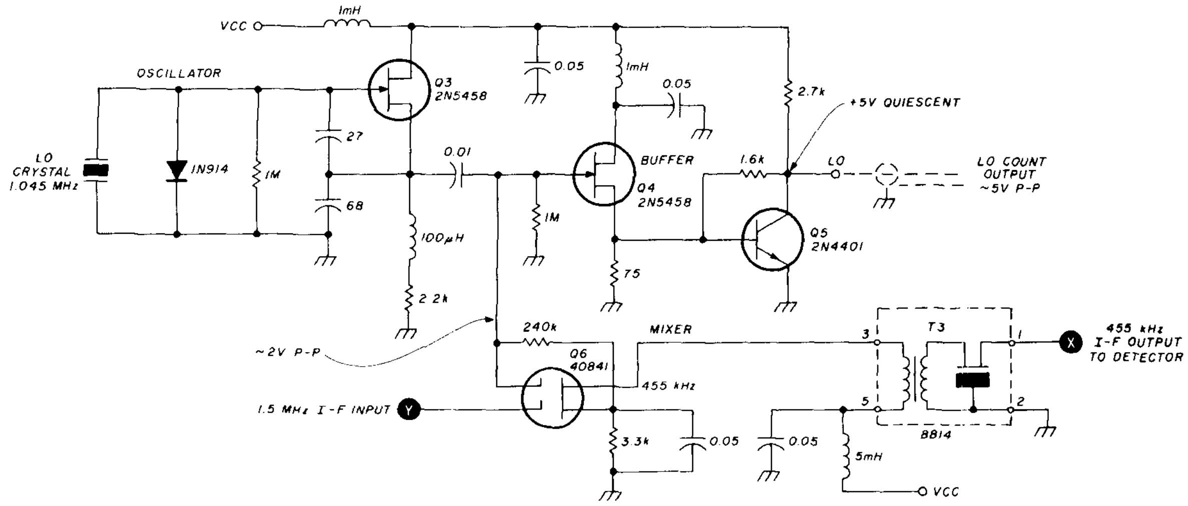


fig. 6. Local oscillator and second-mixer schematic.

U9 through U13 serve as a five-decade count, latch (transfer), and LED display. Notice that a standby switch controls the ±5 volt bus to the LED display. The purpose here is to provide power dissipation limit control but not to interrupt the continued operation of the oscillators. This will avoid the thermal shock that receiver on-off cycling would effect and minimize oscillator drift.

construction and device selection

Construction of the rf section is on one PC-board assembly.* The digital circuits are divided between two boards, the gate segments and the LED display. There's probably an ideal assembly sequence; however, being practical, one never has all the parts available at the initial starting point and we must compromise by building up various sections of the system. Any of the subfunctional blocks on a PC board can be built and tested, or the entire board assembly can be constructed and initially aligned.†

The chassis assembly shown in the photographs should in no way represent an optimum or the only way to do it; however, your chassis approach should include effective shielding if birdie and image rejection is to be achieved. An rf shield around the gate and low-frequency clock circuitry is necessary to isolate the large 10-Hz pulse from being coupled to the audio or causing harmonic interference to the rf stages.

There's no magic in device selection. With a little common sense, each of the discrete devices can be replaced with an alternative that is from a similar family type.

Consider Q1, Q2, and Q3; these are RCA 40841 dual-gate mosfet devices, and there are similar devices within the RCA family and other manufacturers' lines. The key items when considering substitution here is to become familiar with the transconductance characteristics with respect to gate 1 and gate 2 to source biasing and quiescent drain current. Most of the RCA family devices have very similar biasing. Alternatives such as the 40671 and 3N187 are almost identical in operation. Care in the substitution of units such as the Motorola MFE3006, 007, and 008 types is required, because the gate 2-to-source transconductance curve differs. There are a number of mosfet enhancement types available. You should avoid these in this arrangement because of the difficulty in obtaining sufficient reverse bias to properly operate the tuned circuit VVCs.

The 2N4401 device was standardized as the workhorse NPN. This is a general-purpose and switching unit having moderate beta at low collector currents. Any unit having an F_b $\left(\frac{f_T}{\beta}\right)$ greater than 6.25 MHz, or with similar ratings, can be used here. One needs only to put a small 5k pot in place of the present collector base feedback resistor and adjust the pot until collector saturation occurs, with the output amplitude and frequency as shown in the diagrams. For determining the alternative device feedback resistance value, remove the pot, measure its value, and use the closest standard resistor available.

The N-channel fet, 2N5458, is a general-purpose, medium-frequency device. Any reasonably similar type may be used by placing a pot in place of the source-feedback resistance and adjusting it until the output amplitudes coincide with those shown in the

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*PC boards for the receiver are available at nominal cost from the author. Send a self-addressed, stamped envelope for prices.

†A copy of the printed circuit board layouts and parts placement diagrams can be obtained by sending a self-addressed, stamped envelope to *ham radio*, Greenville, New Hampshire 03048.

diagrams. Remove the pot and install a fixed resistor with a value close to that measured on the pot. The output signal should also be examined for obvious harmonic distortion if substitutes are used, because poor mixer gain and increased receiver noise could result.

There are many VVC units that can be used provided they have similar +4 volt ratings and reasonable tuning ratios greater than two. Substitute VVCs should not be self-resonant near these operating frequencies. Alternative ICs to those shown exist in manufacturers' data books by various part numbers; however, the TI74490 is proprietary at the time of this writing. It is my understanding that other manufacturers expect to have equivalent devices on the market soon (*i.e.*, Motorola, National, and Fairchild). Type 74L, 74LS, and 74H units may be used for all or any of the gate circuit elements with slight increases or decreases in current requirements for this assembly. There are CMOS replacements for the low-frequency portions of the gate circuitry; however, their use should be examined carefully to avoid frequency degeneration, since many CMOS equivalent units are inoperable above 5 MHz in the +5 volt V_{DD} condition. Type LM374 may be used in place of the LM373 by including a 1k resistor between pin 9 and V_{CC} on U2. The LM703LH may be used to replace the LM703H provided the ac decoupling cap on pin 5 is moved to pin 7.

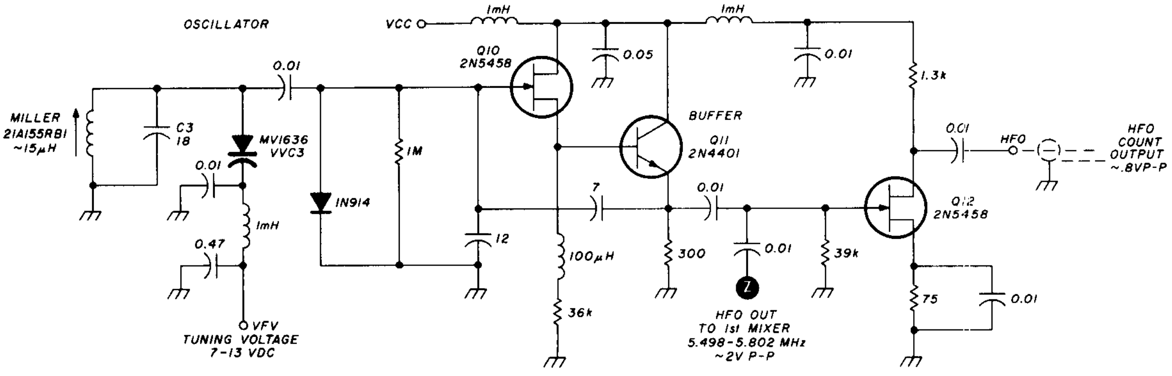

fig. 7. High-frequency-oscillator schematic.

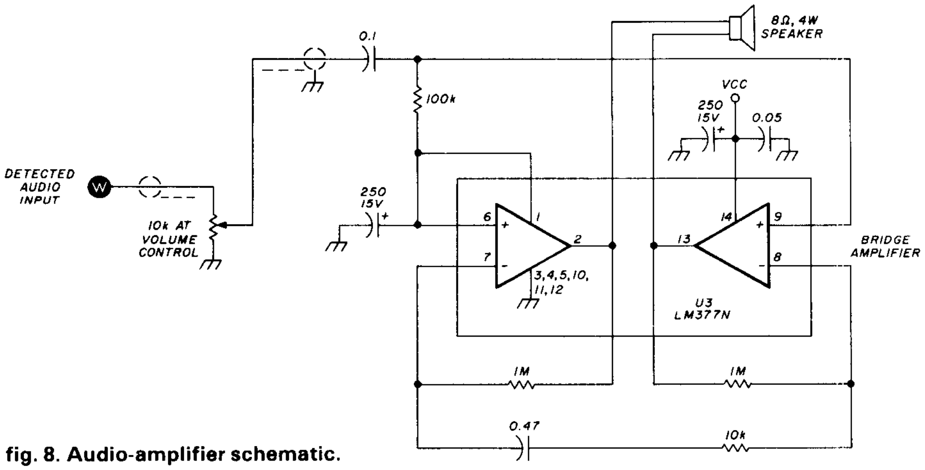

fig. 8. Audio-amplifier schematic.

alignment and test

Before alignment or assembly of the rf section to the chassis, power should be applied and a verification made of the dc voltage levels and oscillator levels in the quiescent state. Incorrect dc-voltage levels are a sure indication of probable malfunction! The dc values for all significant voltage points are shown in the schematics. These values were obtained using a high-impedance VOM and a nominal +15 volt V_{CC} voltage. Type 7815 regulators will provide outputs of ±5 per cent of nominal and may have a mild effect on the expected dc levels; but any circuit having dc values ±10 per cent from those indicated should be closely scrutinized.

test sequence

1. AC couple a 50-mV peak-to-peak 1-kHz signal through a 0.01-μF capacitor to U3 pin 7. With power applied, a 2-volt P-P signal should be apparent on pin 12 with a 10-ohm resistive load substituted for a speaker load. (All signals are referenced to ground.)

2. Adjust the BFO level on U2 pin 6 for $\approx$50 mV P-P. Adjust the i-f gain control for 4.0 volts on U2 pin 1. Ac couple a 10-mV, 455-kHz rf signal (unmodulated) through a 0.01-μF to U2 pin 2. Adjust the 455-kHz rf signal for 1 kHz detected audio on U2 pin 7. The amplitude of the detected audio should be 60 mV P-P; detuned noise should be $\approx$5 mV P-P. A verification of i-f gain control should be made by varying the voltage on U2 pin 1 from $\approx$3.9 to 4.5 volts. Maximum gain should occur at $\approx$4.0 volts.

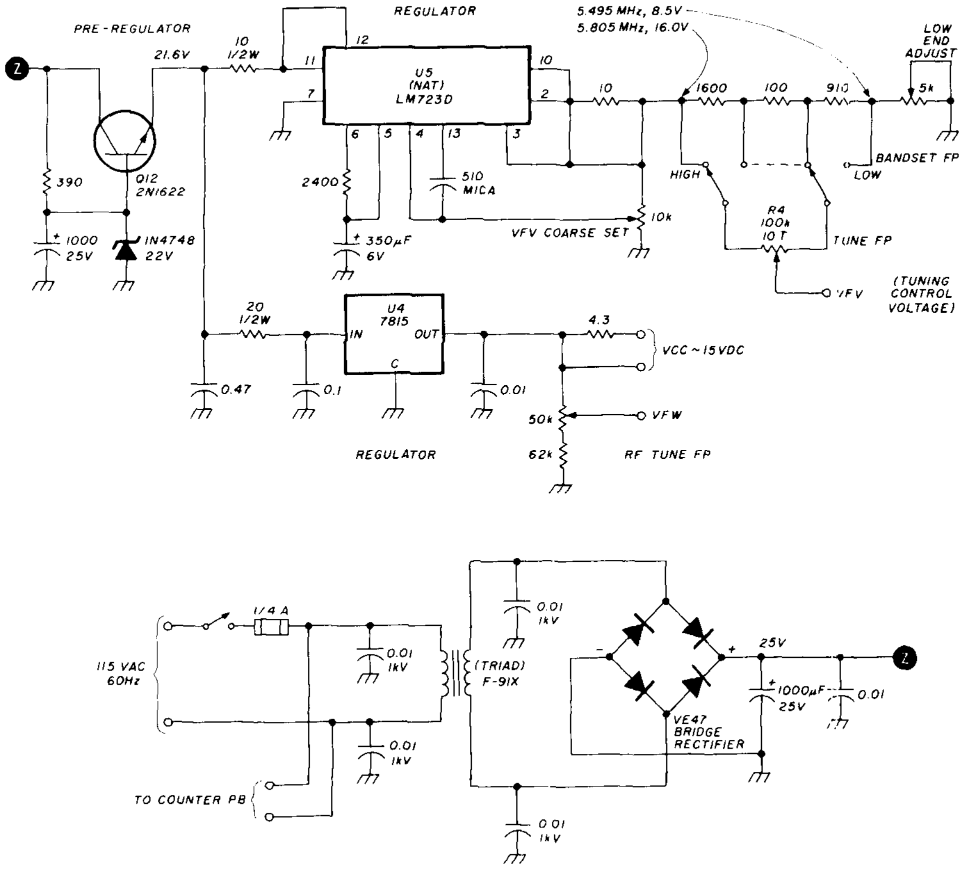

fig. 9. Power-supply schematic.

3. Ac couple a 455-kHz rf signal of 10 mV into T3 pin 3. Peak T3 for maximum detected audio at U2 pin 7.

4. Ac couple a 1.5-MHz rf signal of 100 μV into T1 pin 2 and peak T1, T2, and T3 for maximum. Detected audio signal on U2 pin 7 should be approximately 200 mV P-P. The signal may also be monitored on U2 input where a 455-kHz signal appears and T1, T2, and T3 adjusted for maximum.

5. The HFO V_{FV} fine band-edge voltage levels should be set as shown in the schematic. With the tuning potentiometer set at the indicated levels, or using an external supply, apply $\approx$11 volts to VVC3 cathode. With V_{CC} applied, adjust L3 slug until a 5.65-MHz signal is available on Q2 gate 2. Adjust the source swamping resistor (30k) for the output amplitude shown in **fig. 7**. Traverse the tuning range and alternate the bandset switch so that the output signal frequency is 5.498 MHz at the low end and 5.802 MHz at the high end of the HFO. Monitoring should occur at Q2 gate 2.

6. Ac couple a 7.15-MHz signal of 100 μV into Q2 gate 1 and monitor the audio output while adjusting the HFO for detector conversion. After obtaining detected audio from U2 or U3, with the HFO set for conversion of 7.15-MHz signals, repeak T1, T2, and T3 for maximum. Monitoring the rf envelope on U2 pin 2 may provide an easier point for verification of optimum alignment.

7. Ac couple a 30-μV signal of 7.15 MHz into Q1 gate 1. Adjust the HFO frequency for a maximum 455-kHz envelope. Adjust the rf tuning pot for maximum-detected audio or rf envelope. Inject a 5-μV signal into the antenna input, and peak L1 and L2 for maximum detected signal; then go back through all tuning elements and peak for maximum. The peak detected rf envelope at U2, for a 1-μV input of 7.15 MHz, should be 60 mV P-P (≈95 dB).

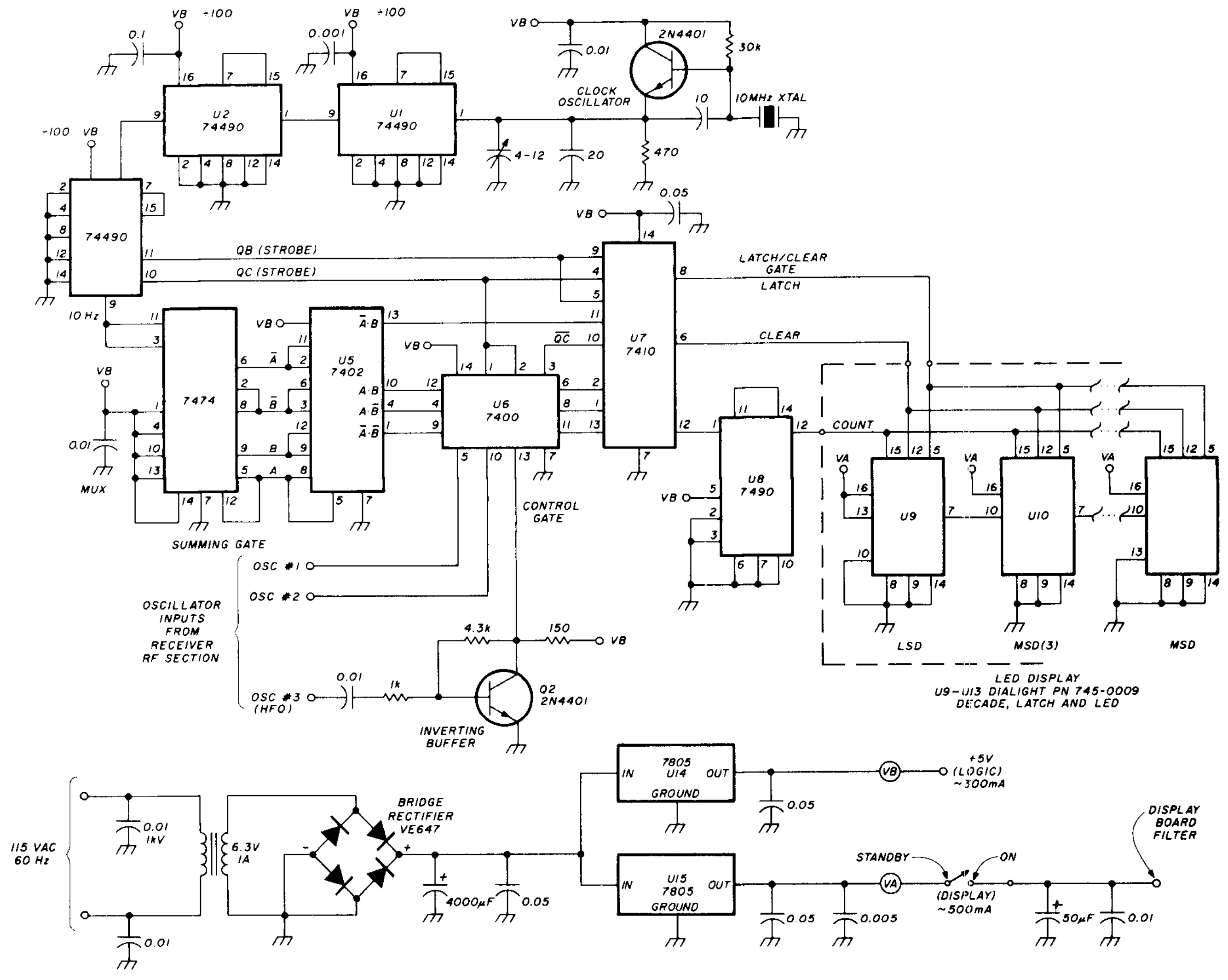

fig. 10. Digital-counter and display schematic.

8. Final alignment should occur after the rf and digital boards have been installed on the chassis, because the effects of stray capacitance will shift all of the oscillators and rf-tuned circuits slightly. This final alignment should occur with rf input signals in the order of 3 μV.

digital-circuit testing

1. Apply power to the gate control board and monitor the frequency of the quiescent states of the logic circuits, as shown in **fig. 10.** It may not be possible to verify the various timing pulses relative to each other as shown in **fig. 11**; however, scope monitoring of the logic pins should indicate that a pulse exists.

2. Inject a 1-MHz rf signal of 100 mV into oscillator no. 3 input, and monitor U7 pin 12 and U8 for oscillator clock train output occurring at regular 250-ms intervals. If the display board is available, temporarily jumper the latch and clear lines. A display of 1.000.0 should result. Inverted latch and clear lines, or unconnected inputs to the display, will result in LED garbage, or an output of 0.000.0. On occasion, when power is first applied to the display, a residual numerical value will appear. In normal operation the next count cycle coming up (250 ms later) should clear the erroneous display.

3. The 10-MHz clock should be trimmed for the closest possible accuracy by zero beating against WWV at 10 MHz, using a separate receiver and a simple link coupling around the clock oscillator.

hints and kinks

Accuracy in a digital system is defined as time base accuracy ± 1 digit. This accuracy must include aging and temperature stability. The digital-counter accuracy will then be as accurate as the basic 10-MHz clock oscillator. All crystal oscillators have both long- and short-term drifts due either to aging, temperature, or circuit-component changes. The display accuracy then will drift with the crystal-oscillator drift.

Without temperature compensation you can expect to have drifts in the order of 5 ppm/°C plus additional aging drifts of perhaps 10 ppm/year. The net result is that faith in the absolute accuracy of the LSD of this counter is wasted. It might impress visitors, but without constant recalibration of the 10-MHz oscillators against WWV or other prime standard at regular intervals, the LSD is only a guess. The actual consistent counter display accuracy is to the closest kilohertz over any significant period of time. This could be improved to ±200 Hz ±1 digit by using TCO (temperature-compensated oscillator) techniques.

Selection of the BFO crystals can be made by measuring the bandpass mechanical filter for the upper and lower 20-dB points. The BFO crystal frequency should coincide with these measured values within ~50 Hz. Most suppliers will furnish the filter with this information, marked either on the filter or with the shipping data package, for a small fee. Significant errors in the BFO frequency will degrade ssb detection.

In the local oscillator and BFO circuits, the amplitude of the output signals from the isolation and CE shaper will vary with different crystal types. Excessive CE saturation will cause a 10-Hz clock pulse to be coupled into the V_{CC} line and will result in pinging of the detected audio. Conversely, low outputs will not stimulate the gate levels required for proper counting.

The suggested cure is to temporarily install a 5k pot in the 2N4401 collector to V_{CC} load position. Adjust each of these until the amplitude of the output signal is just high enough to stimulate the gate input but will not keep the collector circuit in saturation for a period that allows the 10-Hz signal to load the V_{CC} line. The schematic illustrates the peak-to-peak signal levels I found achieved these goals.

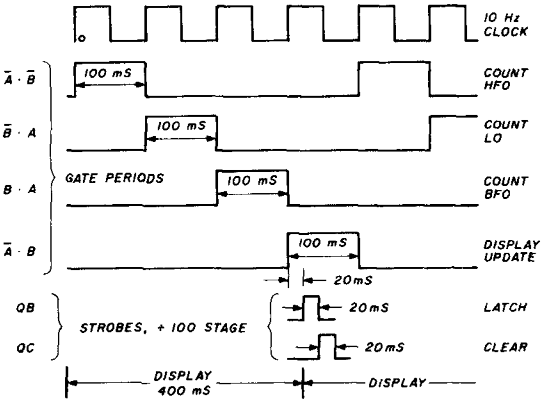

fig. 11. Digital-counter timing diagram.

A 100-kHz clock oscillating crystal may be used as a substitute for the 10-MHz unit in **fig. 11** by deleting the second divide-by-100 (74490) stage. Some adjustment in the oscillator bias values may be required, and counter accuracy will be limited by your ability to trim the oscillator frequency properly.

improvement suggestions

The following are areas of individual user improvement:

1. RF gain. Wide dynamic range is necessary to meet the demands of the Amateur-band receiving environment. Signal levels vary from parts of a microvolt to perhaps a millivolt — a three-decade range. The front end of this receiver is optimized for maximum transconductance (gain). Strong signals $>100\ \mu V$ will cause channel saturation in the rf- and first-mixer stages. Two easily implemented methods of controlled rf gain are available to the builder:

a. source current limiting: use a 2k pot in the source lead instead of a 300-ohm resistor.

b. gate-2 control: lift the gate from the board and add a bypassed flying lead to a 0-5 Vdc voltage.

2. Agc. Positive agc voltage is available on U2 pin 8, proportional to detected audio. The quiescent value, ~4.0 Vdc, is very close to the i-f gain maximum value. As the audio level from the U2 audio mixer stage increases, this agc voltage may be fed through a 1-k resistor to U2 pin 1.

Agc voltage may be fed through a 1-k resistor to U2 pin 1. The i-f gain circuit must be open or disconnected for the agc feedback system to operate.

3. Variable bandwidths. The addition of a diode switch and other Collins-type filters will allow optimization for both CW and ssb reception.

4. HFO stiffness. Thermal — physical isolation of U5 will improve V_{FV} regulation and drift/°C rate; transient — increased beta characteristics for Q12 will aid in improved ripple rejection for both the HFO and general receiver operation.

5. Images. A simple series-tuned 1.5-MHz shunt in the antenna input will improve the image rejection by 20 dB. Provisions are on the PC board for the installation of the components.

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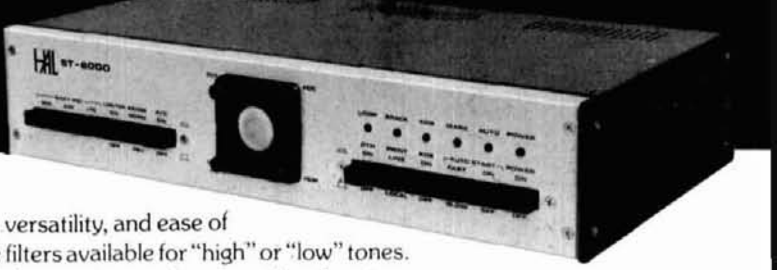

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CW operator's PAL

An accessory for CW operators that reduces difficulties with interference, noise, and fading

Described in this article is a multifunction magic box, called a *Pal*, which I designed to relieve some of the operating problems faced by CW operators. Some of those problems are outlined below. The *Pal* is based upon old, well-known techniques; it is novel primarily because it merges — and modernizes — past solutions to these problems:

1. Many Amateurs endure listening to CW signals but enjoy listening to telegraph sounders. Most Amateurs own or can obtain sounders, which are now obsolete, and old-timers usually retain a nostalgic interest in telegraphy.

2. Received signals vary in pitch, due to frequency drift at the transmitter or receiver, which requires frequent readjustment of the receiver's BFO. Chirps due to poor transmitter power-supply regulation, or jumps due to sudden line voltage changes, can also cause problems for the CW operator.

3. Received signals vary in volume due to fading and due to variable propagation conditions; sometimes gain changes at the transmitter or receiver can occur during warm-up periods. Changing audio volume forces the receiving operator to frequently readjust the gain control to maintain normal volume.

4. Received signals often include annoying interference from adjacent frequencies, even with very selective receivers.

5. Received signals incur variable background noise due to a combination of atmospheric, static, and manmade electrical noise; noise limiters help but don't eliminate all noise.

6. Transmitter operators are said to have poor "fists" when they send imperfect code characters with straight or semi-automatic keys. Poor sending can often be attributed to inability to monitor a transmitter not equipped with sidetone output.

7. Many Amateurs don't own a code-practice set. Such a set is more than a means of learning and teaching code; it permits high-speed, off-the-air sending practice and proper adjustment of a semiautomatic key.

circuit description

In the *Pal* (**fig. 1**), a rectifier-filter circuit is used to actuate a telegraph sounder; a tone oscillator-amplifier is used to provide a code-practice set. Both circuits are used to improve CW reception and transmission. The input voltage to drive the *Pal* is derived from either the receiver (J1) or a voltage picked off the transmitter (J2).

The external power supply may be a 12-volt storage battery, eight dry cells in series, a regulated 12-volt power supply (such as Radio Shack's no. 22-124), or a homemade unregulated power supply which delivers 12-15 volts at about 1 ampere. An internal 9-volt transistor radio battery of moderate size will suffice for code practice.

An ac-connected power supply should have a switch and an indicator lamp — either a small 120-volt type across the switched line or a 6.3-volt, 0.15-ampere pilot lamp in series with a 40-50 ohm, 2-watt resistor across the 12-volt output. Lamp types 40 (screw base) and 47 (bayonet base) are suitable.

construction

Nothing in this circuit is critical. The circuits may be breadboarded and simply placed in a box. A 5 × 5 × 7 inch (12 × 12 × 18 cm) metal box is recommended because it will easily house all the parts except for the sounder (on top) and the key.

By Carleton F. Maylott, W2YE, 279 Cadman Drive, Williamsville, New York 14221

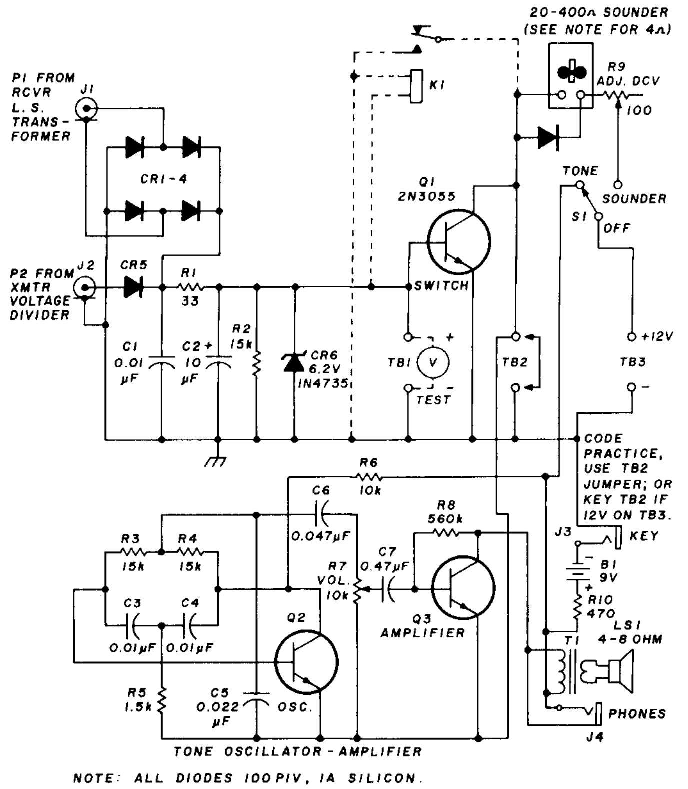

fig. 1. Schematic diagram of the CW operator's Pal. Q1 is preferred and relay K1 (Radio Shack 275-004) is optional because Q1 is more sensitive; it needs less input and perhaps no step-up transformer if fed directly from a high-impedance phone jack instead of a voice-coil input. Relay K1 can be used to operate a 4-ohm sounder (1.5 Volts, 375 mA) by replacing the latter in the given circuit, setting R9 to 500 ohms (or using a 560-ohm, 1-watt fixed resistor), and putting K1 contacts in series with the sounder and a large dry cell or a 27-ohm 10-watt resistor in the 12-volt line. Transistor Q2 is an NPN low-power device.

sounder operation

Any telegraph sounder with 20 to 400 ohms resistance is usable, but the power supply voltage must be dropped with a series resistance. For example, if a 20-ohm sounder operates at 3 volts and 0.15 ampere, the necessary 9-volt drop requires 60 ohms resistance at 1.35 watt, so a 56-ohm, 2-watt resistor will do. Obviously, if various sounders are to be used, an adjustable resistor or rheostat must be used; a common 12½-watt, 100- or 500-ohm rheostat is a good choice.

A communications receiver usually delivers only a fraction of a volt to a 4-8 ohm loudspeaker. The CW-dc converter must have a higher ac input voltage to deliver about 0.8 volt to a switching transistor or 5 volts to a sensitive relay. The dc output voltage of a full-wave bridge rectifier is, theoretically, 90 per cent of the input ac voltage. Under these conditions, it is necessary to use a step-up transformer unless the output from a high-impedance phone jack is sufficient. The step-up transformer can be made from a filament or output transformer used backwards and shunted across the voice coil of the speaker.

A switch in series with the voice coil will disable the loudspeaker while the *Pal* is in use, thus avoiding confusion between CW tones and sounder clicks. Before disabling the loudspeaker, advance the receiver gain control to a point somewhat beyond the triggering level. A lower level invites drop-out during fading, and a higher level invites atmospheric static and manmade noise, which seldom affects the sounder under normal operating conditions.

The pitch and volume of the original CW note are of little importance during sounder operation; in fact, the note may be much higher or lower than desired during normal CW operation. Since pitch is now unimportant, you may zero-beat or drop the frequency of an undesired signal so that only the desired signal passes efficiently through the receiver-to-converter transformer. If a phone signal happens to be received, the sounder will faithfully click on voice peaks; no harm is done, and the effect is rather amusing.

tone-oscillator operation

A center-off switch (to avoid needless energy loss) is used to replace the sounder with a tone oscillator, amplifier, and loudspeaker. The converter circuit is unchanged and the operating procedure is much the same. The pitch of the new CW tone is fixed by the circuit constants; the volume is fixed after a preliminary setting of the gain control. Pitch and volume are now unrelated to receiver output; they should never need readjustment regardless of fading, interference, or noise.

During tuneup, there are two tones because there are two loudspeakers with different inputs. Thereafter, confusion between the fixed and variable tones may be avoided by opening the receiver loudspeaker disabling switch.

transmitter operation

The sidetone for transmitter monitoring is produced in much the same way as in the CW-to-CW conversion just described. A separate half-wave rectifier is connected to a voltage divider shunted across the transmitter output. The half-wave rectifier and the bridge rectifier share a common output and their back-to-back polarity connection avoids interaction unless both inputs are active. Fortunately, transmitter and receiver functions do not occur simultaneously, hence the *Pal* can serve both functions at the same time without changing connections. Thus, both plugs can remain in both jacks (phono and coaxial) regardless of the choice or number of functions desired.

It is easy to obtain the rf voltage needed for the half-wave rectifier input; a typical transmitter which delivers 200 watts to a 50-ohm load also delivers 2 amperes at 100 volts. Even a 50-watt transmitter can deliver 50 volts, of which only a small fraction is required. The proper voltage divider step-down ratio is, therefore, rather large. It is determined easily by experiment but not by calculation. It is useless to know that a half-wave rectifier, in the absence of voltage drop, has a dc output voltage which is 45 per cent of the ac input voltage. Variables include transmitter power, voltage divider resistance, rectifier load, rectifier voltage drop, and rectifier circuitry.

Under these conditions, you can connect the transmitter to a dummy antenna and gradually raise the tap on the voltage divider until the desired dc voltage appears across the relay coil or the switching transistor base-to-emitter input. The dc voltage should not exceed 6 volts or 1 volt, respectively[1].

The voltage divider should be located in or near the transmitter, if possible, to minimize any effect on the VSWR. The divider might be housed in a can, like a filter, in series with a line from the transmitter to the *Pal*. In any case, a short jumper made from RG-8/U or RG-58/U coaxial cable may be provided with PL-259 plugs for attachment to SO-239 jacks at both ends. The transmitter antenna jack may be provided with a T-adapter which has one male and two female outlets (amphenol M-358), one of which will accommodate the jumper.

code-practice operation

There are at least three possible ways of keying the *Pal* as a code-practice oscillator. The key may be connected across the normally open relay contacts or the collector-to-emitter leads of the switching transistor. If you wish, a small battery may be placed in series with the key to the phone jack used for normal receiver connection, making sure that the plug tip is positive.

A third way of keying the practice oscillator replaces the external 12-volt power supply with an internal 9-volt transistor battery in series with a key jack, thus providing greater portability.

conclusion

There you have it, a multifunction magic CW box, which is a *Pal* to me; other amateurs should be able to obtain similar results. I wish to thank W2SSJ and his correspondents, W9KSR and W9YZE, for their suggestions on CW-to-telegraph converters.

reference

1. Lew McCoy, "An RF Actuated CW Monitor," *QST*, November, 1968, page 39, or *The Radio Amateur's Handbook*, 1971, page 183.

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Those people who use high-frequency propagation predictions know them to be accurate. Commercial interests use the predictions as part of the normal course of business. DX contest operators use them, among other things, to determine band openings into high-density population areas to establish a high contact per hour ratio before skip lengthens into the rest of their country. DXCC operators use them to determine at what time and on what band a specific long-haul station will be coming through so as not to waste time when no path is possible.

Propagation predictions became practical after World War II through the work of the Central Radio Propagation Laboratory of the National Bureau of Standards. In 1947, Newell Atwood, W3KTR, published an article[1] describing a method whereby CRPL propagation prediction overlay charts were moved over a basic world map, showing what areas of the world were open from a home station for any band and time of day. That information was issued by CRPL three months in advance and proved very useful, as well as accurate.

Much has been written over the years on the subject. At least one magazine[2] devoted a complete issue to the subject, while another[3] devoted a monthly article for many months to prediction techniques and their results. Finally the ultimate was reached in 1971 when *Telecommunications Research and Engineering Report 13* was issued by the United States Department of Commerce.[4] The report consists of four volumes on ionospheric predictions covering all frequencies, time of day, and day of the year, and by interpolation, for any solar activity with a Zurich sunspot number from 10 to 160.

The use of these predictions is explained in Volume 1, but a detailed description was published by Jerry Hall, K1PLP.[5] Every Amateur interested in DX should have a copy of these four volumes in his library. The use of these predictions is tedious and time consuming, but well worth the effort. First, let me sell you on their importance by illustrating how you can use the prediction techniques to get that long-haul DXpedition when others can't. Then I will explain how I've streamlined the use of them, and made it more fun by use of an SR52 computing calculator with the PC100A printer.

The programs discussed in this article are directly related to the TI-58/59, newer models of the SR52/SR56; by use of the equations and flow charts, they are applicable to the HP-25 and similar calculators. The PC100A printer is a desirable luxury, but is not necessary to the functioning of these programs.

Back in December, 1975, the Northern California DX Foundation sponsored a DXpedition to CR9AK in Macao, a country I needed. Macao is a long way from New Jersey; since I work from 8:15 AM to 5 PM, I had to know when I could expect to work him — to increase the possibility of being there when his signals were coming through.

The sunspot cycle had not reached bottom in 1975, but it was pretty low, around a Zurich sunspot number of 30. Under those conditions, when could CR9AK be workable in New Jersey, and on what band? A propagation prediction using *Report 13* was performed (see **fig. 1**) and, much to my dismay, there was no opening above 13.8 MHz via the short path. Happily, though, there were two openings via long path, one at 2400 GMT (2400Z) with an MUF of

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14.4 MHz and a second one from 1100Z to almost 1600Z with an MUF peaking to 21.7 MHz.

Unfortunately, that particular week of the DXpedition found the ionosphere very unsettled and nothing was heard on either opening the first three days of the DXpedition; the beam was pointed long path, while I listened on 20-meter CW. Little was heard from anyplace that week. On the fourth day, a very weak pileup was heard around 14025 kHz; it had to be CR9AK. When you are desperate you have to try, so a short call was given after the pileup stood by; he came back to me with a "surprised to hear East Coast." We exchanged signal reports, I was 459, and back to the faint (in New Jersey) pileup he went. Although CR9AK made amost 4000 contacts in 71 countries, few of them were with the East Coast. My working him was part luck, but also planning, as finding him would have been hit and miss without the prediction.

In April, 1976, Bill Rindone, WB7ABK, in cooperation with the NCDX Foundation, visited Christmas Island in the Indian Ocean, operating as VK9XX. Christmas Island is 3700 km farther from New Jersey than Macao. In addition, the Zurich sunspot number was lower than it had been in December — so low that I used 10 to allow me to use *Report 13*, Volume 2 directly. Again, when should I listen for him and on what band? A propagation prediction was made, (see **fig. 2**) and I had several choices, both short and long path on 20 meters. I'll take the short path anytime I can, as the distance is about 18,300 vs 26,000 km (9900 vs 14,000 miles) for the short and long paths respectively to Christmas Island. With a TH6DXX antenna at 15.2 meters (50 feet), the signal makes quite a few hops at even 18,300 km (9900 miles).

From **fig. 2**, you can see that the opening at 2400Z is sharply peaked. I have not been very successful with that type of opening and concentrated on the opening at 1200Z. This worked out quite well — I got VK9XX shortly after the band opened on 20-meter ssb before the pack became aware of him. I left for work only 20 minutes after my normal departure time.

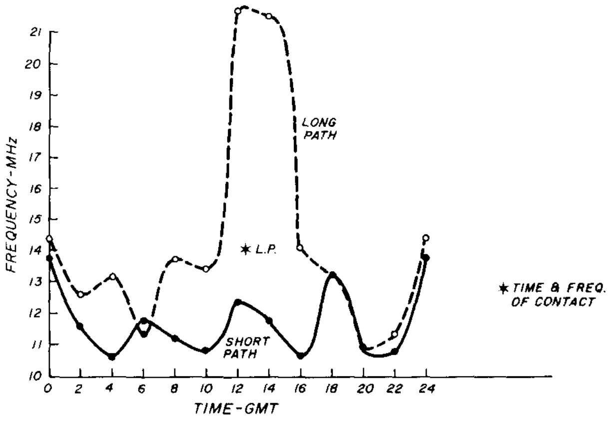


fig. 1. Propagation prediction for the long and short paths to Macao. The prediction showed that for December, 1975, there was no 20-meter short-path opening. However, the long path was open, even as high as 15 meters.

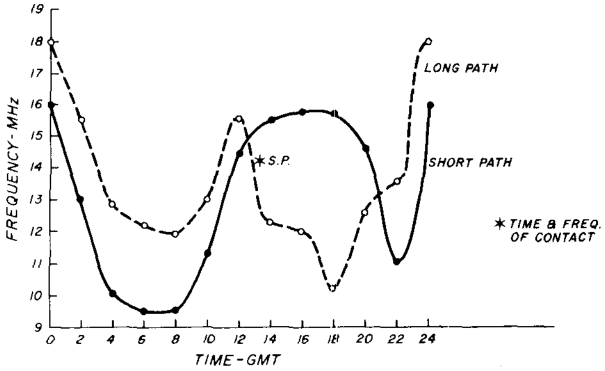


fig. 2. A similar propagation prediction for the path from New Jersey to Christmas Island in April of 1976.

I know that many top DXers rely only on their experience, patience, and operating skill to snag elusive DX. But, when it comes to a DXpedition, you know he's going to be continually working. So, why not take advantage of all the miracles of modern science to get him!

Propagation predictions using *Report 13* and reference 5 are easy to do, but they take a long time. You must first draw on transparent paper a great circle route between your home and the desired DX location and then establish an ionospheric control area distance at both ends of the great circle. The great circle is then over-layed on the MUF (Zero) F_2 map for the solar activity level, month and time desired, and the MUFs at the two recorded locations. The process is repeated using the MUF (4000) F_2.

The F_2 MUF is then determined by means of a chart relating the (zero) and (4000) MUFs, which tends to decrease the MUF for all distances greater than 4000 kilometers. Next, the process is repeated for the MUF (2000) E, as some propagation may be by the E layer as well as the F_2 layer. Actually, the MUF (2000) E exercise is mainly for distances less than 2500 kilometers. By cranking the numbers and filling in the recommended work sheets, you arrive at an MUF for every two hours during the day. A plot such as **fig. 1** or **2** can then be drawn and analyzed. However, if the Zurich sunspot number is not 10, 110, or 160, which are Volumes 2, 3, and 4, the whole thing must be repeated and an interpolation made for the desired sunspot number. It's easy, it's all explained, but it is tedious. I'm sure the dogwork inhibits many from using the technique. Let me repeat: it's well worth it to give you a competitive advantage against the big guns.

Hang in there, though, if you're lazy like me. Pare the work to the bone. Being hams, we are highly op-

```
LATITUDE           22.0      PRT
LONGITUDE        -115.0      PRT

NAUTICAL MILES   7219.2      PRT
BEARING           343.1      PRT

KILOMETERS      13369.9      PRT
```

fig. 3. Example of the printout for the great-circle distance (both in nautical miles and kilometers) and bearing to the desired DX station.

timistic when it comes to band openings. Why not ignore the MUF (Zero) F_2, which tends to decrease the MUF anyhow, forget the MUF (2000) E, unless you're looking for short haul, and record only the lower MUF of the two control areas, since that's the one you are going to use anyhow? That saves a lot of work. Unfortunately, you still have to go through the interpolation procedure until the Zurich sunspot number hits 110, when you can use Volume 3 directly. That won't last long though, as the number will go either up or down.

making the prediction

The first thing you need for the prediction is the great circle distance (in kilometers) between your home station and DX location. *Report 13* shows you how to do it graphically, but why not use the calculator? Other sources of theory include the articles by Marquart[6] and Hall[7]. If you have a Hewlett-Packard HP-55, Chester Brent, WB4GVE, has the program all worked out for you[8].

The program for the SR52 provides the beam heading and distance in nautical and statute miles, as well as kilometers. (The kilometer value is automatically stored in memory 98 of the SR-52, a storage location not cleared except by removal of power.) The program is contained on a magnetic card, requiring only the insertion of the DX latitude and longitude, which are also printed out on the PC100A for reference purposes. The home location is part of the program, although a second location may be manually inserted. The equations for D, distance in nautical miles, and H, heading in degrees, are:

$$D = 60 \cos^{-1}[\sin L_1 \sin L_2 + \cos L_1 \cos L_2 \cos(\lambda_2 - \lambda_1)]$$

$$H = \cos^{-1} \frac{\sin L_2 - \sin L_1 \cos(D/60)}{\sin(D/60) \cos L_1}$$

where

L_1 and λ_1 are your respective latitude and longitude

and

L_2 and λ_2 are the respective latitude and longitude of the other station

When the problem is solved, bearing from true north, distance in nautical miles, and distance in kilometers are displayed and automatically printed on the PC100A tape. In addition, by pushing a button, statute miles may be displayed. **Fig. 3** shows the printout obtained.

Since 200 of the 222 storage locations were used for the bearing and distance program,* another program is required for the prediction. The prediction program, also on a magnetic card, is inserted in the SR-52 without de-energizing the unit; it is necessary to retain the kilometer value stored in memory location 98. The program is then actuated and the PC100A prints out the number of hops required for the path, the reference hop length, and the ionosphere control area distance (see **fig. 4**). These equations are very simple:

Number of hops = Great circle distance/4000,

but increased to the next higher integer; *i.e.*, if the answer is 3.3 hops, increase it to 4 hops.

The flow chart shown in **fig. 5** indicates the steps taken by the calculator to arrive at a whole integer for the number of hops.

Reference hop length =
Great circle distance/number of hops

Ionospheric control area distance =
Reference hop length/2

At this point the smoothed sunspot number is placed into the computer, because from now until 1990, we are going to have to interpolate. There are two buttons to be pushed on the SR-52 while extracting the data from prediction volumes. One will be called Label A for MUFs at the home location, and one Label B for MUFs at the DX location. Open two volumes to the same month, 2400Z and MUF (4000) F_2. The two volumes cover the sunspot numbers which are to be used for interpolation to the desired sunspot number.

The great-circle overlay is placed on the lower sunspot number chart first, carefully aligning the equator and the vertical line reference. I use a Greenwich meridien vertical line as well as a vertical line on the right border of the charts. The MUF under the home control area point is determined, inserted in the key-

```
NO. HOPS            4.0      PRT
HOP LENGTH       3342.5      PRT
CONT. AREA DIST. 1671.2      PRT
```

fig. 4. Printout of the hop length, number of hops, and control area distance as computed by the calculator.

---

*User instructions and complete coding forms may be obtained from *ham radio* by sending a self-addressed, stamped envelope.

board, and Label A pushed. The MUF under the DX control area point is next determined, inserted into the keyboard, and Label B pushed. The printer will print out the lesser of the two numbers, since the lower MUF is the controlling one for the path. A little time may be saved during this procedure. If the MUF at the DX location is higher than that of the home location, it will be discarded by your calculator. Therefore, rather than inserting that data into the keyboard, just push Label B and it will use Label A data which shows on the computer display.

Next, the great circle overlay is placed on the higher sunspot number chart and the above procedure repeated. The printer tape will now display MUFs for the lower and higher sunspot numbers. I have found it helpful to print these two numbers to verify that the interpolated number which follows appears correct. If a printer is not used, the numbers can be recorded from the display.

The program for this calculation is quite simple and may be performed on any hand calculator that has "If Pos" and "If Flag" features. I wanted to use the same labels for the minimum solar activity and the average solar activity (or the maximum activity). Label A is used for the home location MUF, and Label B is used for the distant location MUF. For any given solar activity we want to save for interpolation use the lower MUF for the high and low sunspot values.

Fig. 6 is the flow chart for the calculation. Check it out with actual numbers. From Volume 2 (SSN of 10) the MUF is 18 MHz at the home control point and goes into the calculator as MUF A, from the same volume, 25 MHz at the distant control point, and goes into the calculator as MUF B. The difference is $18-25 = -7$, a negative number. Since it is not "If Pos," it goes to a flag check. Since the test for flag is the first time through, none exists, and the proper MUF for SSN 10 is 18 MHz. The value is displayed and stored in memory. At the same time, two flags are set to prevent using this path for the MUF determination at SSN of 110.

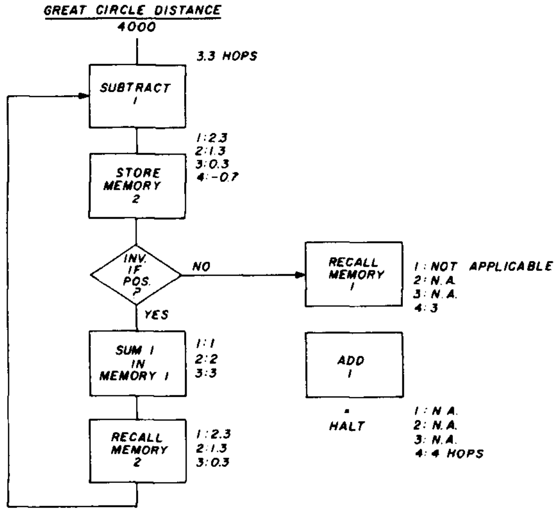


fig. 5. Flow chart for use with the TI-series calculators to determine the number of hops. Hewlett-Packard calculators with the INT function can perform this entire program in five keystrokes.

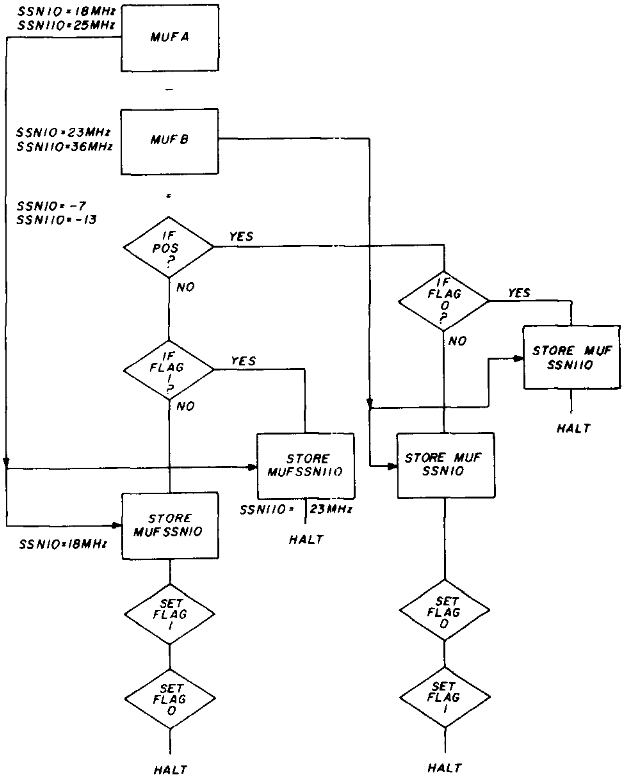


fig. 6. Flow chart to determine and store the lesser MUF value at the home or DX location. This information is needed to interpolate the MUF for the current sunspot number.

Transferring to the SSN 110 chart, the MUF at the home location is 23 MHz and 36 MHz at the distant location. Again the "If Pos" check reveals a negative number and the calculator drops to "If Flag 1." Since that flag has been set, the value of 23 MHz is displayed and stored for interpolation purposes. If MUF A had been higher than MUF B, the calculator would have branched to "If Flag 0" with similar results.

To get the final interpolated answer for the hour used, press the Label D button. When that button is pushed, the calculator solves the equation:

$$MUF = MUF_{10} + 0.01(MUF_{110} - MUF_{10})(R_{12} - 10)$$

where

R_{12} *is the current sunspot number*

MUF_{10} *and* MUF_{110} *are the MUFs at the low and high sunspot numbers.*

When the sunspot numbers exceed 110, Volumes 3 and 4 will have to be used, and the program changed to use:

$$MUF = MUF_{110} + 10.02(MUF_{160} - MUF_{110})(R_{12} - 110)$$

Either of the interpolation equations can be programmed in your calculator or done by hand. With the programmable calculator, it is necessary only to recall MUF_{10} and MUF_{110} from memory and wait for the answers. It is at this point that the two flags have to be reset so that the program can start over for the next time increment.

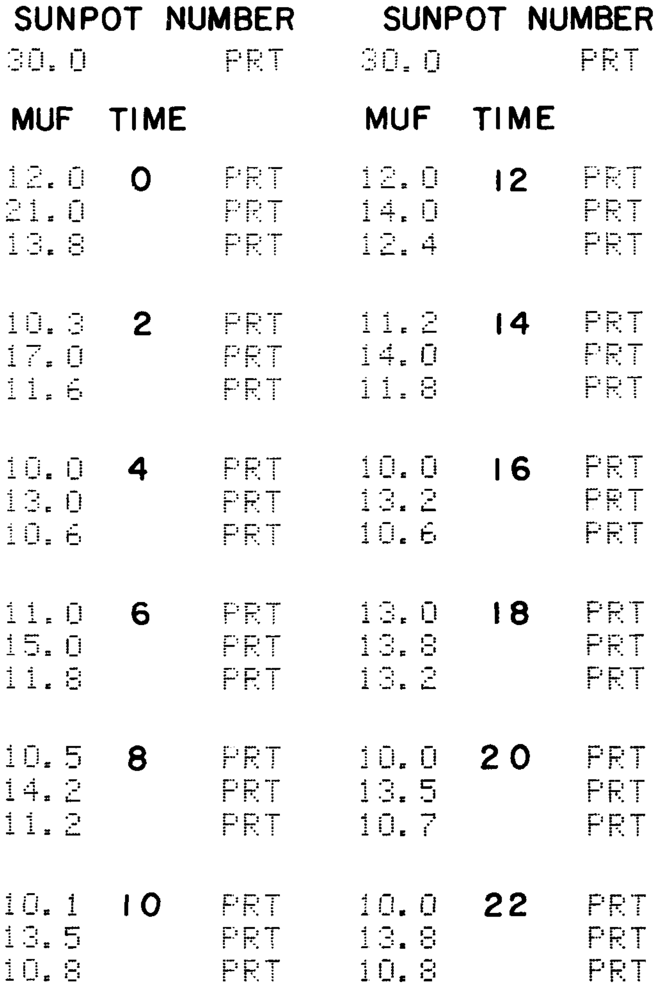
If a printer is not used, the information is recorded from the calculator's register. When finished, you will have twelve points to plot for a 24-hour period, and the printout will be as shown in **fig. 7**. Plot them on graph paper using a suitable scale for the frequency to give a presentation such as shown in **fig. 1** or **2**. You will now have a 24-hour short-path MUF presentation to your desired DX location. Altogether the procedure has taken 10 to 15 minutes.

Don't stop at this point, however; also compute the long paths. Some of them are rather exciting. Don't be confused by the referral in Volume 1 to short and long paths. When they say short path, they mean a short distance, like Washington, DC, to Ottawa, Canada. A long path to them is Washington, DC, to Berne, Switzerland.

The long-path computation is treated just like the short path computation, but with different numbers and control area points. Since you already know the short-path distance, subtract it from 40,000 km to determine the long path distance. For our purposes, 40,000 km is sufficient for the world circumference.

Now when using the program, the calculator will give long path information to establish the ionospheric control area distance. The long path great circle route and the new control area must be put on the transparent overlay. Before repeating the prediction procedure, it will be necessary to reinsert the sunspot number as that information was cleared from memory when the program was activated. The "clear memory" operation is automatically accomplished when the prediction program is actuated to eliminate stored information from being carried over from the bearing/distant program.

Repeating the prediction exercise will give twelve new points to plot on your graph for the long path.



```
SUNPOT NUMBER              SUNPOT NUMBER
30.0          PRT          30.0          PRT

MUF   TIME                 MUF   TIME

12.0   0      PRT          12.0   12     PRT
21.0          PRT          14.0          PRT
13.8          PRT          12.4          PRT

10.3   2      PRT          11.2   14     PRT
17.0          PRT          14.0          PRT
11.6          PRT          11.8          PRT

10.0   4      PRT          10.0   16     PRT
13.0          PRT          13.2          PRT
10.6          PRT          10.6          PRT

11.0   6      PRT          13.0   18     PRT
15.0          PRT          13.8          PRT
11.8          PRT          13.2          PRT

10.5   8      PRT          10.0   20     PRT
14.2          PRT          13.5          PRT
11.2          PRT          10.7          PRT

10.1   10     PRT          10.0   22     PRT
13.5          PRT          13.8          PRT
10.8          PRT          10.8          PRT
```

fig. 7. Calculator printout for the MUFs during two-hour time increments.

Now, you have a complete picture and can establish your strategy and working hours to snag that elusive DX. Sure, you're still going to fight the pileups at times, but you will be there waiting for him with the beam pointed in the correct direction, and get him before many of the other guys wake up to his presence. Or, you'll be listening at a time and on a band which is most favorable to you.

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4. *Ionospheric Predictions*:

Volume 1, The Estimation of Maximum Usable Frequencies from World Maps of MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E, stock number 03000318, price $0.30.

Volume 2. Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E for a Period of Minimum Solar Activity, $R_{12} = 10$, stock number 03000319, $3.00.

Volume 3, Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E for a Period of Minimum Solar Activity, $R_{12} = 110$, stock number 03000320, $3.00.

Volume 4, Maximum Usable Frequencies MUF (Zero) F_2, MUF (4000) F_2, and MUF (2000) E, for a Period of Minimum Solar Activity, $R_{12} = 160$, stock number 03000321, $3.00.

Write to the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402.

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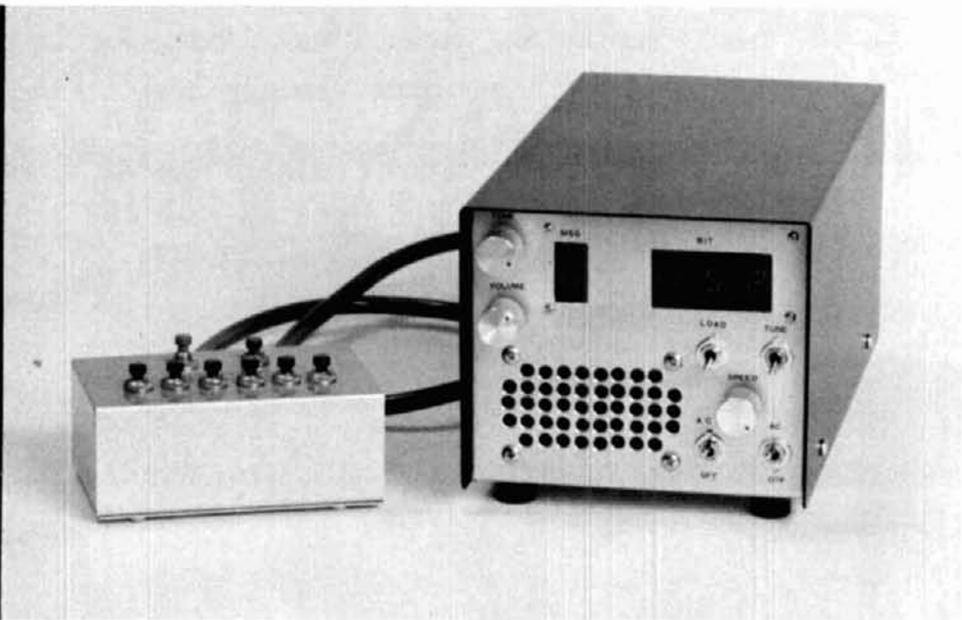


deluxe memory keyer

with 3072-bit capacity

A revision of the famous WB4VVF Accu-Memory, featuring improved automatic character spacing, six-message memory, and remote control

A memory keyer with the right characteristics offers many advantages in contests, net-control operations, and DX chasing. It can relieve an operator of much of the routine repetitive sending, generally reducing fatigue and providing time for efficient log keeping, checking for duplicate contacts, and occasionally taking a sip of coffee, even during rapid-fire contest exchanges.

The Accu-Keyer,[1] as originally described in *QST*, was a nonmemory keyer with many desirable features. These included self-completing dots and dashes, next-dot and next-dash memories, iambic operation, and optional automatic character spacing (ACS). Later, WB4VVF and W4YUU described the companion Accu-Memory,[2] which, when used along with the Accu-Keyer, permitted storage of up to 2048 bits (about 200 CW characters) keyed in from the keyer. These could be read out as four individually selectable messages of up to 512 bits each, or as one or more longer continuous messages.*

---

*The term *bit* is used throughout as the element of information contained in a dot or a space in a CW character. A bit interval is the time duration of a single dot or space. A dash is three bits long, the standard space between characters is three bits, and a normal word space interval is seven bits long.

By Robert C. Cheek, W3VT, c/o Ebara-Infilco Co., Ltd., 1-1-1 Hitotsubashi, Chiyoda-ku, Tokyo 100, Japan

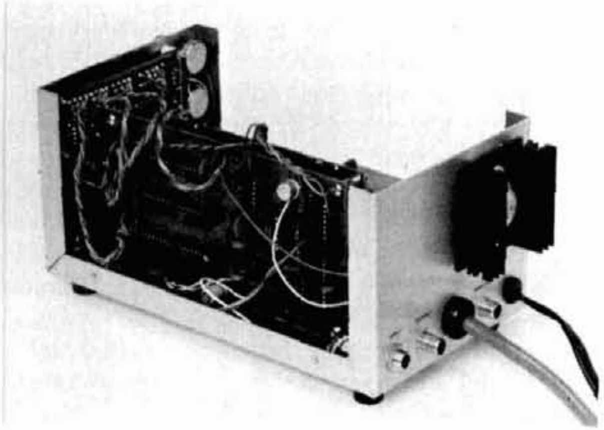
I built a modified version of the Accu-Keyer with the Accu-Memory, incorporating those modifications I thought desirable. My own experience with it, and the comments and questions of others, indicated that further improvements and enhancements could be made. I have now gone through the redesigning and building process five times, with the memory keyer described in this article the final result.

The memory capacity of this keyer is 3072 bits, in six selectable, 512-bit segments or in longer segments up to the full capacity of the memory. The entire system, except for the power supply and the display LEDs, is contained on two circuit boards measuring 80 by 150 mm (approximately 3-1/8 by 6 inches). ICs are used throughout, except for the output, which employs a keying transistor. When the memory is being loaded, the IC clock is synchronized to manipulator movement. Readout from memory may be automatically interrupted at any time by manual keying. A choice of keyer output circuitry to accommodate either grid-block keying or positive-line keying is provided for in the circuit-board layout. A cable-connected, remote-control unit is used for manual start, stop, and message selection functions. It can be conveniently placed on the operating table to be played with the left hand like the buttons on an accordion while the right hand makes log entries and operates the keyer paddle for station calls, nonstandard message insertions, and so on.

system description

A functional block diagram of the complete system is shown in **fig. 1**. Those who wish to follow the explanations in detail should also refer to the schematics of the two main boards, **figs. 2** and **3** for the keyer-driver-monitor and memory boards, respectively. This explanation assumes a knowledge of the specific IC characteristics and truth tables given in IC data manuals.

Rear view of the six-message memory keyer showing the driver-monitor board; an adequate heatsink must be used on the 5-volt regulator.



Keyer. The keyer proper (included in **fig. 2**) is basically the well-known Accu-Keyer, with all of the original features retained. Keyer operation was explained in the original article[1] and will not be repeated here, except for those portions of the circuit that have been changed.

The original clock pulse generator, made up of discrete transistors, has been replaced by an NE555 IC clock located on the memory board. As will be explained later, the clock output approximates a square wave rather than sharp pulses. This characteristic is used to greatly reduce the error rate in manual keying when the ACS feature is in use.

The ACS circuitry inserts precision, three-bit spaces between characters, correcting premature timing of one character following another in the manipulation of the keyer lever. Without ACS, this shows up as a shortening of the proper three-bit character interval. For example, *DE* may sound almost like *B* if the *E* is keyed too quickly after the *D*.

However, the original ACS feature is difficult to use at higher speeds without errors, which show up as unintentional extra spaces. This happens because lever manipulation must not only be fast, but must actually slightly lead, or at least never lag behind, in the formation of characters in which a dot is followed by a dash, such as an *A*. Those of us who grew up with the bug or semi-automatic key tend not to lead while keying, but to insert our own approximately correct one-bit interval between the dot and the dash in the formation of such characters. But, if the dash lever is tapped just a fraction of a bit space too late, an *A* becomes *ET* in the original keyer when ACS is in use. This occurs because the outpin at pin 6 of U5A (see **fig. 2**) is always low during the one-bit space following a completed dot or dash, and then always goes high for the next one-bit interval. It is this high transition that transfers a waiting dash or dot from the next-dash or next-dot memory, U1A, U1B, or U2C, U2D, as seen through the iambic gates, to the present dash or dot memory, U3A or U3B, starting the output of the dash or dot. If the lever is tapped even very slightly late, this transition is missed and two additional bit spaces are inserted.

U8 in the keyer has been added to provide some

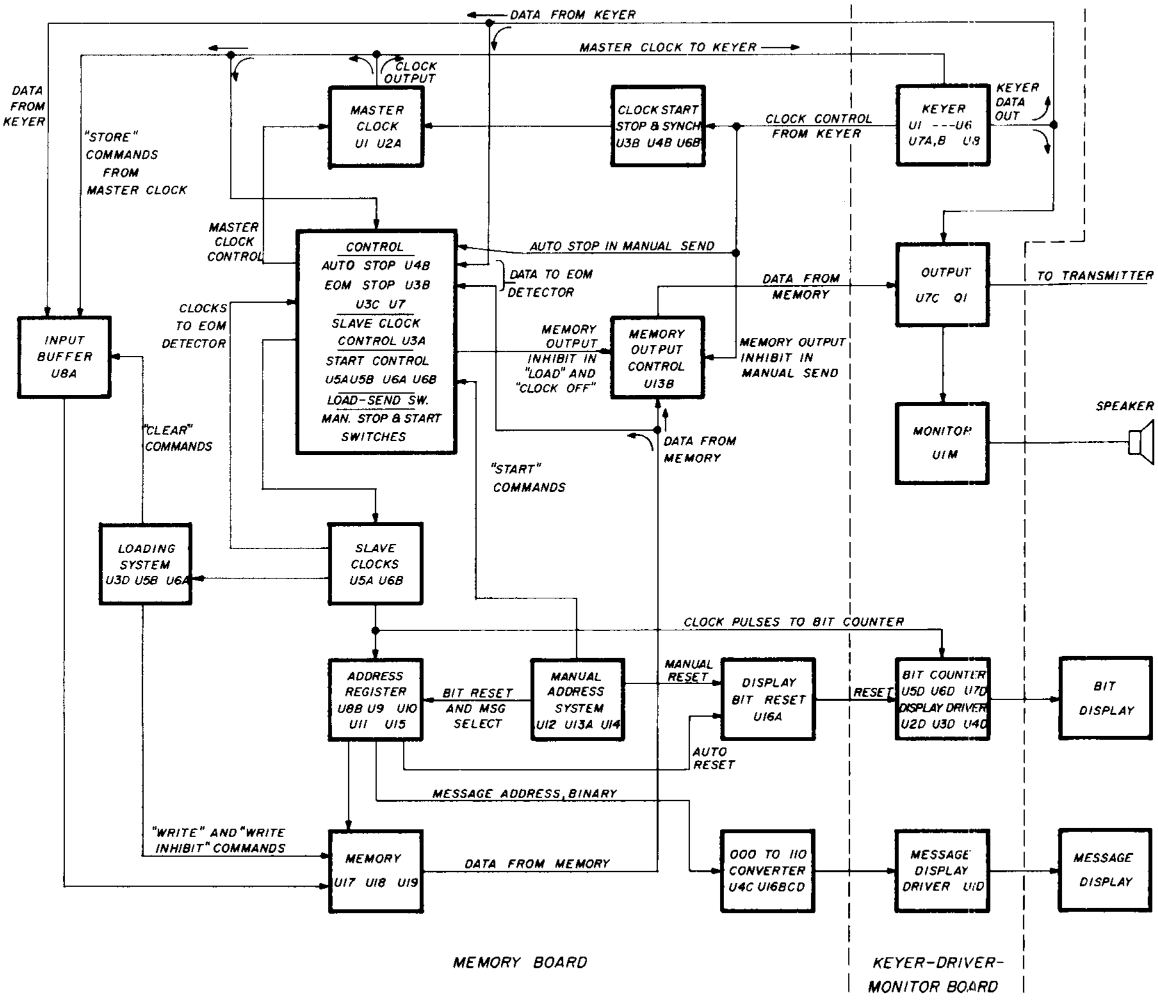


fig. 1. Functional block diagram of the memory-keyer system. Details of the keyer block are not shown; these are given in reference 1.

tolerance for slight manual timing delays in the keying of dashes following dots as described above. The approximate square wave from the clock is high at pin 1 of U8 for the first half of the bit interval after pin 6 of U5 goes high. If neither a dash nor a dot has already been called for, pin 8 of U6 will also be high. If the dash paddle is tapped during this half-bit period, U1D, pin 11, will go low. This low, inverted by U8B, will cause U8A, pin 6, to go low because all four inputs are high. The low will set U3A, pin 6, high and start the dash, although a little late. However, even though the interval between the dot and the dash may be lengthened by as much as half a bit and the dash correspondingly shortened by this amount, the result is almost undetectable to the ear and is far better than the three-bit interval that otherwise occurs, causing a completely erroneous character to be formed. Also, since data as it exists at the *end* of the bit interval is what is recorded into memory in the loading process, the correct one-bit interval between the dot and the dash is restored when the message is sent from memory.

The keying of a dash following a dot *after* the half-bit tolerance interval provided by U8 results in the insertion of the normal three-bit character interval, retaining the desirable features of automatic character spacing. At the beginning of the third-bit interval during a character space, pin 6 of U5A again goes low when the clock pulse starts. The RC filter at pin 1 of U8 delays the clock pulse slightly so that pin 6 of U5A has time to go low before the clock, as seen by U8A, goes high. This prevents a shortened dash from prematurely appearing if the dash lever has by that time been tapped for the next character.

By adding another IC equivalent to U8, it would be possible to provide the same tolerance for manual keying of a dot or dots following a dash. However, nearly all the tendency for errors at high speed is the result of the very short and critical interval for the keying of a dash following a dot. Since the available interval is inherently much longer going the other way, the extra complication would not be worthwhile.

Clock. The clock, as shown in **fig. 4**, is the heart of the keyer and the memory. Both require a clock signal which is at zero level in the quiescent state, remains at zero for the duration of the first bit interval after the clock is called for, and then provides a steeply rising positive pulse or square wave at the beginning of each bit interval thereafter. With the control scheme adopted, the combination of the NE555 timer and an inverter provides the necessary characteristics, including the square wave required for the ACS error-reducing circuit in the keyer.

Pin 4 of U1, which is normally used as the control input in NE555 applications, is maintained high at all times, keeping pin 3 normally high, with the clock output being low. When pin 11 of U3B is low, calling for the clock to be off, the timing capacitor is essentially discharged through CR1. Pin 3 of U1 remains high and the output of U2 stays low. When the clock is called for by either the keyer or the memory, pin 11 of U3B goes high and the timing capacitor begins to charge at close to a linear rate. The output from the inverter remains low during this process. When the charge reaches 2/3 of V_{cc} (the supply voltage), pin 3 of U1 goes low, forcing the inverter high, marking the end of the first bit interval. Thereafter, the charge on the timing capacitor oscillates between 1/3 and 2/3 of V_{cc}, and the output at U2 goes high at the end of each subsequent bit interval until pin 11 of U3 again goes low, stopping the clock.

CR1 and CR4 permit independent discharge of the timing capacitor for clock synchronizing purposes, as will be explained later.

Memory system (load mode). The basic memory loading system of the Accu-Memory[2] is essentially retained. In the load mode, with a message address manually selected, the clock control line from the keyer goes low at the moment keying begins and pin 11 of U3B (see **fig. 3**) goes high, starting the master clock. At the same time U6B, pin 1, goes low, causing a pulse at pin 13 of U6. This pulse is passed along by U5B and then appears at pin 5 of U6A, from where it is applied to pin 3 of U7, the end-of-message detector. This pulse at pin 3 performs an AND function with the data at pin 2 of U7, setting the output at U7 pins 1 and 11 to zero, whereupon pin 8 of U3C will go high and pin 12 of U2B low, keeping pin 11 of U3B high. The high at U3C pin 8 also enables slave clock control U3A. Pins 9 and 10 of U3C are both high only when U7 is at a count of nine. Thus, even though the clock control line from the keyer may go high, U7 will keep the clock running until it counts nine consecutive pulses at its pin 14 from slave clock U5A, pin 5, without being reset by the pulses at pin 3 coincident with data at pin 2.

At the end of the first bit interval, the leading edge of the master clock pulse transfers the data at the D input of the 7474, U8A, pin 12 to U8A pin 9. Halfway through the second bit interval, the trailing edge of the first master clock pulse, having been inverted by U3A and applied to pin 10 of U5A and U6A, causes positive pulses at both Q outputs. The write inhibit, provided by gate U3D, is removed by the pulse from pin 5 of U6A, and the first-bit data is recorded into the memory. The pulse from U5A is somewhat longer in duration. Its negative-going trailing edge, occurring later, advances the address register by one, and at the same time the positive-going $\overline{Q}$ output at pin 12 of U5A causes pin 4 of U5B to pulse low, clearing the buffer, U8A, to zero. At the same time, pin 13 of U5B pulses high, and its subsequent trailing edge causes another positive pulse at pin 5 of U6A, again removing the write inhibit and recording the previously mentioned zero from U8A in the second memory address. All of this takes place during the latter part of the second-bit interval.

At the end of the second-bit interval, the loading process is repeated for the second data bit, and the zero previously stored in the second address is changed to whatever the contents of the data buffer may be for the second bit. If there is a pause in keying, the end-of-message counter U7 will stop the master clock when it counts nine consecutive bits without being reset by pulses from U6A along with the data. The ninth bit will be changed to a one if keying is resumed, and the pause, no matter how long, becomes a word space (slightly lengthened to eight bits) when the message is read out in the send mode. However, if the message has really ended and no additional data is loaded, the zero remains in the ninth bit position, and in the send mode U7 will stop the clock at that point.

The master clock free runs during the word intervals in the load mode, and it is necessary to synchronize the clock to the paddle movements during loading so that shortened dots and dashes do not appear, as they would if the paddle were tapped to start a new word just before a clock pulse was about to occur. Synchronization is accomplished by U6B, U4A, and U2C. When the paddle is tapped during loading, the clock control line from the keyer goes low, calling for the clock (which normally is already running). This low, applied through CR5 to pin 1 of U6B, causes the $\overline{Q}$ output (pin 4) of U6 to pulse low. This pulse forces the output of U4A high and U2C low. The timing capacitor is momentarily discharged through CR2 by this pulse and then allowed to charge normally, ensuring full bit intervals for the

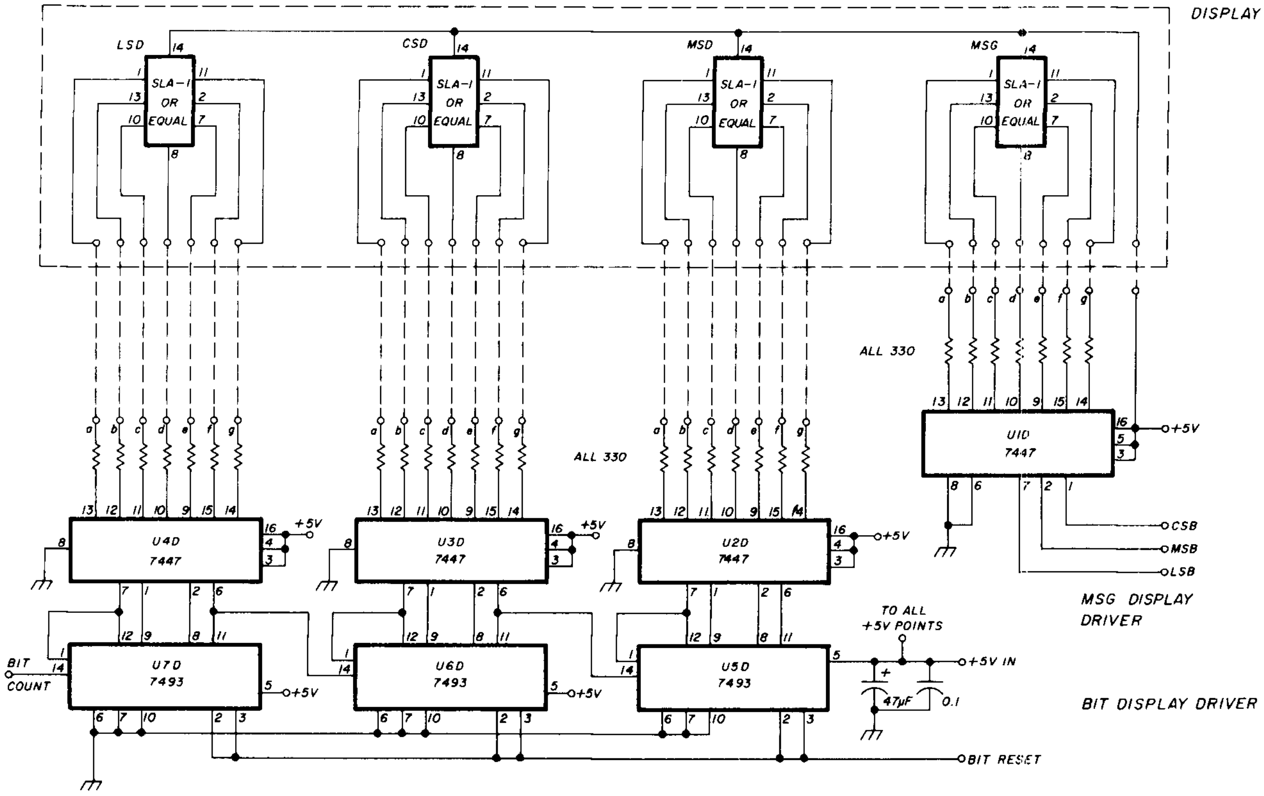


fig. 2. Schematic of keyer-driver-monitor board and display. All resistors are ¼-watt. Capacitors with polarity indication are upright electrolytic, 15-volt minimum rating; all others are disc ceramic, 25-volt minimum rating unless otherwise specified. The diodes are small signal silicon, switching diodes, 1N4148 or equivalent. Dotted lines indicate connections to external switches and controls. Small circles with labels, except +5V to ICs, indicate similarly labeled points on the board or power supply. In the output circuit for negative grid-block keying, Q1 may be a 2N4888 or any high-voltage audio, switching, or rf silicon pnp transistor with a V_{CEO} rating adequate for the open-circuit voltage to be keyed. The unit pictured uses a Japanese type 2SA510 (RCA SK3025 is also a recommended equivalent). For positive-line keying, an equivalent npn transistor is used at Q1.

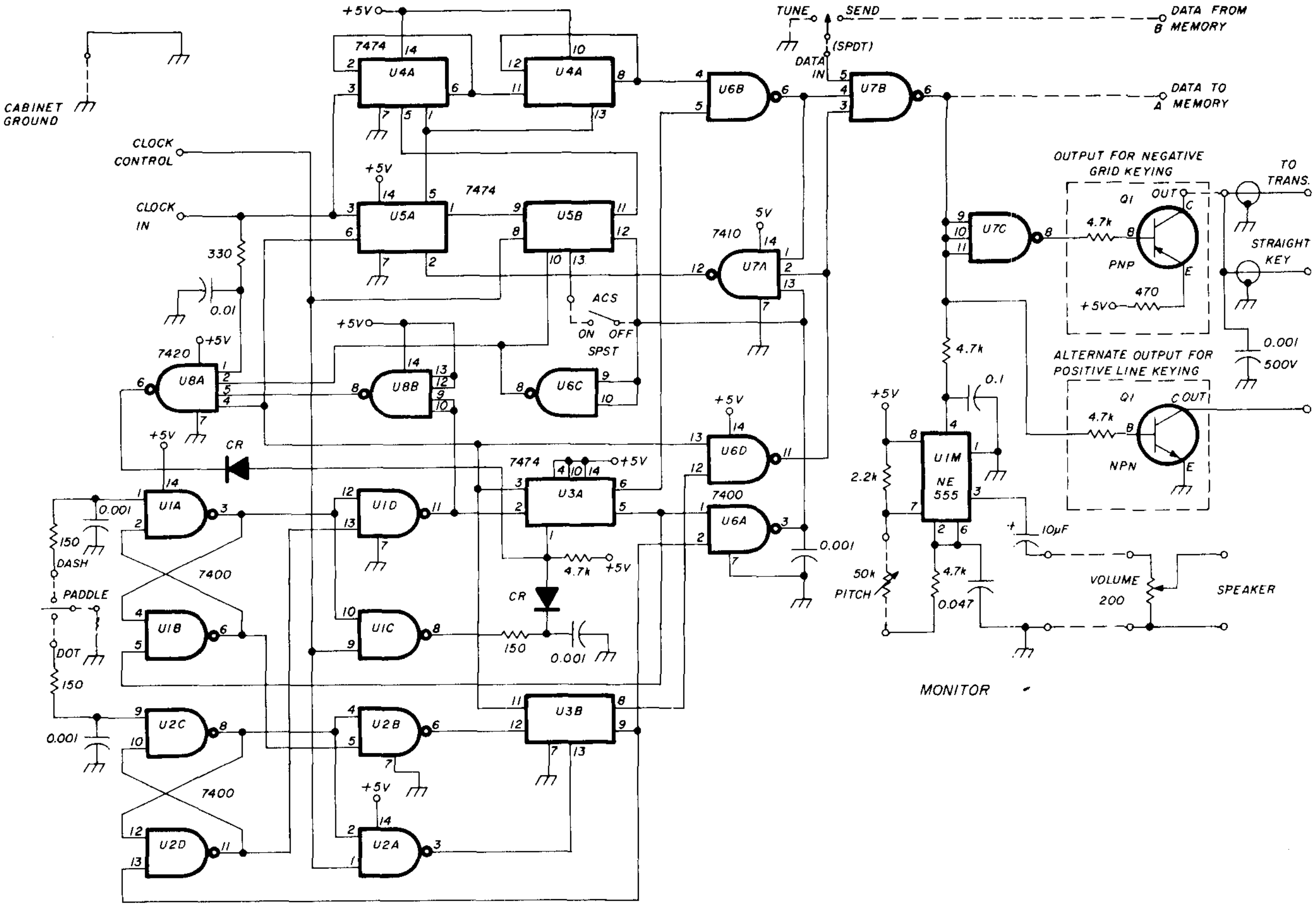
first bit and subsequent initial bits of manual sending.

In the load mode, the load/send switch grounds pin 12 of U13, preventing any output of random or previously stored data in the memory from reaching the output data line. While loading is progressing, only the new data from the keyer can key the transmitter.

Address register and manual addressing system. The address register consists of U9, U10, and U11A for the lowest nine binary address bits (512 of the 1024 positions on each memory chip). The tenth address bit, determined by U11B, becomes the least significant message address bit. Thus, U11B divides each 1024-bit memory chip into two 512-bit message segments, selected by making the output from pin 9 either a zero or a one. U8B, U15A, and U15B constitute a shift register or ring counter which activates one memory chip at a time for loading or readout by placing pin 13 of the selected chip low, while maintaining pins 13 on the other two chips high.

The entire address register is controlled by the manual addressing system, consisting of U12, U13A, and U14. When a message button is pushed (assuming the clock is not running) pin 6 of U13 goes high, setting U9, U10, and U11A to zero. Pin 9 of U11 is set either low or high, depending upon the message selection, and the Q output of one of the three flip-flops in the ring counter is set low to address the desired memory chip. The other two Q outputs in the ring counter are simultaneously set high. Whenever a manual message selection is made, the display bit counter is simultaneously reset to zero by U16A.

Since memory-chip selection is made by setting one of the Q outputs in the ring counter low, the $\overline{Q}$ outputs of U15A and U15B are used as the central significant bit and the most significant bit, respectively, of the message address. When both these Q outputs are high ($\overline{Q}$ outputs both low), the memory chip that is activated by a low from the Q output of U8B is in use, and the message address is either 000 or 001 binary (decimal 0 or 1), depending upon the



KEYER

MONITOR

Q output at U11B. The other binary message addresses are 010, 011, 100, and 101 (decimal 2, 3, 4, and 5). Since it is better in operation to have the six messages numbered 1 to 6 rather than 0 to 5, U4C and U16B, C, D were configured to convert 000 binary to 110 binary, leaving the other binary indications unchanged. Thus, message 0 becomes message 6 for all selection, display, and operating purposes. The entire addressing system is a continuous counter, so that a message as long as the entire 3072-bit memory capacity can be started at any selected beginning address and loaded continuously in memory, with automatic transition through the consecutive message segments back to the starting address.

Send Mode. Before a selector button is pushed, the clock is stopped by a low at pin 11 of U3. This low also inhibits U13B, blocking the continuous output that would otherwise appear if the address register happens to be stopped at a memory address where a high exists. When a selector button is pushed, the high at U13, pin 6, which resets the sit address register to zero, is inverted by U2D, simultaneously causing a low at pin 1 of U6B. The resulting pulse at U6, Pin 13, triggers U5B which in turn causes a pulse at pin 5 of U6A.

A high always exists at memory bit address 000 when a message segment is loaded. This high is applied to pin 2 of U7, and the pulse from U6A is applied to pin 3. The AND function at these two points resets U7 to zero, unblocking the memory output and starting the clock. The high at pin 11 of U3B also causes U13A to return to low, even if the message selector switch is kept closed, removing the reset on the address register. Slave clocks U5A and U6A are enabled through U3A when U7 resets. U5A advances the address as described for the load mode. During each bit interval, a pulse from U6A is applied to U7, resetting U7 whenever data appears at pin 2 of U7. Unless the stop button is pushed, sending from memory continues until U7 counts the nine consecutive bits with zero data which mark the end of the message.

Automatic Stop. A message readout in progress

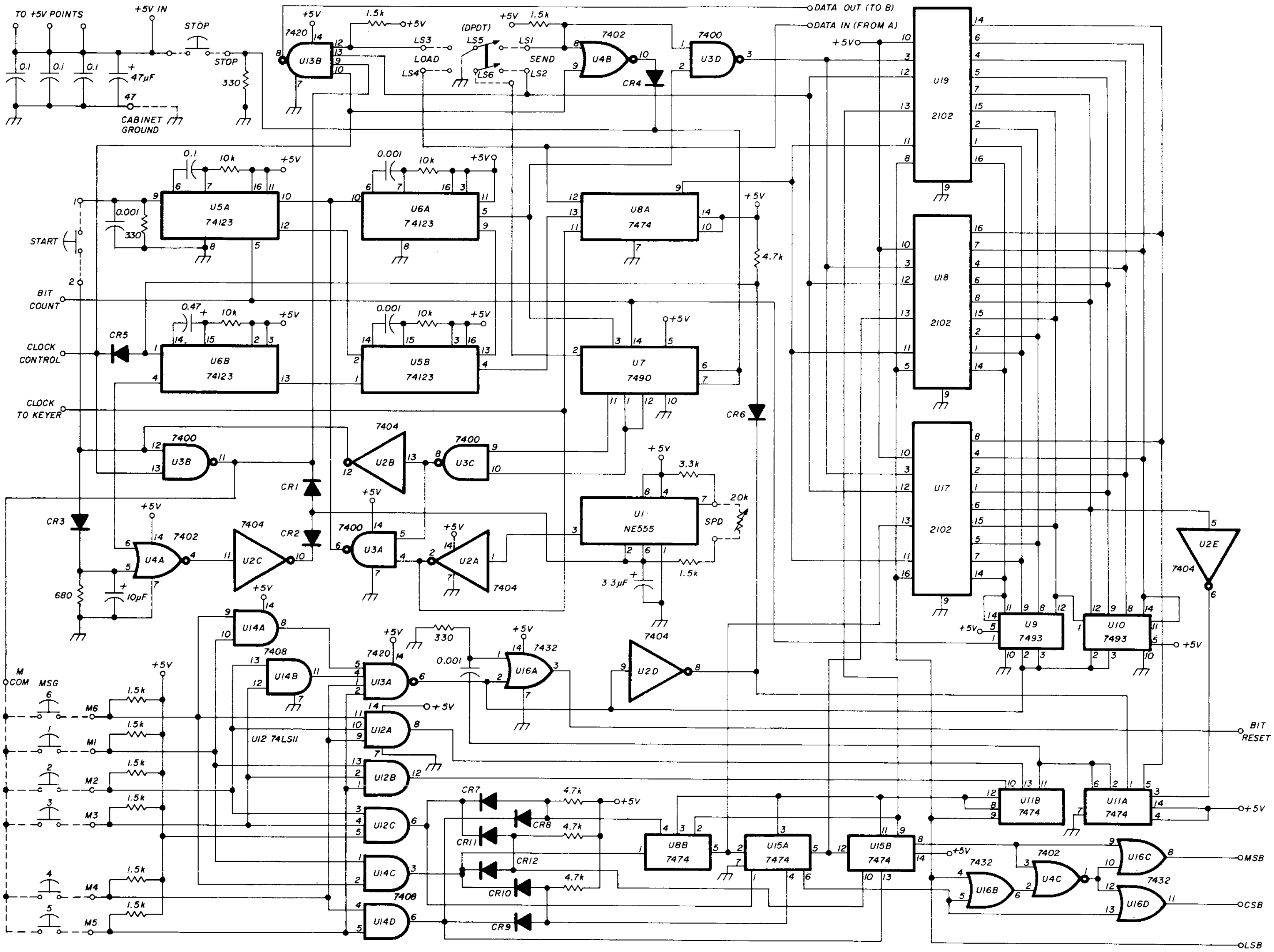


fig. 3. Schematic of memory board. See fig. 2 for explanation of component designations and external connection symbols. All diodes are small signal silicon switching diodes, 1N4148 or equivalent. START, STOP, and message selector switches are all momentary push-button type.

can be interrupted by manual sending. At the first tap of the paddle, the clock control line from the keyer will go low, forcing U4B pin 10 high. This high, applied through CR4 to U7, forces a nine count, instantly turning control of the master clock over to the keyer and stopping the input to pin 10 of slave clock U5A. Meanwhile, the low on the keyer clock control line causes U6B, pin 4, to pulse low. This pulse causes U4A to pulse high (provided pin 5 is low) and U2C to pulse low, synchronizing the master clock to the first bit of manual sending. The forced nine count at U7 causes U3C to go low and U2B to go high, but the 10-μF capacitor at pin 5 of U4 keeps that pin momentarily low, allowing the synchronizing pulse to pass. Thereafter, the capacitor charges, U4A stays low and U2C high, and the clock is under normal control of the keyer.

Power Supply. The system requires 700 to 800 milliamperes at 5 volts. A schematic of an adequate 5-volt, 1-ampere supply using an LM309K regulator is shown in **fig. 5**. Based upon its specifications, the TTL 5-volt, 1-ampere regulated supply kit offered by Jameco should be equally suitable.

construction

Assembly of the circuit boards* (**figs. 6, 7,** and **8**) offers no unusual problems, but the usual cautions apply — use a low-wattage iron and small diameter 60-40 solder;· be sparing of solder, but cover the eyelets completely; above all, make sure there are no *unintentional* solder bridges between IC socket terminals or between adjacent separate conductors in the vicinity of component terminals.

The use of IC sockets or Molex pins is recommended in preference to soldering the ICs directly to the boards. There is a big difference, however, among IC sockets, so be sure to use the compact, slender type rather than any of the more bulky types, as the latter may cover up a few of the component or jumper eyelet holes adjacent to the ICs. The digital display LEDs are intended to be soldered directly to the display board. If you prefer to use sockets, the unused socket pins must be clipped off to permit their mounting in the holes provided.

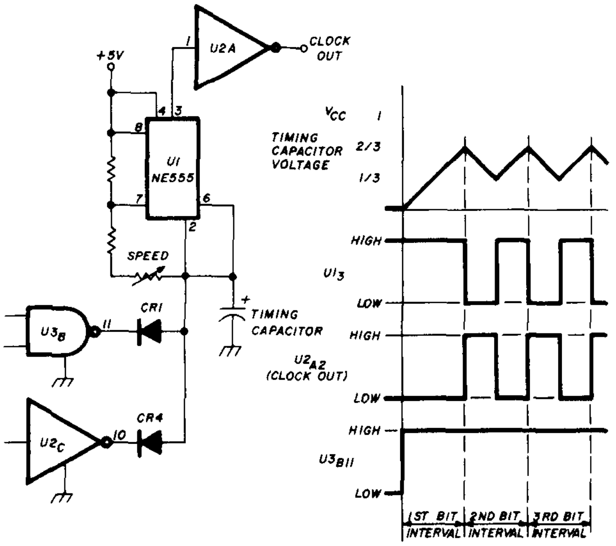


fig. 4. Diagram of the basic clock circuit and clock control logic.

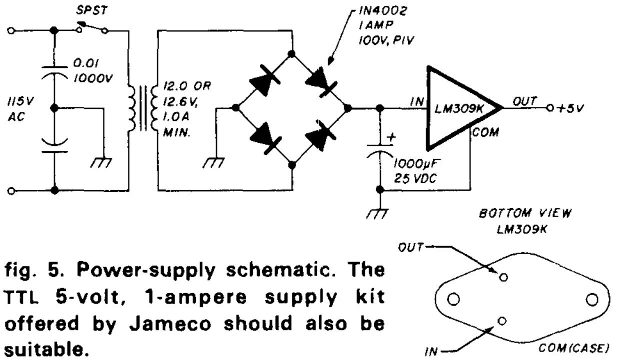


fig. 5. Power-supply schematic. The TTL 5-volt, 1-ampere supply kit offered by Jameco should also be suitable.

Some of the wire jumpers on the memory board must be routed around the IC sockets. Mount the sockets first, and then install the jumpers before mounting the other components. All jumpers are installed on the component sides of the boards. The short, isolated, straight jumpers can be made of bare wire to save space, but obviously the longer ones and those which cross over uninsulated jumpers or come close to other component leads should be of insulated wire.

For interconnections between the boards and leads to be connected to external switches and controls, use small-gauge flexible (stranded) wire to prevent strain on the circuit board eyelet terminals. In making the main board-to-board interconnections, lay the two boards side by side, foil sides down, with the top edges adjacent to each other about 4 cm (1-1/2 inches) apart. Recheck this spacing each time you cut a new lead to be sure its length is sufficient to permit the leads to come over the top edges of the boards when they are finally installed.

All connections to the main boards are made with leads inserted from the component side. In the case of the display board, however, all connections are inserted into the eyelets and soldered from the *foil* side, with the excess wire clipped off flush on the component side. In this way, the leads do not interfere with the mounting of the LEDs against the transparent, colored plastic strip that covers the inside of

---

*A complete set of circuit boards can be obtained from G. R. Whitehouse. For more information on the circuit boards and other components, write to G. R. Whitehouse and Company, Newbury Drive, Amherst, New Hampshire 03031.

the front-panel cut-outs. The use of color-coded wire will permit the seven leads for each seven-segment LED and the one +5 volt lead to be twisted together before final soldering, providing a much neater job than the mess that will result if each of the twenty-eight wires is run separately between the keyer-driver board and the display.

After assembly and interconnection, the two main

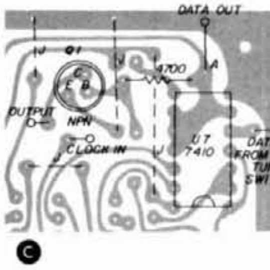
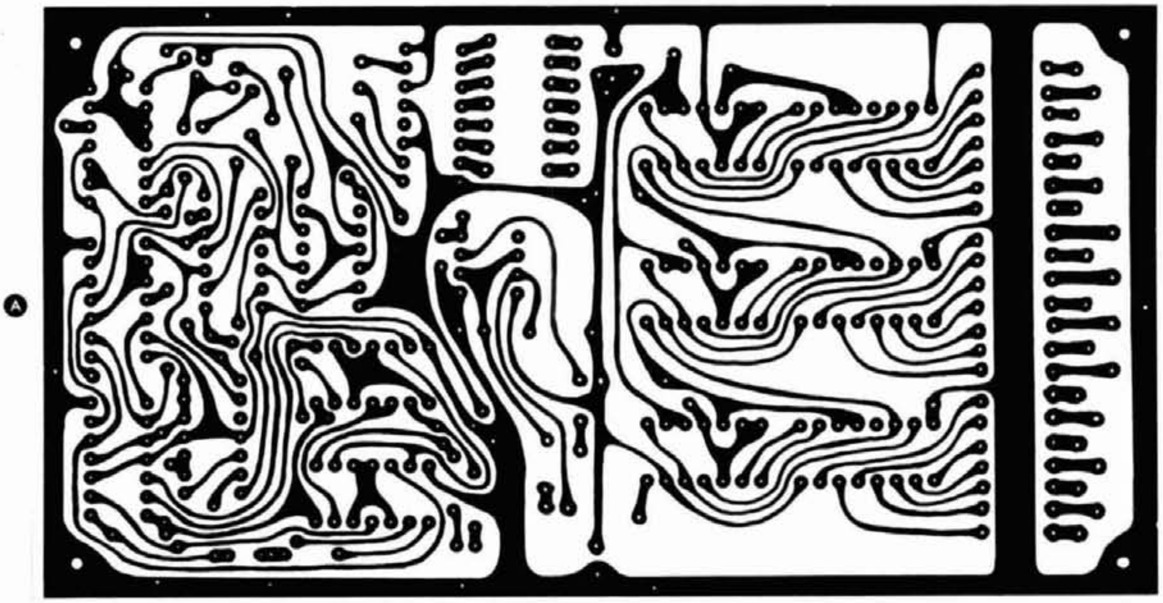
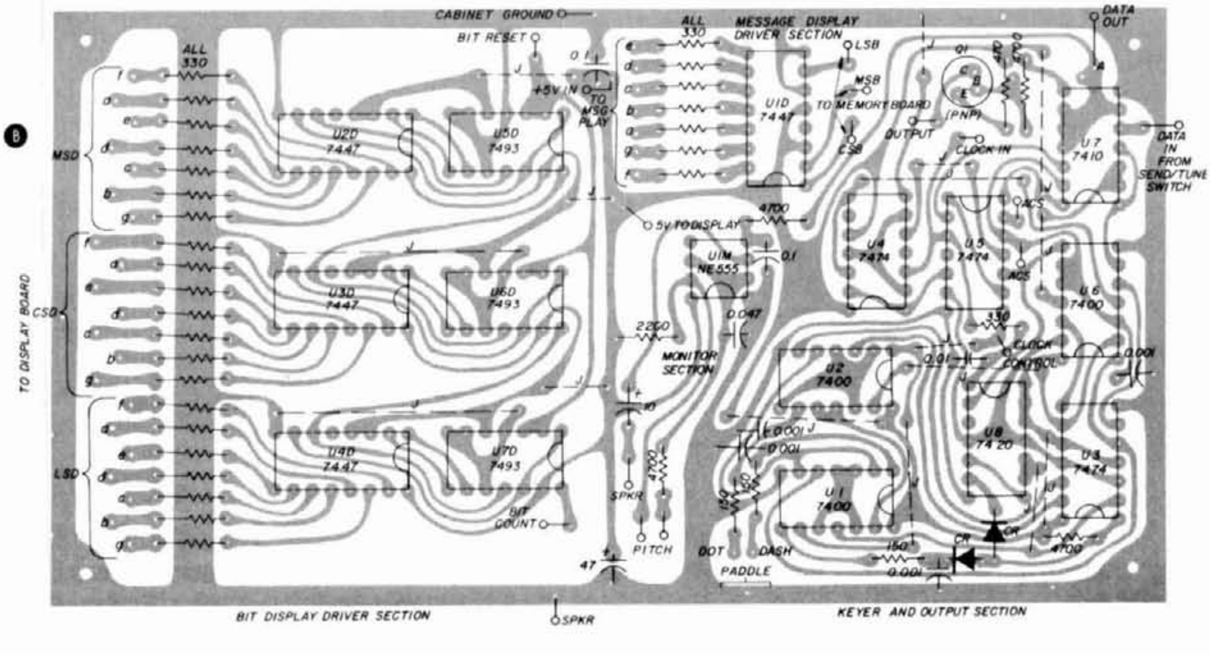


fig. 6. Circuit pattern of keyer-driver-monitor board (A), and component placement for negative grid-block keying (B). Fig. 6 (C) shows alternative component placement for positive-line keying.





circuit boards are mounted to each other with foil sides facing, separated by 9-mm (3/8-inch) cylindrical spacers at each corner. This assembly is mounted to the bottom of the cabinet with four L brackets, one on each side of the bottom corners of the board assembly. This makes a very solid unit, easily accessible for replacement of ICs. The board assembly is mounted on the right-hand side of the cabinet to make room for the power supply at the left rear.

The only precaution worth mentioning in connection with the power supply is to use an adequate heatsink with the LM309K regulator, which is mounted on the rear panel outside the enclosure. The regulator output terminal, projecting inside, will serve as the +5 volt source terminal for the leads to the main boards and the +5 volt lead to the stop switch. Note that both sides of the ac line cord should be bypassed to ground near the point at which it enters the cabinet.

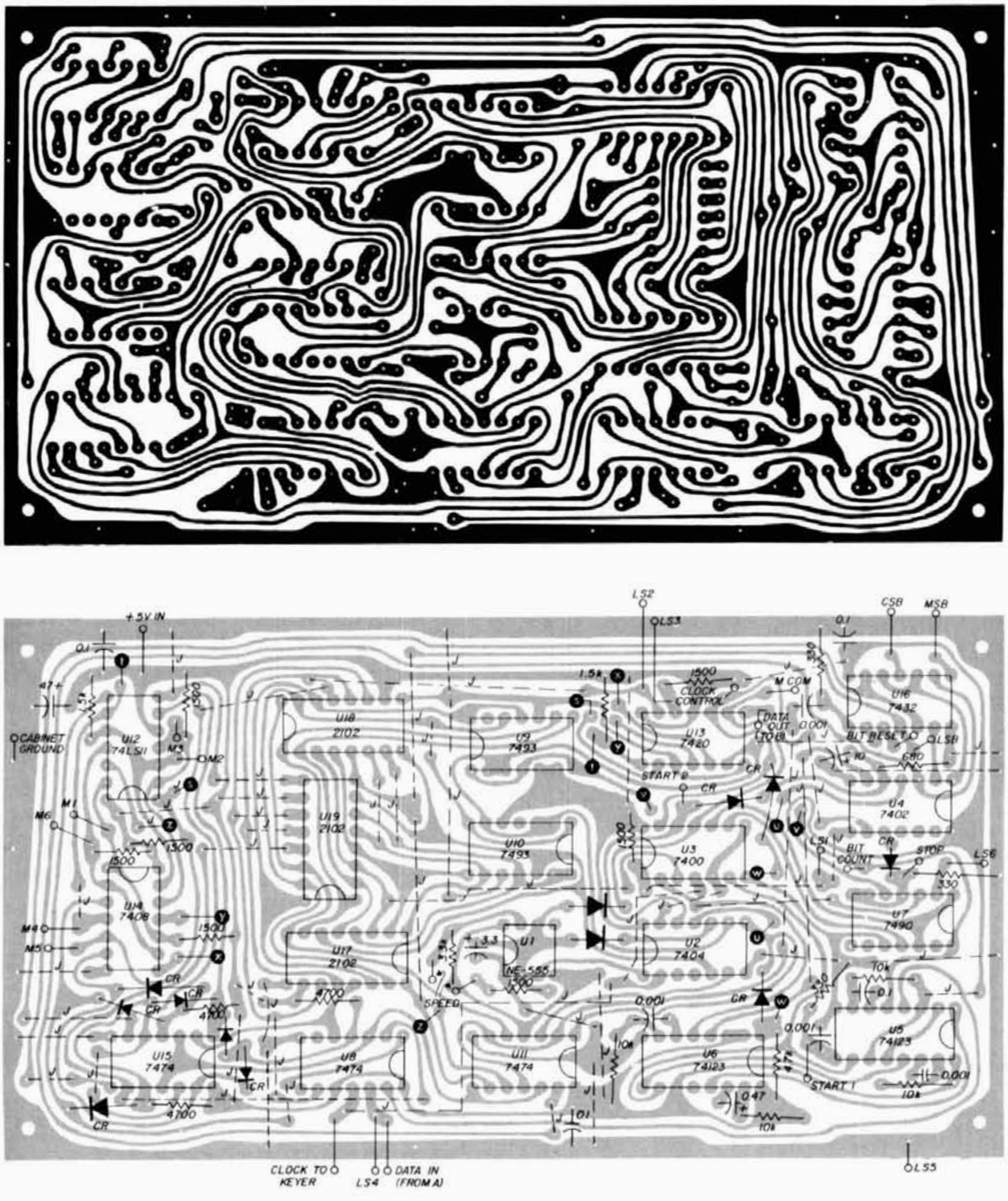


fig. 7. Memory-board circuit pattern (above) and component placement (below). For clarity, several jumpers were omitted from the parts placement diagram. These can be added to the board by connecting the jumper between the lettered designators.

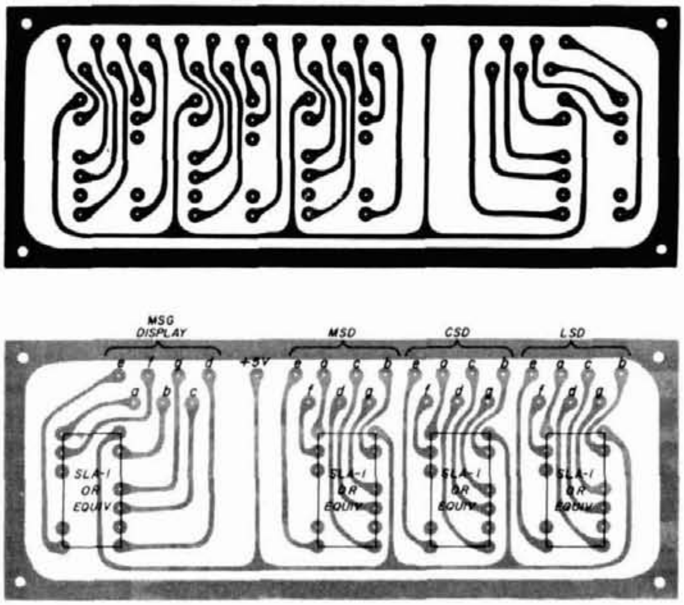


fig. 8. Display-board pattern (above) and LED placement (below). The display LEDs may be SLA-1 or any of several equivalent 8-mm (0.3-inch) common-anode displays with 14-pin DIP pin spacing and the same pin-out configuration for the seven segments. Decimal points are not used and their pin-out locations are not important.

RF shielding of all external leads to the main boards and the use of an all-metal cabinet are absolutely necessary. The leads to the remote-control box (which also must be a metal box) should consist of a shielded cable (11 conductors plus shield are needed) with the shield grounded both in the box and in the main cabinet. The keyer-paddle leads should also be shielded. A two-conductor shielded cable, with the shield used as the ground lead, will be satisfactory.

One last precaution — be sure to insert the correct ICs in the correct locations with the positioning notches in the proper direction, and then recheck each one before applying power. This is very obvious, but I've ruined a few ICs myself by being a bit careless on this point.

initial tests

After assembly and wiring are completed and checked, set the tune/send switch and the mode switch to send, all controls at about two-thirds scale, and apply power by turning on the ac switch. If some gibberish comes out of the monitor, it is a good sign. Pushing the STOP button should stop the output. Check that the output terminal of the regulator is +5 volts. Each display LED should show some digit at this point, although the actual indication is not important.

Next, check the clock, keyer, and monitor for proper operation as follows: Alternately ground the dot and dash inputs to the keyer. Continuous dots and then continuous dashes should be heard from the monitor. Grounding both inputs simultaneously should produce alternating dots and dashes. Switching the tune switch to TUNE should produce a continuous tone. Return it to normal. Now set the mode switch to LOAD and press each message selector button in turn. Bit count indication should be 000 and message numbers 1 through 6 should be displayed in sequence as the associated buttons are pushed. Pushing the START button should cause the bit address to advance by ten bits and then stop.*

Again selecting message 1, ground the dot input to the keyer, causing it to send continuous dots. The bit indication should advance as these dots are loaded into memory and after bit 511 the message address should change to 2 000. Continue loading dots part way into message 6, stop the keyer by ungrounding the dot input, set the mode switch to SEND and again select message 1. The entire series of dots should play back on the monitor without errors or omissions, and it should be possible to stop the playback at any point by pushing the STOP button or by momentarily grounding the dot or dash input of the keyer. Next, in the LOAD mode, again select message 1. While holding the STOP button down, push the START button down once and release it to get a bit count of 001. (Because of contact bounce it may be necessary to try several times to get just a one-bit advance when the button is pushed.) Load continuous dots from 1 001, again part way into message 6. In the SEND mode, select message 1 and this time push the START button to start playback. Again, the whole series of dots should play back smoothly. These two tests will indicate whether all memory cells in this address range of the memory ICs are okay. Perform similar tests by loading dots beginning at addresses 6 000 and 6 001, continuing into message 1 in each case, to check the remaining range of memory addresses.

using the keyer-memory

The memory may be loaded with a message starting at any selected message address. Simply switch to the LOAD mode, select the message number, and start keying. When the message is completed, another may be begun at a different message address until all beginning addresses are used. (If a

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*This may seem contrary to the previous statements that U7 stops the clock after counting nine bits with zero data. Actually the START button causes the address register to advance by one bit at the start of the first bit interval as the count at U7 goes from 9 to 0. U7 then counts nine additional bits and stops the clock.

message is too long to fit into a 512-bit memory space, it will continue into the next message address.) In the SEND mode, pushing the message selector button for a particular message will instantly start transmission of that message from memory. In contest work, a CQ SS or CQ DX can be loaded as message 1, with several different replies (different reports, etc.) as the next few messages, and *R TU QRZ* . . . etc. as message 6. For contests in which the exchange requires a variable element such as a QSO serial number, load the standard information up to the point where the variable element must be inserted. Then, holding the stop button down, advance the memory two or three bits by pushing the start button. Continue loading the subsequent standard information. In the SEND mode the memory will stop at the point where variable information is required. After this is keyed manually, the rest of the standard message will continue when the START button is pushed. This technique can also be used to load a request for a repeat after a standard number transmission, both in the same message address, to be used in case you miss the number sent back to you.

If you are going to call a station on a schedule, you may wish to send his call sign more times than provided in the usual 3 × 3 calling format. To do this from the memory, load his call just once at the message starting address and pause until the bit indication stops advancing. Then continue with *DE*, your own call several times and *AR*. In the SEND mode, hold the message selector button down continuously instead of pressing and releasing it. The called station's call sign will be repeated as long as you hold the button down. When you release it, the rest of the message will continue to conclusion. (You can send long CQs this way too, but they're not recommended!) If you want to reduce the number of times your own call is signed, simply manually send *AR* or *K* at the appropriate point to stop the memory.

To call the same station often, as in a DX pileup, switch to the LOAD mode, select a message segment, and proceed to make your first call. Then, in the SEND mode, press the selector button and the memory will do all the subsequent repetitive calling for you. The above are only a few of the possibilities. Once you have gained some familiarity with the unit's capabilities, many other operating applications will suggest themselves.

references

1. James M. Garrett, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August, 1973, page 19.

2. James M. Garrett, WB4VVF, and D. A. Contini, W4YUU, "The Accu-Memory," *QST*, August, 1975, page 11.

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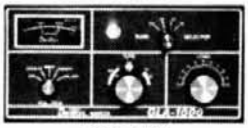
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for RTTY

Construction details for an active bandpass filter which will help eliminate interference and provide improved RTTY copy

This active bandpass filter was designed to work ahead of the NS-1A PLL Demodulator,[1] but it will provide improved copy when used ahead of any RTTY terminal unit. The filter is connected between the audio output of a receiver and the input of the terminal unit.

A bandpass filter is just what the name implies. It will pass only a specified band of frequencies, attenuating all frequencies above and below the specified limits. The objective is to make this attenuation as high as possible, passing nothing other than the desired signals. Without complex circuits, this is not always possible. Simpler filters, like the one described here, will always pass some unwanted frequencies, attenuated enough, however, that they are of no consequence.

For RTTY, we want to pass the standard narrow-shift tones, 2125 and 2295 Hz. For this 170-Hz shift the bandpass should be about 200 Hz wide, centered on 2210 Hz.

circuit description

As can be seen from **fig. 1**, the filter consists of two separate, two-section filters using cascade-connected operational amplifiers. It uses only one IC, the LM3900, which contains four so-called "Norton" amplifiers. These op-amps will do almost anything the standard op-amp will do. The principal advantage is that the LM3900 requires only a single power supply. Also, the LM3900 differences the input currents, whereas the conventional-type op-amp differences the voltages. The noninverting input function has been made possible by using what is called a current mirror circuit.

Several formulas are used to determine the required values of the resistors and capacitors. The

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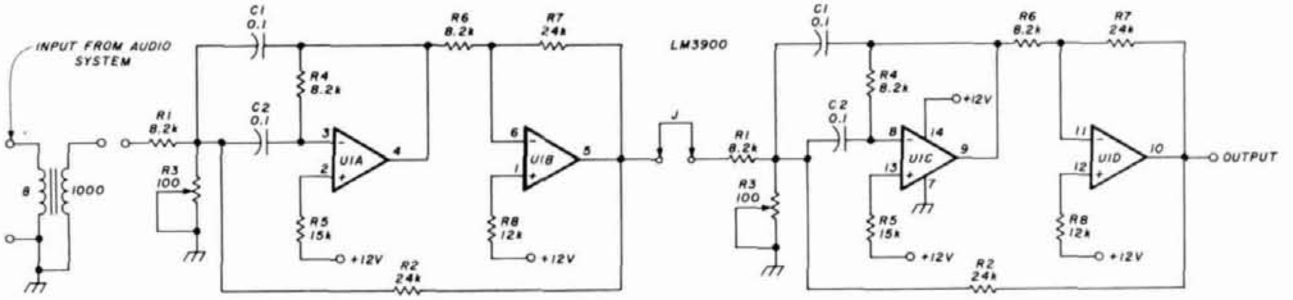

fig. 1. Schematic diagram of the active bandpass filter. The center frequency is 2210 Hz, with a −3 dB bandwidth of approximately 160 Hz. A single LM3900 Norton op-amp is used for the filter.

nearest standard values of 5 per cent resistors are shown in the diagram. The capacitors can be 10 per cent tolerance. R2 sets the *Q* of the circuit (and also the gain) and works out to be 12.5 kilohms for a *Q* of 11, or a −3 dB bandpass of 200 Hz centered on 2210 Hz. When both filters are connected together, however, the increased *Q* makes the bandpass too narrow. The value of 24 kilohms was chosen by trial and error to give the best passband.

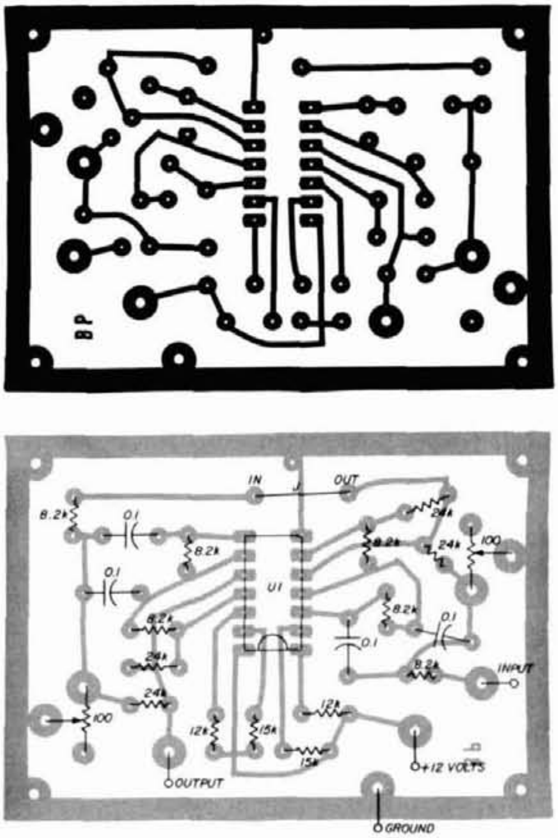

fig. 2. A circuit board foil pattern for the active bandpass filter is shown above. The filter can also be built on a small piece of perforated board using 5 per cent resistors and Mylar capacitors.* *The bottom illustration is the parts placement diagram for the circuit board.*

The complete filter, with standard resistor values, actually gives a −3 dB passband of 160 Hz, or a *Q* of 14. This may seem a little narrow for 170-Hz shift, but on-the-air tests have shown this not to be the case. The bandpass can be easily widened by changing R2 or by stagger-tuning one filter 50 Hz below and the other 50 Hz above the center frequency. The filter turns out to be about 800 Hz wide at the −20 dB points. There is no insertion loss; in fact, there is a small gain.

alignment and connection

Before connecting the jumper between the two filters, separately tune each filter to 2210 Hz. This can be done by injecting a 2210-Hz signal and adjusting R3 for peak output as shown on a VTVM or high-impedance ac voltmeter. The jumper is then connected between the sections. Another method of tuning the filters, though not as accurate, is to tune in a strong RTTY signal with the filter in the line and adjust R3 on each filter for peak output or best copy.

The transformer is used to step up the input from the receiver speaker output to more closely match the input impedance of the filter. If the output impedance of your receiver is at least 500 ohms, the transformer will not be needed.

A good way to check the performance of the complete filter is to have some means of switching it in and out of the circuit. You will find that the filter will sometimes be the difference between copy and no-copy. High-impedance headphones (2000 ohms) connected across the output will enable you to hear the difference.

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*A complete kit of parts and wired/tested units are available from the author. See Flea Market ads.

reference

1. Nat Stinnette, "Update of the Phase-Locked Loop RTTY Demodulator," *ham radio*, August, 1976, page 16.

ham radio

new audio amplifier for the Drake R-4C

A new audio amplifier for the Drake R-4C, suitable for direct substitution in all R-4C versions

Improvements in the Drake R-4C receiver, up to now, have been confined mainly to the i-f and detector systems.[1,2,3] One remaining area which needs improvement is the audio strip, which suffers from buzz and higher-than-desirable distortion; it also dissipates 7 to 10 watts of heat near the PTO. The audio amplifier, diagramed in **fig. 1**, eliminates these problems. While intended as an R-4C retrofit, this circuit performs so well that we also recommend it for other communications uses.

Our circuit is designed around National Semiconductor's LM383T, which, with the R-4C low-voltage supply, can deliver in excess of 2 watts into a 4-ohm load. The LM383 and associated components* can be mounted on a copper-clad board 3.8 cm (1½ inches) square, or another appropriate small heatsink (for a V_{cc} of 16 volts or less). It should be installed just behind the front-panel phone jack, between the passband-tuning capacitor and long i-f shield on which the Sherwood CF-600/6 may be mounted. This location provides access to the speaker lead and detected signal at the audio gain pot. It also keeps the circuit away from power transformer hum fields in the chassis.

circuit precautions

The secret of making the LM383 an unconditionally stable audio amplifier (suitable for field installation in various layout configurations) is our output stabilization network. Proper stabilization is accomplished by connecting a 1.0-μF monolithic ceramic capacitor (such as Sprague 5CZ5U105X0050C5) with 19-mm (3/4-inch) leads directly between pins three and four of the LM383. Use of a lower-value capacitor with significantly longer *or* shorter leads will virtually guarantee oscillation problems. Tantalum or aluminum electrolytics *cannot* be substituted for the monolithic capacitor.

Other circuit values have been chosen to tailor the audio response for greatest communications intelligibility. As in the original R-4C circuit, low frequencies are rolled off at one end of the needed spectrum; high-frequency shaping is similar to that of our suggested modification.[1] The feedback network has been chosen to provide nearly 40 dB of power-supply ripple rejection, minimizing the need for abnormal amounts of filtering. Gain at 1 kHz is 40 dB.

component selection

As with any high-gain amplifier, feedback and hum loops between the input and output should be avoided. Return all signal and power leads to pin 3, except for V_{cc} bypass, which should be returned to the IC tab with a solder lug.

To reduce component size, the 0.22-μF and 10-μF capacitors can be 16-volt (or greater) tantalums. The 200-μF electrolytic at pin 2 can have a 3-volt rating. The 300-μF output capacitor should have a minimum rating equal to V_{cc} (20 volts maximum). Sixteen volts is adequate for the R-4C. As mentioned above, a small heatsink is used for a V_{cc} less than 16 volts; above 16 volts a large heatsink. Never exceed a V_{cc} of 20 volts.

installation

To disable the existing amplifier, lift the output

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*A parts kit will be available from G. R. Whitehouse, Newbury Drive, Amherst, New Hampshire 03031.

transistor's collector lead at its solder lug. Also, remove its base or emitter wire, and/or disconnect one end of the driver's 100-ohm collector resistor. Connect the new amplifier's output to the phone jack terminal *with* the sleeved wire from the audio output transformer, and bypass with a 0.01-μF capacitor. This secondary is still used to provide the needed step-up for anti-vox system.

The only ground return should be a short, thick, insulated wire, run from pin 3 directly to the cable-braid terminal of the audio gain pot (rear section of the dual control.) Do not allow the board or heatsink to touch any other part of the receiver ground. Next, add a small wire between the audio gain-control braid terminal and a close chassis ground. Disconnect the existing wire from the gain pot center wiper terminal, and connect the new amplifier input to this lug. Connect V_{cc} to the original audio-strip printed circuit board terminal with the blue wire from the audio-output transformer primary.

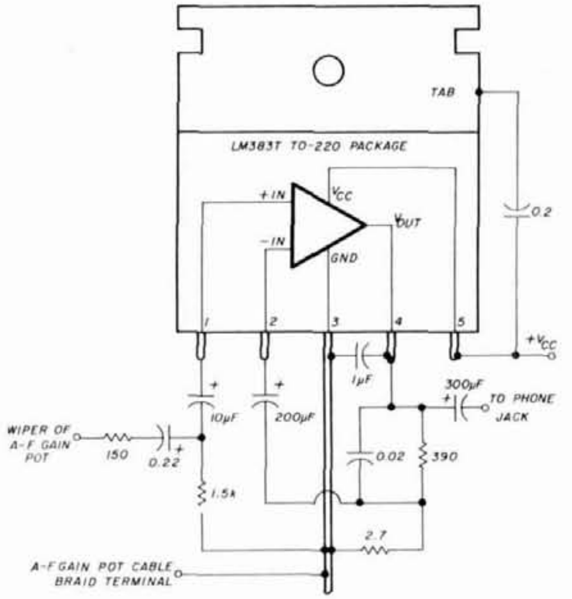


fig. 1. Schematic diagram of the new audio amplifier for the R-4C, based upon the LM383T audio amplifier IC. Resistors are ¼ watt. As pointed out in the text, all ground returns *must be made* through the connection to pin three to eliminate hum and feedback problems. The 1-μF monolithic ceramic capacitor *must* have 19-mm (¾-inch) leads (see text).

references

1. R. Sherwood, WB0JGP, G. Heidelman, K8RRH, "Present-Day Receivers — Some Problems and Cures," *ham radio*, December, 1977, page 10.

2. R. Sherwood, WB0JGP, G. Heidelman, K8RRH, "New Product Detector for the R-4C," *ham radio*, (ham notebook), October, 1978, page 94.

3. R. Sherwood, WB0JGP, G. Heidelman, K8RRH, "New Product Detector for the R-4C," *ham radio*, (short circuits), February, 1979, page 94.

ham radio

the verti-loop a folded whip antenna for vhf mobile operation

Do you want a simple, inexpensive way to improve your 2-meter mobile performance in the fringe areas? How about a 3/4-wavelength vertical on your car roof? The concept sounds mind boggling — but what if it were only 1 meter (3 feet) long? Sound interesting? Try building one of these verti-loops, a name coined for a 3/4-wavelength vertical ground plane compressed to less than 1/2-wavelength long by folding the bottom section into a horizontal loop. The advantage is 2-3 dB gain over the 1/4-wavelength whip, and it still lets you keep the car in the garage. It's an easy weekend project and should cost less than $3.00 if you already have a roof-mounted antenna or an old discarded mag-mount CB whip. The results will be well worth the effort.

theory

Most vhf mobile operators know that a 1/4-wavelength whip is ideally located in the center of an unobstructed car roof. A compromise, because of physical size, is to use a 5/8-wavelength gain antenna on the trunk lid, where, because of obstructions on most cars, the 2-3 dB of potential gain is usually lost; results may even be inferior to those of the smaller roof-mounted whip.

A 3/4-wavelength ground-plane antenna is resonant, therefore nonreactive, and has a high-current feedpoint. The length of such an antenna may be calculated as half the length of a 3/2-wavelength dipole[1]:

$$L = \frac{149.95\,(N - 0.05)}{f} \qquad \textbf{(1)}$$

where L is the 3/2-wavelength dipole length in meters, N is the number of half wavelengths on the antenna (3), and f is the frequency in MHz (146).

$$L = \frac{149.95\,(2.95)}{146}$$

$$= 3.03 \text{ meters } (119.3 \text{ inches}).$$

The vertical length would be half of this, or 1.52 meters (59.6 inches). The antenna impedance would be half of the 3/2-wavelength antenna impedance (105 ohms), or 52.5 ohms.

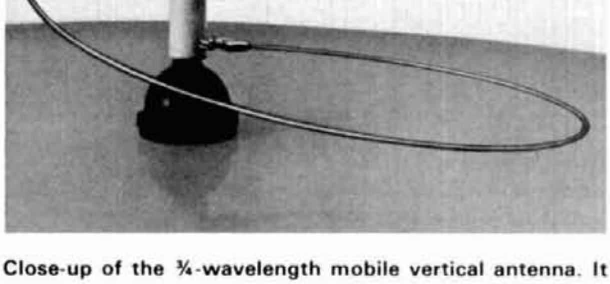
design

To shorten the 3/4-wavelength vertical to a reasonable size, it was divided into 1/4- and 1/2-wavelength sections, and the 1/4-wavelength section was folded into a loop. The result was the upside-down "halo" antenna depicted in **fig. 1**. The top section was shortened slightly to 91 cm (36 inches), and the bottom section was lengthened slightly to 61 cm (24 inches) for convenient fabrication sizes. When folded into a loop, the bottom section results in a 19-cm (7.5-inch) diameter.

construction

The antenna was constructed very simply using a 2.4 mm × 91 cm (3/32 × 36 inch) stainless-steel welding rod (available at any welding supply shop) as the 1/2-wavelength section, and an ordinary tv uhf loop (available from Sears) as the folded 1/4-wavelength section.

To one end of a 5-cm (2-inch) section of plastic tubing (PVC works fine), a threaded stainless steel nut, which matched the threaded end of my rooftop 1/4-wavelength antenna, was epoxied. A piece of wood dowel (or plastic filler) was cemented into the other end of the tube and was drilled for a snug fit for



Close-up of the ¾-wavelength mobile vertical antenna. It resembles an "upside-down" halo.

By Herb Bresnick, WB2IFV, 16 Creekside
Drive, Honeoye Falls, New York 14472

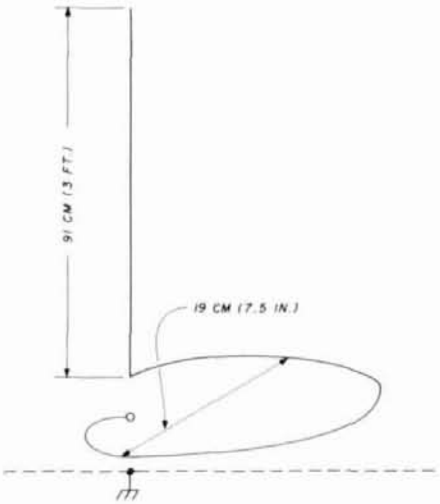


fig. 1. Design of the verti-loop antenna. Antenna is a ¾-wavelength vertical divided into ¼- and ½-wavelength sections; the ¼-wavelength section is folded into a loop. Total extended length is ¾ wavelength on 2 meters. Antenna mounts on your car roof with a plastic base made of PVC tubing.

the stainless rod. A little epoxy will keep it from loosening (**fig. 2**).

The Sears loop comes with hinged terminals, which were left on to permit fine tuning the antenna. One wire terminal was simply pulled out of its hinge, leaving a spring opening through which the stainless rod was forced. It makes for a snug fit, but permits sliding the loop down to where the rod joins the plastic tube.

The other wire terminal on the loop was cut off (leaving the hinge) and the wire remaining was passed through a small hole drilled in the side of the PVC tube just above the nut, where it was compressed by the threaded roof stud when the antenna was screwed on.

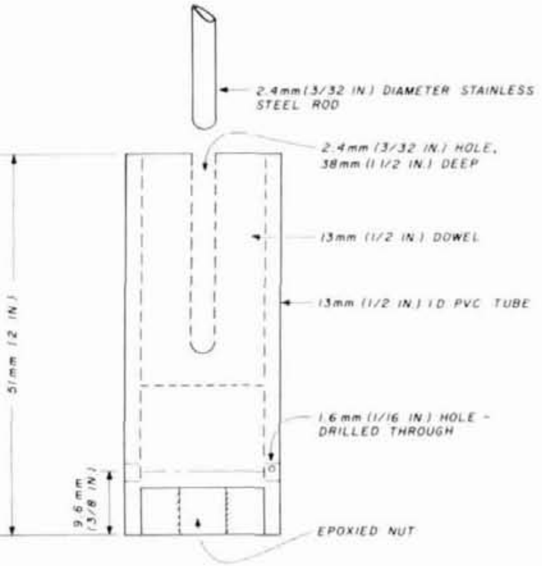
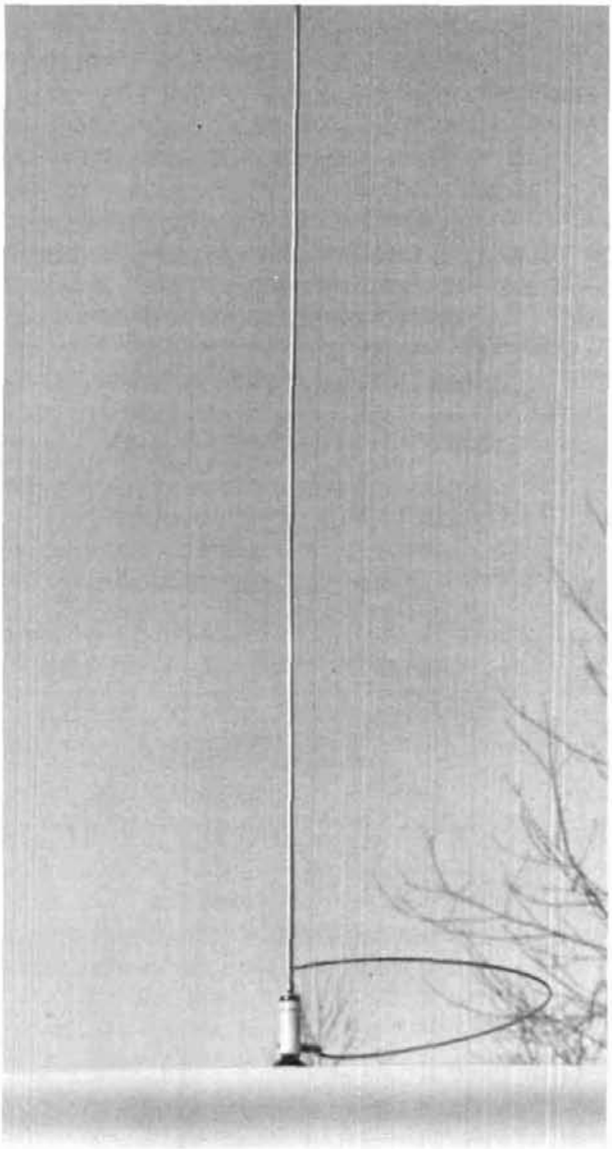


fig. 2. Details of the mounting base. An ordinary tv uhf loop was used for the ¼-wavelength section, or you can use stainless-steel wire.

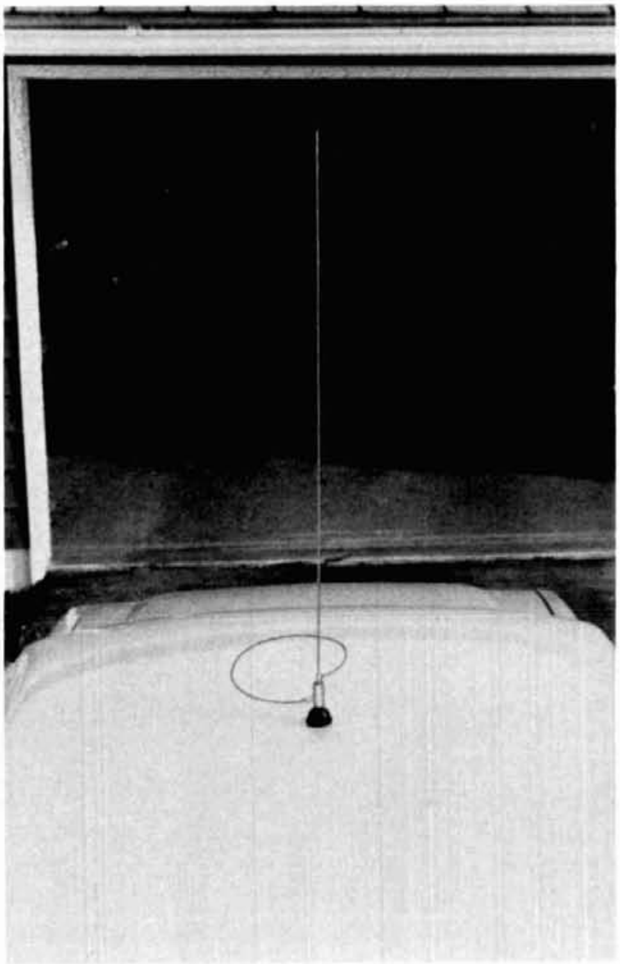
The antenna rod should be insulated from the stud except through the loop connection. The design can be modified easily to fit other base mounts or for new installations using a PL259 coax fitting, or even to a discarded CB magnetic mount if you don't want to drill a hole into your car roof.



View of the antenna at car-top level.

results

With the transmitter on an unused frequency and a SWR meter attached, the loop position was moved on its hinges for minimum SWR. It's not a critical adjustment if the dimensions are followed. I obtained 1.1 to 1.0 SWR with the loop about 2.5 cm (1 inch) from the car roof with the first try. The hinges are stiff enough to maintain the loop position even when



Author's antenna is a full-fledged ¾-wavelength vertical but fits under garage opening without dismantling. The stainless-steel radiator bends easily to 30-40 degrees from vertical.

driving over bumpy roads. I've built three of these antennas at a cost of $2.00-$3.00 each. All have resonated with practically no adjustment.

Actual dB measurements on the antenna have not been made; however, tests indicate a marked improvement over the 1/4-wavelength whip. One repeater that was barely readable is now almost full quieting into my receiver. Reports from friends are that noise on my signal (from a portable HT in the car) was all but gone and no evidence of mobile flutter occurred.

Best of all the car still fits into my garage, and, although the top of the antenna does hit the entrance, the stainless-steel rod is flexible enough to recover from a 30- or 40-degree bend. The antenna has a somewhat futuristic look to it — at least it looks different from your good buddy's 11-meter job!

reference

1. *The ARRL Antenna Book*, 12th edition, 1970, page 33.

ham radio

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VSWR at resonance	1.3:1	**Wind loading @ 80 mph**	114 lbs.
Impedance	50 ohms	**Assembled weight (approx)**	37 lbs.
F/B Ratio	20 dB	**Shipping weight (approx)**	42 lbs.
Boom (O.D. x length)	2" x 14' 4"	**Direct 52 ohm feed or balun**	
No. of elements	3	**Maximum wind survival**	100 mph
Longest element	27' 4"		

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Gain	11 dB	**Turning radius**	37.13"
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Impedance	50 ohms	**Surface area**	44 sq. ft.
Boom (O.D. x length)	1" x 64"	**Wind loading @ 80 mph**	5.5 lbs.
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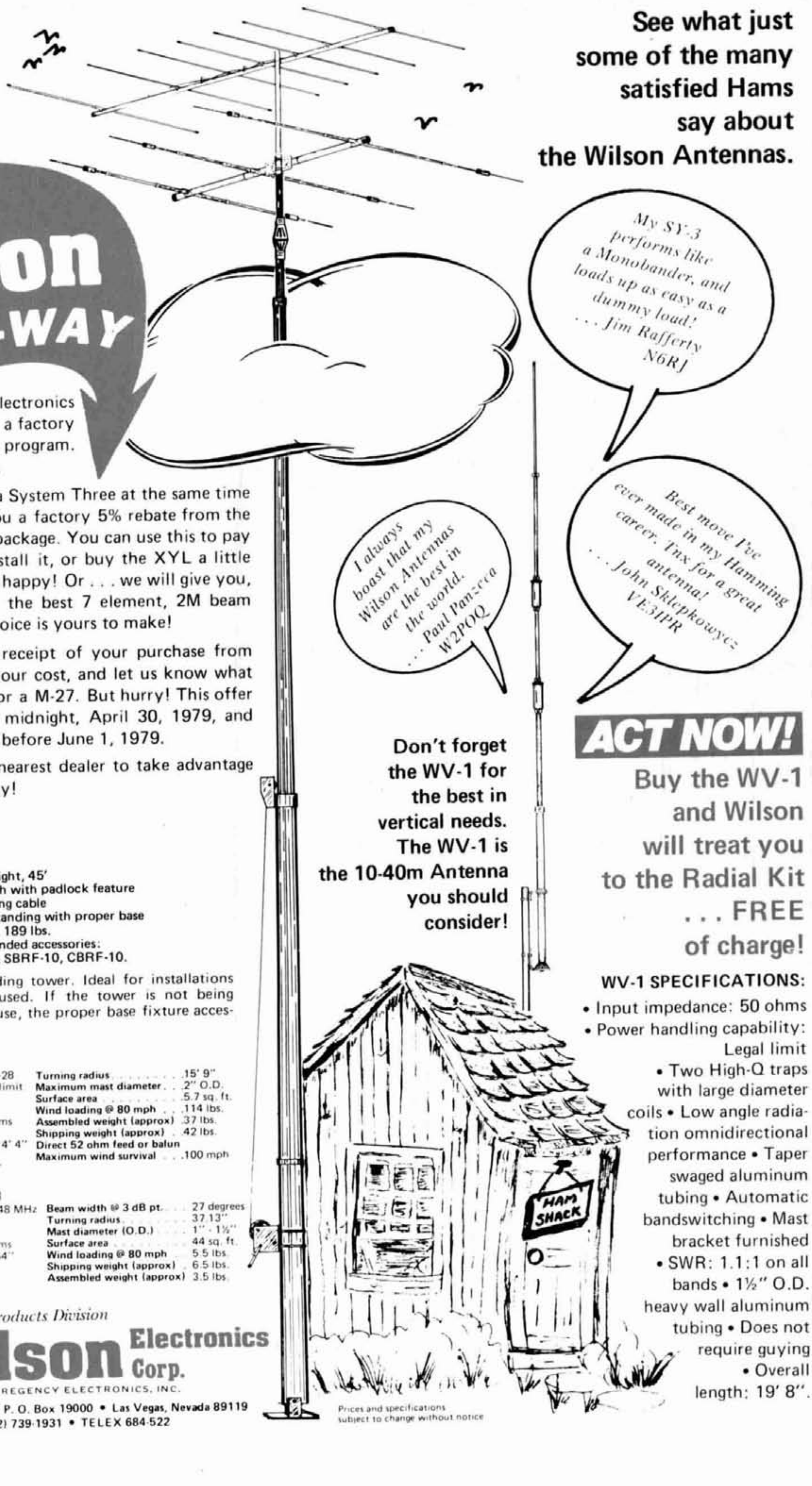

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Ω	.2 to 20mΩ Range x 1 x 10 x 1k x 10k	± 3% arc
dB	+ 10db~+22db for 10VAC	± 4% fs
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the jammer problem:

some interesting solutions

The incidence of jammers is increasing — here's an interesting approach to the problem

In this paper we use the basic ideas of the scientific method in solving some interesting problems. It is important to have a clear understanding of, and agreement on, the nature of observation, cause and effect, hypotheses and testing of these hypotheses.

Science begins with observation. The scientific worker uses his mind to imagine how something might be constructed or how a problem may be solved. He knows that only by examining his observations can he determine if his imaginings correlate with the real world.

statement of the problem

The locale is a large city in the United States. In this city many Radio Amateurs abound. The problem we are about to address (and solve using the scientific method) concerns specific specimens. These specimens are Radio Amateurs (and others who are not legitimate Radio Amateurs) who persist in jamming the 2-meter repeater stations in this city. These specimens particularly delight in causing a disruption in radio communications over the repeater links by using various schemes.

The problem is to locate these specimens using three scientific methods: mathematical methods, methods from theoretical physics, and methods from experimental physics.

mathematical methods

Several mathematical methods are available to us to ferret out the offending 2-meter jammer. The mathematical methods we shall enumerate are easily seen to be applicable, with obvious formal modifications, to similar situations in other parts of the globe. As with other branches of knowledge to which mathematical techniques have been applied in recent years, the mathematical techniques of ferreting out 2-meter ham jammers has a singularly unifying effect on the most diverse branches of the exact sciences.

Hilbert, or axiomatic method

We place the culprit in an imaginary cage, which we have previously defined in time and space by using established techniques. We then introduce the following logic system:

Axiom 1. *The class of jammers in the cage is non-void.*

Axiom 2. If there is a jammer in the city, there is a jammer in the cage.

Rules of procedure. If p is a theorem, and p implies q is a theorem, then q is a theorem.

Theorem 1. There is a jammer in the cage.

Conclusion. Approach the cage with caution. Lift the jammer gently out of the cage, shave his beard, cut his mike cord, and put him in a playpen.

By Nuryev Sidelbandsk, UX3PU

method of inverse geometry

We place a spherical cage in the city, enter it, and lock the cage. We then perform a geometric inversion with respect to the cage. The jammer is then inside the cage and we are outside.

Conclusion. We proceed as in the method of Hilbert, *above*, except in this case we call in federal troops with Zen guns, who quickly dispose of the specimen. We then donate his equipment to a worthy organization.

"Call in the troops."

"Theorem 1 — there is a jammer in the cage."

method of projective geometry

Without loss of generality, we may regard the city as a plane. We project the plane into a line, then we project the line into an interior point of the cage that contains the jammer. We then perform a geometric involution of catharsis, which in turn projects the jammer into the city jail (without bail).

Bolanzo-Weierstrass method

In this method we bisect the plane by a line running north-south. The jammer is either in the east portion or in the west portion; let us assume he is in the west portion. We bisect this line with another line running east-west. The culprit is either in the north portion or the south portion. Let us suppose he is in the north portion.

We continue this process indefinitely, constructing a sufficiently strong force about the chosen portion at each step. As the volume of the chosen portion approaches zero, the jammer is ultimately surrounded by a fence of arbitrarily small perimeter. The specimen (jammer, if you will) is thus a victim of the confusion syndrome. He begins to believe he has, at last, become rational. It is now relatively easy to approach the jammer. We place a cowbell into the jammer's left hand and into his right hand we place a telegraph key, which is connected to a code-practice oscillator. It is now a very critical moment for the jammer. He begins to send messages to himself with the cowbell and answers himself with the telegraph key. We realize that the messages make no sense because the jammer is not comfortable with the Morse code.

Conclusion. To reduce the confusion factor in the jammer's mind, we gently tap him on the head with a small mallet. As we manipulate the mallet (which we do using the Morse code), the jammer begins to get the message, to wit: "Jamming is a no-no. Jamming is a no-no. Jamming is a no-no." I shall leave as an exercise for the *student* the decision as to how long we subject the jammer to this action.

Cauchy, or function-theoretical, method

In this mathematical method we consider a jammer-valued function, $f(z)$, where z is nondimensional but critical to resolving the problem.

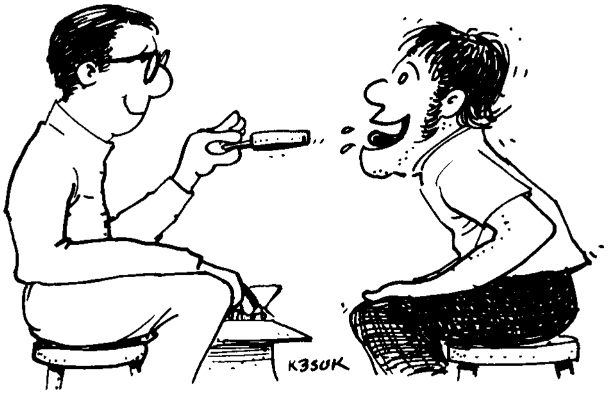
Let there be a function, $f(z)$, and let S be the cage

"The jammer is not comfortable with the Morse code."

enclosing the jammer. Consider the integral

$$\frac{1}{2\pi 1} \; {}_{0}\int^{C} \frac{f(z)}{z-s}\, dz \qquad (1)$$

where C is the boundary of the plane of encounter. Its value, after appropriate manipulation of **eq. 1**, is that of $f(S)$. Therefore, we have a culprit in the cage. We then dispose of him appropriately. In this case we ignore his stupid mouthings on the repeater but send a male nurse armed with ten-dozen Popsicles. The nurse plies the jammer with Popsicles until the jammer decides to throw up (a very effective problem-solving method).



"Ply him with popsicles."

methods from theoretical physics

We now examine some theoretical methods for dealing with the 2-meter jammer. These methods have not been proved in practice but offer some interesting food for thought for the theoretician.

Dirac method

We observe that jammers are known to use wild methods to attract attention. Therefore these jammers must be considered wild men. But, according to the scientific method, we must temper our observations with caution. Perhaps the jammer in question is not really wild but is somewhat tame with a rather limited mentality. Perhaps he needs help. Why punish him if, indeed, he's not really responsible for his actions? The author has a proven formula which, if applied to these situations, will produce good results in many cases. For those interested I shall be happy to furnish the method and appropriate action. Please send a self-addressed, stamped envelope for an immediate reply.

"Perhaps the jammer is not really wild."

methods from experimental physics

Using the methods from experimental physics, we are able to deal positively with the problem of 2-meter jammers. It should be noted that these methods are also valid for other inconsiderate jammers in the other Amateur bands.

thermodynamic method

We construct a semipermeable membrane, permeable to everything except jammers, and sweep it across the plane of interest (*i.e.*, the city in question). We also add a pot of hot water to the apparatus. As the semipermeable membrane is swept across the city, it picks up the jammers. The jammers are enmeshed in the membrane and, having nothing else to do, immediately start brewing coffee. We immediately detect the aroma of brewing coffee, descend on the culprits, and capture them in a loony net. We then enlist the aid of local authorities to dispose of the problem.

magnetomatic method

In implementing this technique we plant a large bed of popcorn arranged in an ellipse, whose major axis lies along the direction of the horizontal component of the earth's magnetic field. At one focus of this ellipse we place a cage (with strong iron bars). We then distribute over the plane in question large quantities of magnetized spinach (*spinacia oleracea*), which has high ferric content.

The magnetized spinach is eaten by various denizens of the locality, which are in turn eaten by the 2-meter jammers. The jammers, having partaken of commodious amounts of popcorn, are then oriented parallel to the earth's magnetic field. This causes a beam of jammers to be focused onto the cage. Since the jammers are full of our specialized popcorn and, by indirect ingestion, magnetized spinach, it is an easy matter to zap the jammers. The benefit of this method is that we can dispose of the culprits in large numbers.

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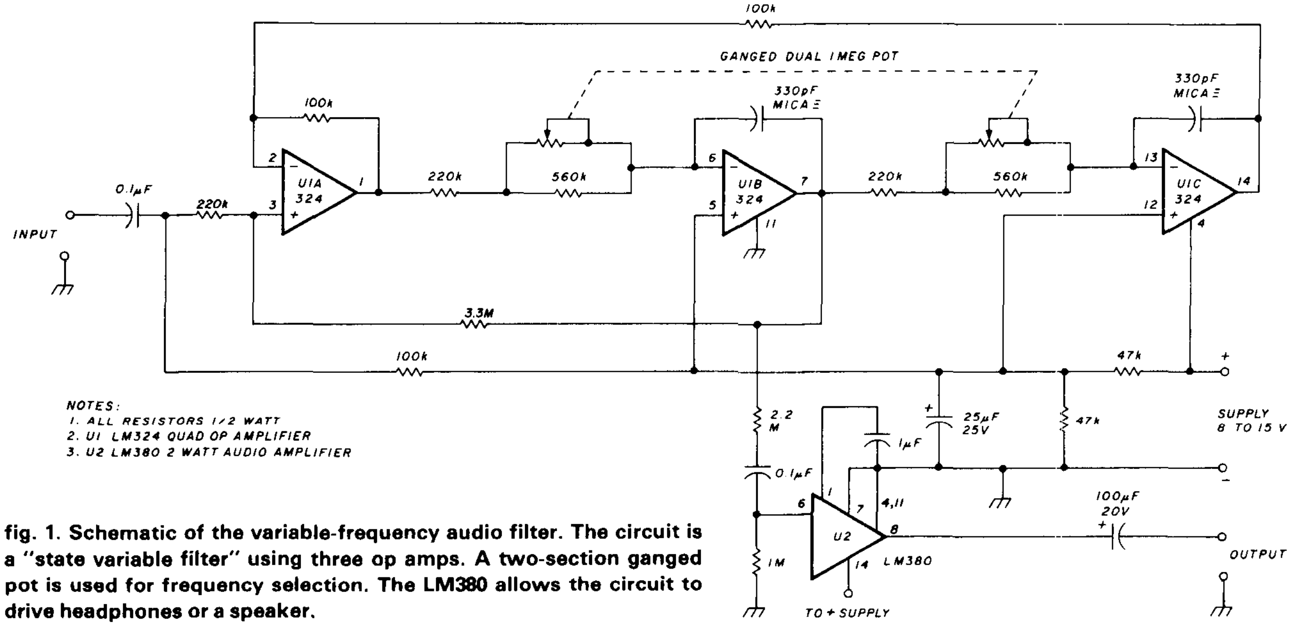
variable-frequency audio filter

Here is a tunable, inexpensive, two-IC audio filter that plugs into the receiver phone jack. It will drive headphones or a low-impedance speaker up to two watts if necessary. It operates from a single power supply that requires no regulation, thus it could be easily incorporated into some of the direct-conversion receivers used in QRP rigs.[1,2] It will also add to the selectivity of any receiver that lacks a sharp CW filter. Used with an ssb receiver, the tunability of the filter allows you to eliminate an unwanted signal by using the high or low skirt of the ssb filter and then peak the desired signal at whatever frequency results.

The circuit is that of the state-variable filter,[3] using three op amps (**fig. 1**). The LM324 IC used in this design has four op amps on the same chip, so the extra amplifiers don't add much to the cost or wiring complexity. Two identical integrator circuits control the frequency response, so only a two-section ganged pot is necessary for frequency selection. The Q, or sharpness, of the filter is controlled by a single resistor. The 3.3-megohm value shown in **fig. 1** was selected as a compromise between sharpness and ringing. Circuit Q comes out to 15. While no ringing is noticed, the noise definitely takes on the pitch of the frequency selected. The value of this resistor may be changed to suit your needs without altering the tuning range. The LM380 was added to the filter circuit to drive hi-fi headphones or a loudspeaker directly; the filter will drive almost anything that the receiver will drive.

With the values shown, the tuning range is from 800 Hz - 2 kHz. Voltage gain is close to unity overall, so the receiver gain control sets the signal level.



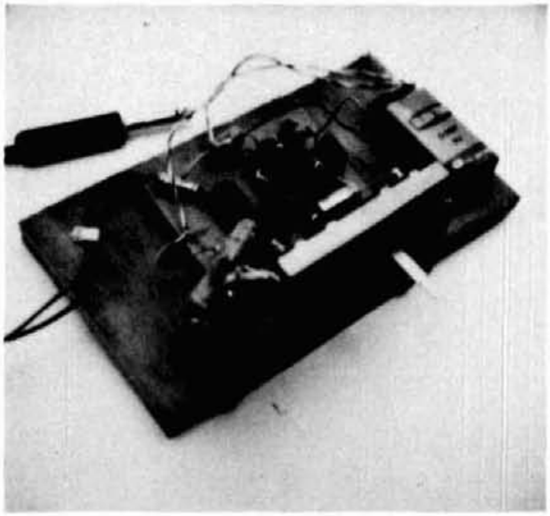
NOTES:
1. ALL RESISTORS 1/2 WATT
2. U1 LM324 QUAD OP AMPLIFIER
3. U2 LM380 2 WATT AUDIO AMPLIFIER

fig. 1. Schematic of the variable-frequency audio filter. The circuit is a "state variable filter" using three op amps. A two-section ganged pot is used for frequency selection. The LM380 allows the circuit to drive headphones or a speaker.

audio-frequency characteristics

The frequency-controlling network is designed to operate with a dual 1-megohm pot. In one model that was built, a dual linear-action slide pot was used and its action was very pleasing. Two single slide pots could probably be mechanically linked to give the same effect. The extra resistors connected in series and parallel with the ganged pots give audio bandspread by allowing the full resistance change of the ganged pots to alter the frequency over less than two octaves. They also make tracking of the ganged pots less critical. If a pot with a logarithmic taper is used,

By Lew Fitch, W4VRV, 109 Robin Street, Clemson, South Carolina 29631



In this model of the variable-frequency audio filter, a dual, slide action linear potentiometer has been used. Because of the high current drain, the 9-volt battery will have to be replaced with an ac supply for permanent installation.

it should be connected for the most linear frequency variation.

If you want to design for a different frequency range, or use a different value of dual pot, the filter peak frequency is given by $f = 1/2\pi RC$. Plugging in the maximum and minimum resistance values will give the frequency extremes. For example, in **fig. 1** when the 1-meg pot is at a maximum, total resistance between pins 1 and 6 is 220k in series with 560k, paralleled by 1 meg. This is 578k ohms, and gives 812 Hz in the formula. When the 1-meg pot is at zero resistance, the 560k resistor is shorted and the resistance in the formula is 220k. This gives 2192 Hz. These values are all for capacitors of 330 pF.

The circuit draws about 11 mA at 12 volts with no signal. This current increases to about 35 mA for fairly loud signals. It would be possible to operate the filter from a 9-volt transistor battery, but not for long periods of time. The feedback loops in the circuit and the 47k divider resistors stabilize the operating point of the integrators at half the supply voltage, so regulation isn't necessary.

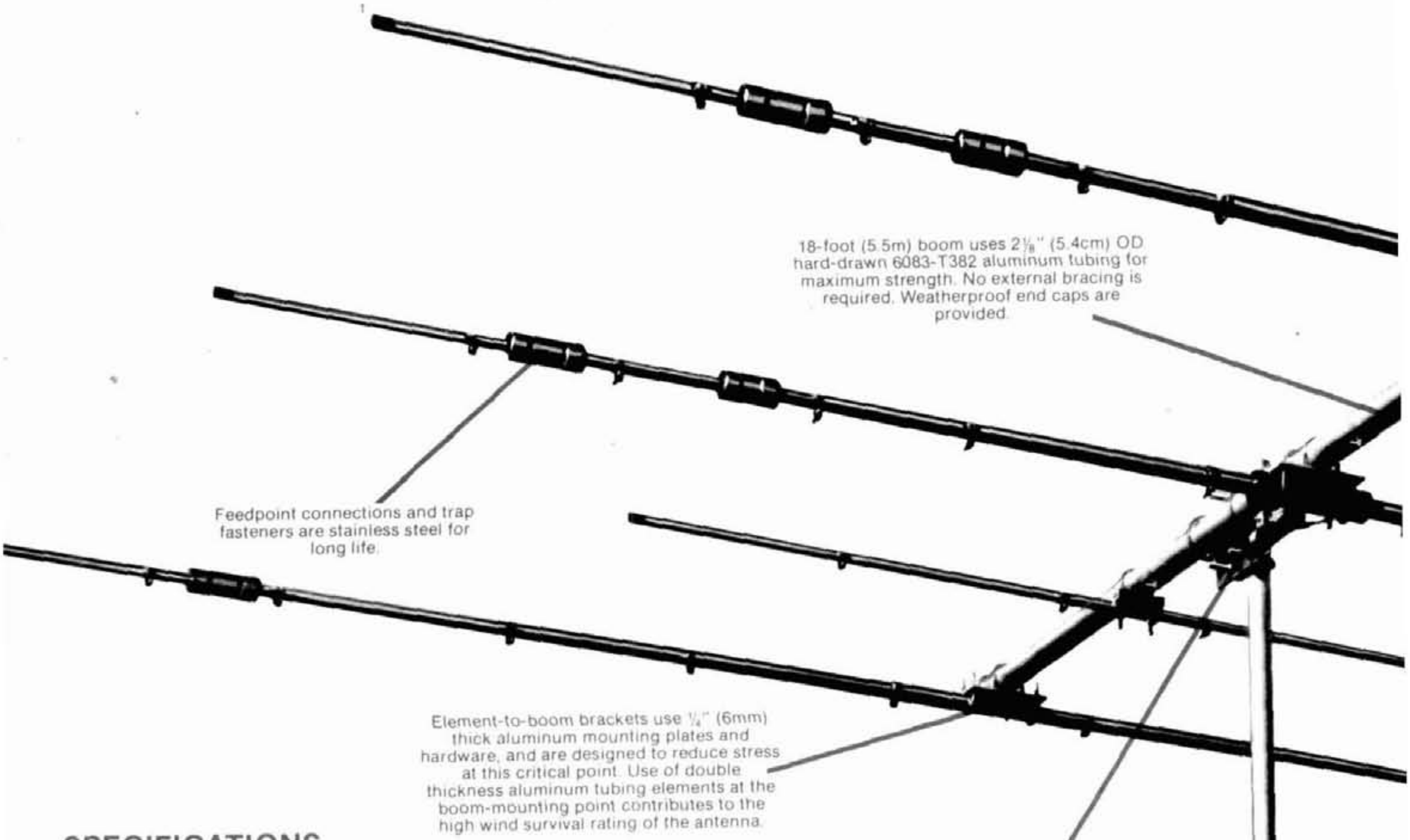
This filter is not the solution to all interference problems, but it will allow you to selectively boost one frequency 10 dB more than another only 200 Hz away. That can make the difference between a useful contact and a marginal one.

references

1. Jay Rusgrove, W1VD, "A 20-Meter High-Performance Direct-Conversion Receiver," *QST*, April 1978.
2. "A Direct Conversion Kilogram," *The ARRL Handbook*, 1977.
3. Howard Berlin, W3HB, "The State-Variable Filter," *QST*, April 1978.

ham radio

HF Cushcraft 3-band antennas may be imitated, but their erformance won't be duplicated.



18-foot (5.5m) boom uses 2⅛" (5.4cm) OD hard-drawn 6083-T382 aluminum tubing for maximum strength. No external bracing is required. Weatherproof end caps are provided.

Feedpoint connections and trap fasteners are stainless steel for long life.

Element-to-boom brackets use ¼" (6mm) thick aluminum mounting plates and hardware, and are designed to reduce stress at this critical point. Use of double thickness aluminum tubing elements at the boom-mounting point contributes to the high wind survival rating of the antenna.

Heavy-duty mast-mounting bracket has been carefully engineered for maximum strength and minimum weight. Use of bright zinc-plated hardware eliminates problems with rust and corrosion. A 2" (5cm) OD heavy-wall mast is recommended.

SPECIFICATIONS

Model	**ATB-34**
Gain 20 meters	7.5 dBd
15 meters	7.5 dBd
10 meters	7.5 dBd
Front-to-back ratio	
20 meters	30 dB
15 meters	22 dB
10 meters	18 dB
3dB Beamwidth	≈ 62°
Normal Input Impedence	50 Ω,
Power Handling	2000 watts PEP
Boom length	18' (5.5m)
Longest Element	32'8" (9.95m)
Turning radius	18' 9" (5.7m)
Wind area	5.4 ft² (0.5 m²)
Weight (assembled)	42 lbs (19.1 kg)
Maximum mast OD	2.5" (6.3 cm)

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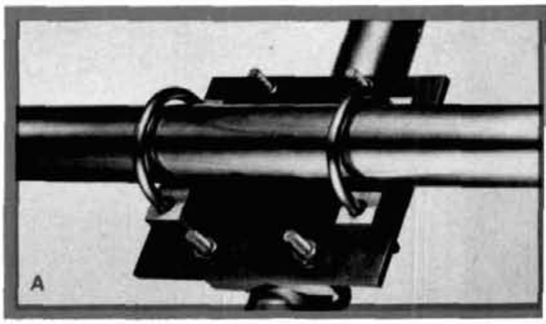
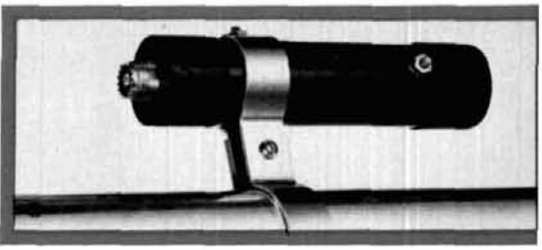
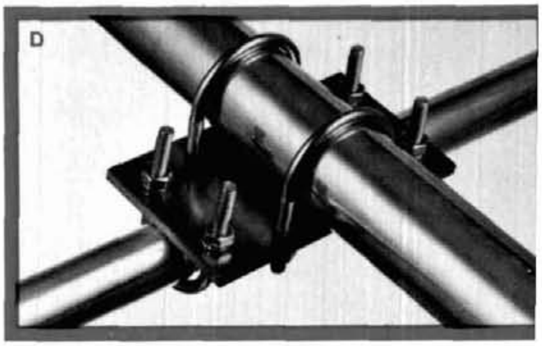
The ATB-34® gives better performance than most single-band beams, yet is easier to install and keep in the air than larger, heavier beams. Forward gain of the ATB-34® is 7.5 dBd on all bands; front-to-back ratio is nominally 30 dB on 20 meters; 22 dB on 15 meters; and 18 dB on 10 meters. Nominal input impedance is 50 ohms; VSWR is 1.5:1 or less at resonance.







A. Heavy-duty mast-mounting bracket for Cushcraft 3-band antenna uses solid aluminum pillow blocks and bright zinc-plated steel hardwa for maximum strength and reliability.

B. High-Q coaxial traps are wound with no. 10 aluminum wire (epox coated) on filament-wound fiberglass forms and have extremely low ohmic losses for longer, full performance elements rated at 2000 watt PEP. Weatherproof covers are provided.

C. Cushcraft 3-band antennas are factory adjusted for a 50-ohm coaxial feedline and come complete with a low-loss, high-performance 1:1 baluns which takes a PL-259 connector.

D. Rugged element-mounting brackets are designed with thick aluminum plates and zinc plated steel hardware to reduce stress and provide trouble-free operation in severe weather. Elements and boom use high-strength hard-drawn 6063-T832 bright-finish aluminum tubing.

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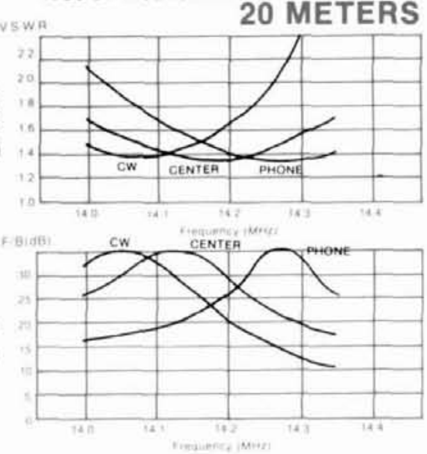
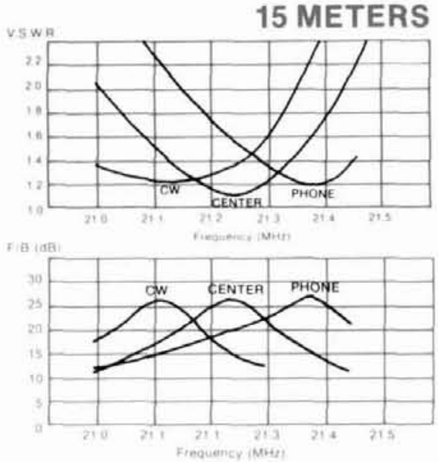
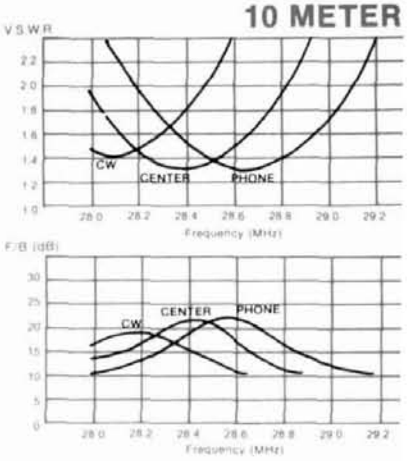






split-frequency operation with the Drake R-4B receiver and the TR-4 transceiver

Modifications you can make to your Drake receiver and transceiver to enhance operating convenience and versatility

This article describes how I increased the versatility of my Drake TR-4 transceiver for split-frequency operation using a Drake R-4B receiver as an external VFO.

I purchased a used R-4B receiver, and the thought kept passing through my mind that it would sure help if I could use it to control the TR-4 transmit frequency (as in the R-4 TX-4 combination).

analysis

The Drake R-4 receiver and the TR-4 use different i-fs; therefore there's no direct interface as with the TX-4. After studying the circuit diagrams, it occurred to me that it should be very easy to duplicate the control circuit of the Drake RV-4 remote VFO and make use of only the VFO portion of the R-4B to control the TR-4.

The control circuit described was built and minor modifications were made to the R-4B to interface with the TR-4. The R-4B can be used as a separate receiver, then switched to remote VFO operation.

Circuit modifications don't interfere with R-4B normal operation. The added components can be removed in a few minutes at resale time to restore the receiver to its original condition.

The main problem in the project is the fact that there's a 45-kHz difference in VFO operating frequency between the TR-4 and R-4B. This is apparently because of the 50-kHz second i-f of the R-4B. There's no reason why the VFO frequency can't be decreased 45 kHz to match the TR-4 VFO output; then the crystals in the premixer oscillator can be changed to oscillate 45 kHz higher to compensate for the VFO downward shift. I breadboarded the circuits to test the idea and everything worked as planned.

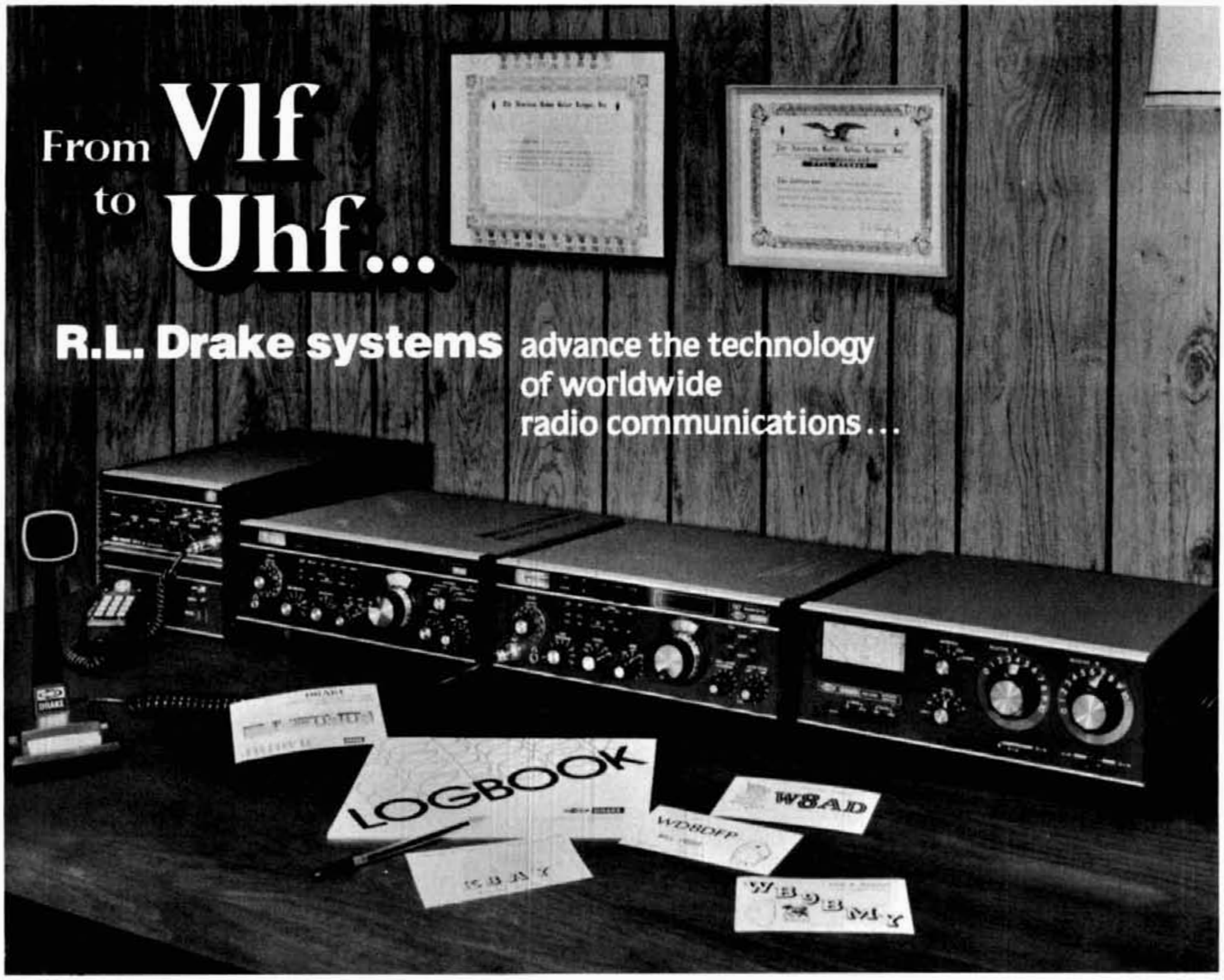
control circuit

I duplicated the control circuit of the RV-4 (**fig. 1**) and mounted it in a small minibox. This minibox sits on top or alongside the TR-4 to interface the R-4B. Only two critical items are in the control circuit. One is peaking coil L1 at the grid of V1 (12AU7); it must match the coax cable from the VFO to the 12AU7 grids.

The diagram in my manual didn't give a value for L1 so I experimented a bit. I reclaimed a vhf coil form from some surplus gear and wound 70 turns of no. 30 (0.25 mm) enamel wire on the 6.5-mm (¼-inch) slug-tuned form. Too few or too many turns would not peak the 5-MHz signal, so the number of turns must be adjusted. A scope is a great aid in tuning L2.

The other critical item is the four-position special switch, A1. After tracing the functions of the Drake switch, I discovered that a standard four-position, five-pole switch would do the job. These parts could be ordered from Drake, but I used all junkbox parts.

By William P. Winter, Jr., WB8JCQ, O'Higgins 3168 Buenos Aires, Argentina



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The 1-mH rf choke from TR1 collector is from an old TV set. Any coil will work as long as it has a good number of turns. Transistor TR1 turns the VFO B+ on and off through the rf choke and the coax line going to the R-4B receiver.

control switching

Position 1, separate receive. The 330-kilohm resistor in **fig. 1** is connected through S1D to B+, turning TR1 on, causing its collector to saturate. Thus TR1 collector goes low, pulling the 10-Vdc B+ line to the R-4B VFO toward zero, which turns the VFO off. The slide switch on the side of the R-4B must be forward to disconnect the control unit, thus defeating the above action. The VFO output is now sent to V8 (premixer) cathode. The R-4B operates normally.

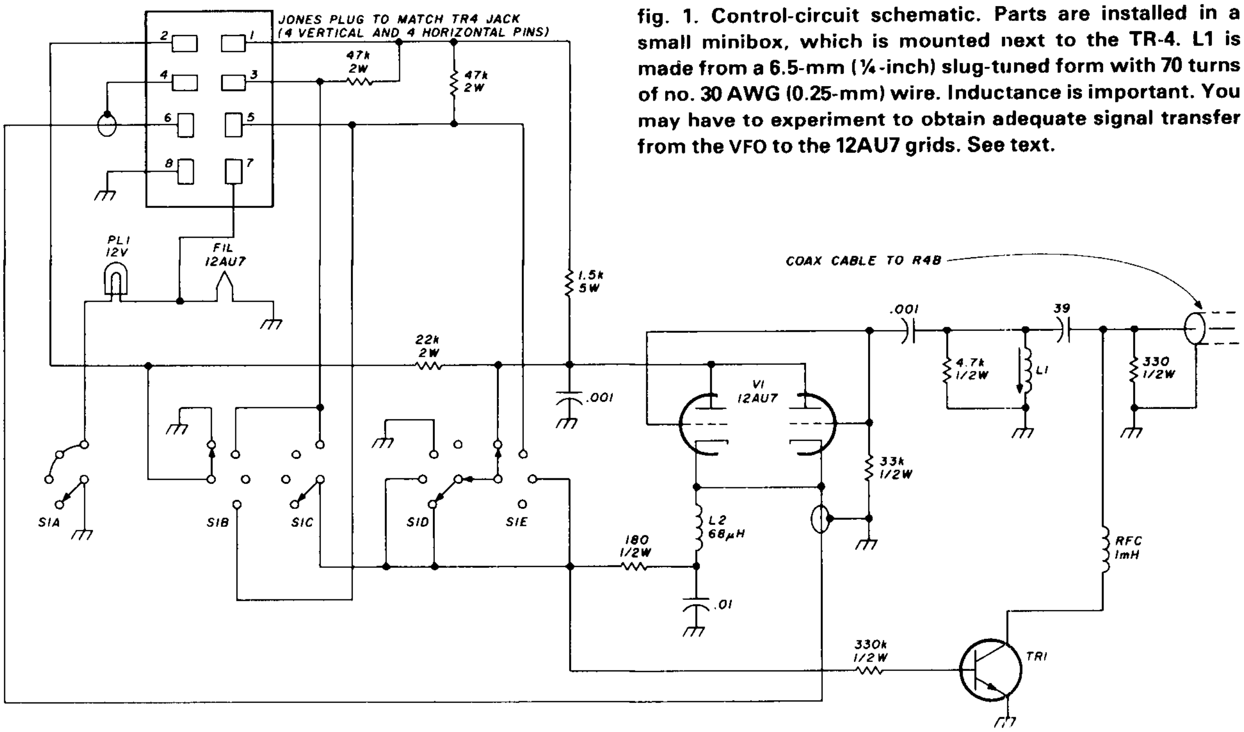

fig. 1. Control-circuit schematic. Parts are installed in a small minibox, which is mounted next to the TR-4. L1 is made from a 6.5-mm (¼-inch) slug-tuned form with 70 turns of no. 30 AWG (0.25-mm) wire. Inductance is important. You may have to experiment to obtain adequate signal transfer from the VFO to the 12AU7 grids. See text.

Position 2, receive. The 330-kilohm resistor is connected through S1D and S1E to pin 5 of the Jones plug. This is receive cathode ground. During receive TR1 base goes low, cutting it off, and the B+ to the VFO goes high, which allows the VFO in the R-4B to operate during receive only. During transmit the receive cathode goes high, turning the R-4B VFO off. Note that the slide switch on the side of the R-4B must be toward the rear to switch the VFO output from V8 mixer to the control unit.

Position 3, transceive. The 330-kilohm resistor is connected through S1D and S1E to ground, turning TR-1 off, thus allowing the VFO B+ to rise to normal. This allows the VFO in the R-4B to operate on transceive.

Position 4, transmit. During transmit only, the 330-kilohm resistor is connected to ground through S1C and pin 3 of the Jones plug to the transmit cathode. TR1 is cut off allowing the VFO B+ to rise, enabling the R-4B VFO. During receive, the transmit cathode goes high, turning on TR1, which pulls the VFO B+ down so that the R-4B VFO can't operate.

The 12AU7 is a cathode follower with both sections in parallel. It functions as an impedance transformer giving a low impedance output to feed the TR-4. At the same time, the 12AU7 acts as a switch to disconnect the signal from the remote VFO. The cathode is switched high or low at the appropriate time according to the switch position and the TR-4 transmit-receive relay.

control-circuit layout

Nothing is critical about the control-circuit layout, since it is primarily composed of dc switching lines. However, some attention should be paid to the rf components related to the 12AU7 grid. Leads should be short, and accepted rf-wiring practice should be used.

Transistor TR1 can be any silicon npn device with a voltage rating of about 40 volts. The coax cable from the R-4B to the control unit, and the cables

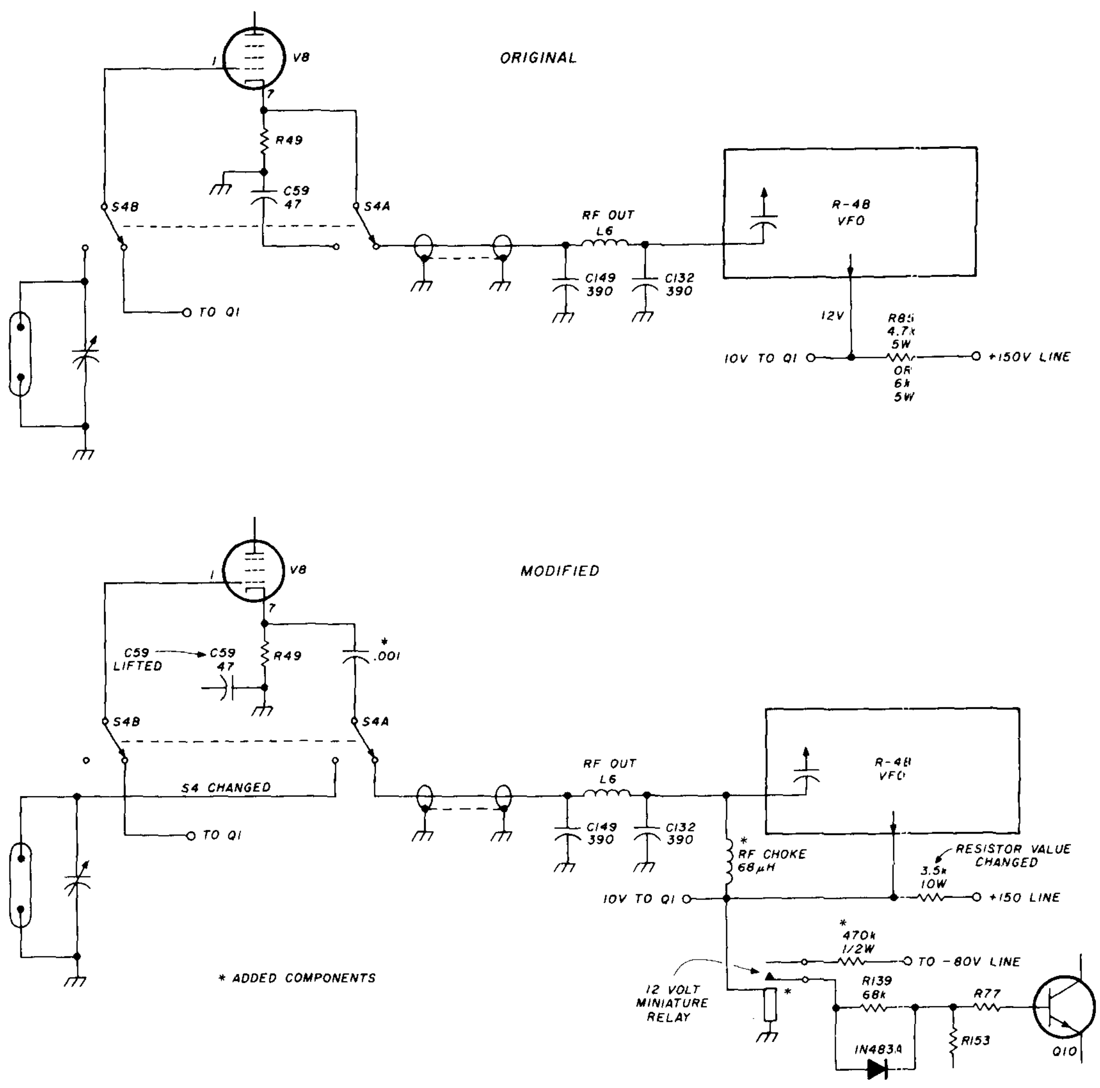

fig. 2. Before and after mods to the Drake R-4B receiver.

from the control unit to the Jones plug, should be as short as possible. *Caution*: Do not insert or remove the Jones plug in the TR-4 while the rig is on; otherwise the TR-4 transistors may be damaged.

R-4B modifications

Of the five modifications described, only the first two are needed to make the R-4B receiver function as a remote VFO. Modifications 3 through 5 are optional. The modifications to be described involve:

1. Changing the value of power resistor R85
2. Changing the position of three wires on S4
3. Adding four components to switch the neon VFO indicator lamp
4. Changing the VFO dial
5. Changing the crystals and adding padding capacitors if desired

Fig. 2 shows the R-4B circuit before and after modification.

Modification 1. Provide VFO output from the fixed-channel crystal jack on the side of the R-4B by following these steps:

1. Lift C59 from S4A. Bend C59 so that it doesn't touch anything. Leave it there for restoration at resale time.

2. Locate the jumper that comes from the fixed-channel crystal jack to S4B. Lift the connection from S4B and solder it to the lug on S4A where C59 was removed.

3. Locate the wire that comes from V8 pin 7. Lift if from S4A middle lug.

4. Lift VFO output coax center conductor from S4A and move it to S4A middle lug (vacated in **step 3** above).

5. Place a small 0.001-μF ceramic capacitor in series with the wire removed in 3 above. Solder the other lead of the capacitor to S4A where the coax formerly was connected. This blocks the 10-Vdc control voltage, which we will place on the VFO output line, from reaching V8.

We have now modified S4 to provide the following switchings: Forward: regular R-4B receive mode. Back: VFO output to crystal jack, receiver disabled. The fixed-frequency crystal socket is now the VFO output jack. A length of RG-59 carries the R-4B VFO signal from this jack to the control box. I soldered the pins of a defunct crystal to the ends of the coax so I could plug the coax into the crystal socket.

Modification 2. Dc switching of the R-4B VFO is accomplished from the control unit by placing an rf choke from the 10-volt VFO B+ line to the rf output line from the VFO. A convenient spot is between the two solder tabs on the circuit board just behind the audio gain control. I used a 68-μH choke from my junk box. The value isn't critical as long as it presents a high reactance at 5 MHz. Check that resistor R85 can handle the additional dissipation when the 10-volt line is shorted to ground by the control box. R85 must drop the entire 150 volts from the +150-volt line. The control unit now enables the TR-4 to turn the R-4B VFO on and off in the same manner as in the remote RV-4 VFO.

Modification 3. Switching the neon VFO indicator lamp is optional but enhances operating convenience. I purchased a surplus 12-Vdc, 10-mA miniature relay and mounted it on a long bolt, which I installed where the audio output transformer is mounted. Your mounting will depend on what relay you have available. (See the comments on relay selection.) The following steps are necessary for this modification:

1. Connect the relay coil from the +10 volt line to ground.

2. Place a 470-kilohm resistor from the normally open contact (normally open with coil energized) to the −80 volt line. A convenient spot is the solder tab that has a white wire with a green stripe (on my R-4B). This is the negative lead of C91, which is the 8-μF, 150-volt filter of the −80 volt line. It's located on a small vertical board behind the notch-depth control, which is mounted on the right-side panel.

3. Locate the board containing Q10. It's mounted below the VFO and just behind the front panel. Connect the other relay contact to the junction of R139 (68k) and D15 anode.

The relay functions as follows. When the TR-4 control box is set to inhibit a signal from R-4B VFO, the +10 Vdc line goes low (3 volts or lower). This action de-energizes the relay just installed, connecting negative cutoff bias to transistor Q10, which causes NE2 to be extinguished.

4. R85, 4.7-kilohm, 5 watts, should be changed to a 3.5-kilohm, 10-watt resistor if you use a 10-mA relay as I did. Extreme care should be used here. With the original 4.7-kilohm resistor, the additional current drain caused the normal 10-Vdc regulated voltage to drop below the zener regulation point. Reducing the value of R85 allows an additional 10 mA of current to be drawn, and the zener will still regulate at 10 Vdc. Do not operate the VFO with the lower resistance value without the relay coil connected, as the additional current will probably blow the 250-mW, 10-volt zener mounted inside the VFO enclosure. Easy does it!

If you use a relay with a coil other than 10 mA, 12 Vdc, adjust the value of R85 accordingly or perhaps provide a separate supply to drive the relay by a transistor. An alternative method would be to put a preregulator on the line (12 to 15 volts) with a zener of enough power-handling capability to do the job.

Modification 4. If you wish dial calibration on the R-4B identical to that on the TR-4, it will be necessary to order a TR-4 dial from Drake. (The price in June of 1978 was $2.00, plus $1.00 handling, plus postage.) It's installed in the following manner:

1. Remove all front-panel knobs (some slip on; others have a set screw).

2. Remove nut on the function selector switch.

3. Remove four screws at corners of the front panel. Be careful to catch the fiber spacers behind the panel (they're hard to find when they fall and roll across the floor).

4. Remove the two screws holding the metal shield over the neon bulb. Be extremely careful *not* to bend the neon-bulb leads, because then break very easily; many standard replacement neon bulbs will not fit. I had to cut a hole in the shield to allow the end of the bulb to stick out.

5. Very carefully remove the front panel without unduly bending the neon-bulb leads. Lay aside the clear Plexiglas sheet with the red line. Do not lose the pressure washer. Note how it came off so it can be replaced in the same way.

6. Remove two screws holding the pilot light shield mounted behind the panel.

7. Remove the two C-rings from the VFO shaft. You'll need a C-ring tool; it can be purchased in auto parts shops. Be careful! The C-rings are tight and are hard to remove. (I broke my tool and had to use a wheel puller to remove the first C-ring). The second C-ring wasn't so tight. *Important*: Note the position of these C-rings so they can be replaced in the exact spot.

8. Locate, identify, (make a drawing), and remove the three leads coming out of the VFO.

9. Remove the three nuts holding the VFO in place. Lift and remove the VFO assembly from the R-4B.

10. Turn the VFO shaft to find the final stop. Noting *exactly* where it was positioned (pencil mark), remove the dial by removing the large C-ring. Place the TR-4 dial in the same position.

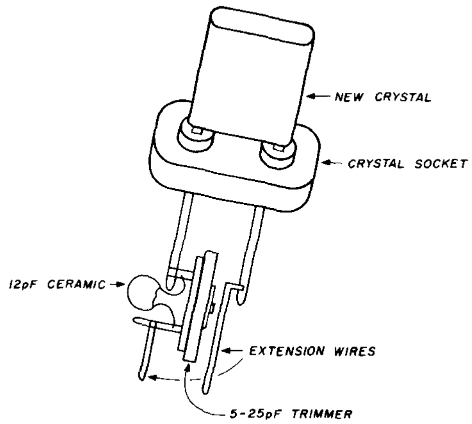

fig. 3. Construction of the crystal paddler, which plugs into the R-4B receiver crystal socket.

11. Reassemble, following the steps above in reverse order.

Modification 5. To track the VFO for TR-4 use and separate receiver use, I ordered new crystals that are 45 kHz higher in frequency than the originals. For the 20-meter band to tune backward, as in the TR-4, I substituted a crystal frequency (14.645 MHz), which is added, rather than subtracted, to the VFO output to obtain the proper injection frequency.

This, by the way, is the same crystal needed for 80 meters. Therefore, by installing an appropriate jumper and one other change, the same crystal can be used for both 20- and 80-meter operation.

To keep expenses down I ordered only three crystals (my cost locally was $20.00 for the three crystals): one for 80 meters, which doubles for 20 meters; one for 15 meters, and one to receive WWV on 15 MHz. Thus I've covered my needs at the present. For the same crystal to be used for both 80 and 20 meters, the following steps are required:

1. Find C53 (68 pF) in parallel with R47 (1.5 kilohms). This network is the collector load for Q1 at 25.1 MHz. Lift these two components from S5F and dress them out of the way so they don't touch anything but are handy for replacement at resale time.

2. Place a jumper from C55 (68 pF) to the switch tab vacated above. Solder the connections.

3. Place an insulated jumper wire from the 14.6-MHz crystal jack to the 25.1-MHz jack. (The 25.1-MHz crystal will no longer be used.) Put it away so it will be handy at restoration time.* The jumper should be placed in the holes toward the front panel. The new crystal, 16.645 MHz, can now be plugged into the 14.1-MHz jack. Eighty-meter operation is as normal; 20-meter operation now tunes backward, as in the TR-4.

crystal padding

I found it necessary to put a trimmer in series with the new crystals to trim the R-4B receive frequency to exactly the same dial calibration as when using the R-4B as a remote VFO. I used a 12-pF ceramic cap in parallel with a small 5-25 pF variable, which I removed from some surplus equipment (see **fig. 3**). The cap is smaller than the standard-size trimmer and will fit between the solder tabs of a standard-size crystal socket. The capacitors are in series with one lead of a crystal socket adapter, which is plugged into the main crystal socket.

dial calibration

Dial calibration is as follows:

1. Calibrate the TR-4 dial as usual, using the TR-4 internal crystal calibrator.

2. Switch the control unit to receive from external VFO. Calibrate the external VFO dial (R-4B dial) in the same manner as the TR-4.

3. Switch the control unit to separate receive (don't forget to throw the slide switch for separate receive). The R-4B should now operate normally. Adjust the trimmer mounted below the new crystal to zero beat.

summing up

Now you can receive a station on the R-4B and then switch immediately to transceive on the same frequency by simply selecting transceive on the control box and moving the slide switch on the side of the R-4B. You're now transmitting with R-4B VFO control. It's a good idea to check the dial calibration from time to time to make sure that everything is still calibrated, otherwise you'll have a small frequency offset when switching to transmit from the R-4B.

You have greatly increased the versatility of the TR-4/R-4B combination, and, if your experience is like mine, you'll also have greatly increased your operating pleasure.

ham radio

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*The steps in these modifications should be kept in a file with your equipment literature. Appropriate annotations to the steps will come in handy when you decide to restore your radio for resale. **Editor**.

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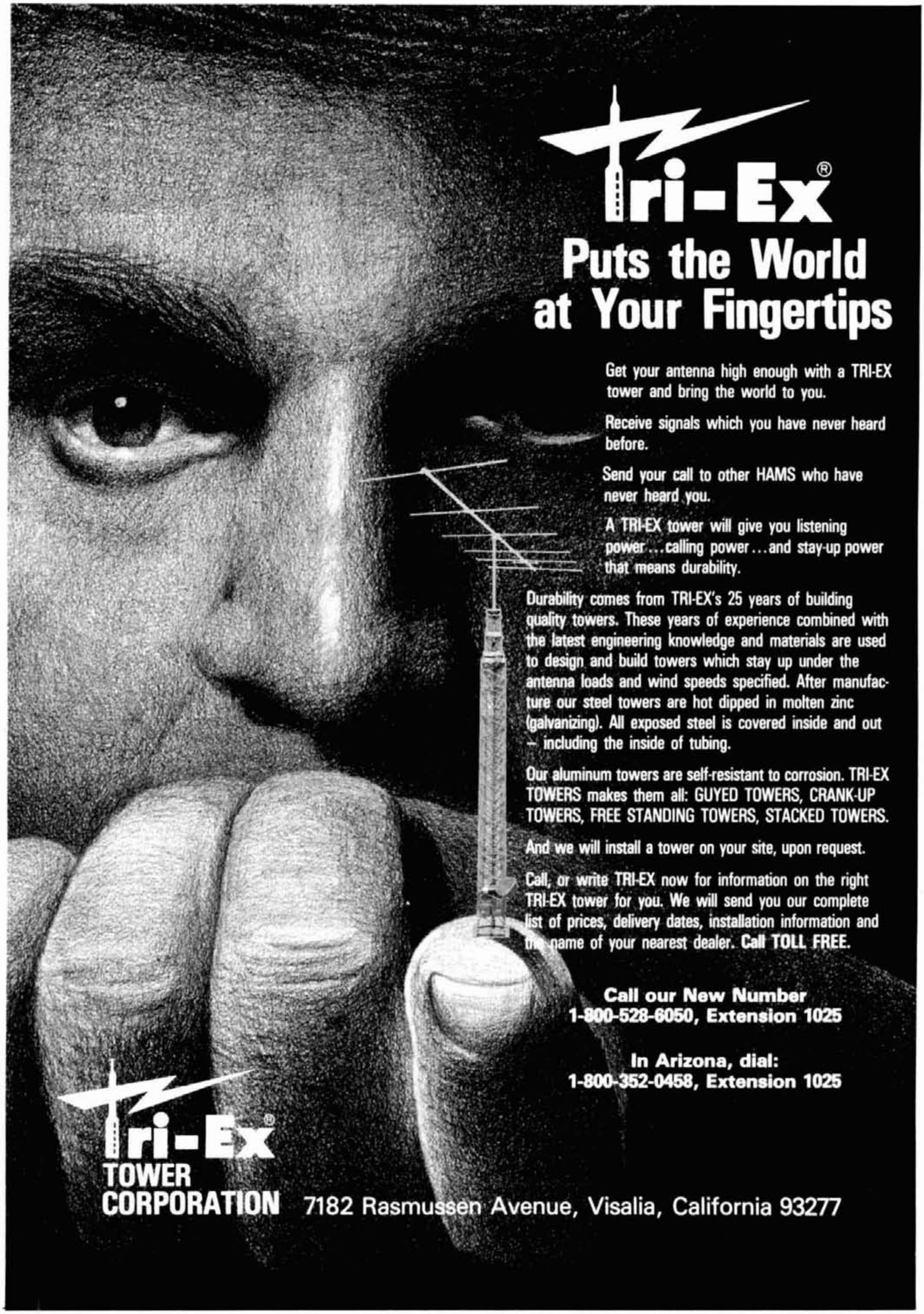

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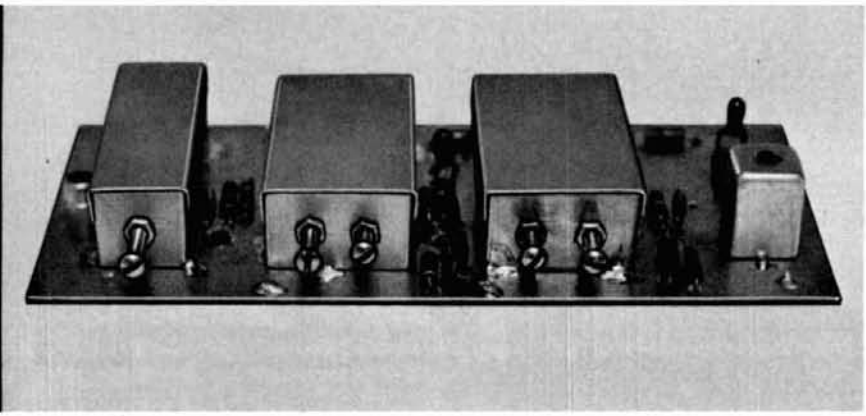


high-performance 432-MHz converter

Complete construction details for a 432-435 MHz receiving converter that is easy to build and tune up

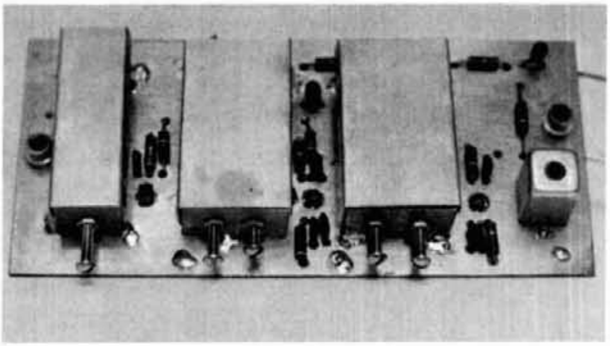
The 432-MHz receiving converter described here is the result of several month's work. As anyone who has attempted to build solid-state uhf gear knows, layout and construction technique is vital, not only to coming up with a good basic design, but also to building a converter that works properly.

Now that economically priced solid-state equipment and devices are available for use above 432 MHz, the cost of building uhf equipment has been reduced considerably. This converter uses three transistors (six if you build the oscillator chain for copying Oscar 8). Three Texas Instruments dual-gate 3CT225A mosfet transistors are used in the converter (90 cents each in small quantities). The 3CT225A transistors are rated at 900 MHz with a 4-dB noise factor; at 432 MHz with 12 Vdc drain voltage and 4 volts on gate 2, gain will be about 22 dB with a noise factor of 2 dB or less. These transistors, and the stripline tuned circuits, makes this a very good converter for weak-signal reception.

Two versions of the converter were built using coils in the tuned circuits, but the results were far below that achieved with striplines. A local-oscillator chain was not built on the converter board because the 404-MHz injection frequency was coupled from the oscillator chain in my 432-MHz transmitting converter. However, a diagram of an easy-to-build oscillator chain will also be described. I use it with the converter to copy Oscar 8 Mode J.

layout and construction

Component arrangement, component spacing, and wiring have been optimized for best results, based on several earlier versions. The spacing between the tuned circuits in the amplifier and mixer



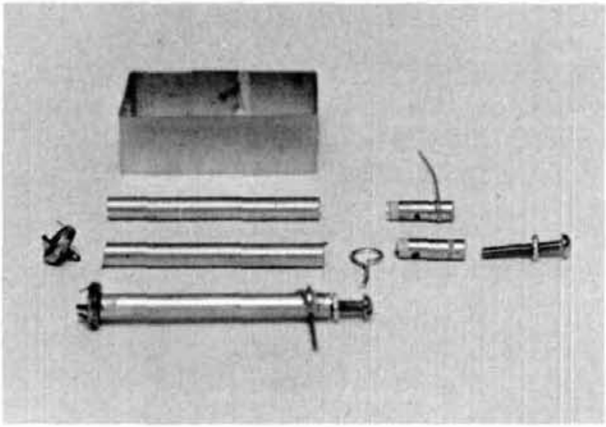
Top view of the 432-435 MHz converter showing placement of the enclosures with the tuned lines. Input is to the left, output to the right. Transformer T1 is in the shield can near the output connector.

By C. H. Robinson, N9KD, 89 Nottingham Road, Springfield, Illinois 62074

was laid out to provide satisfactory interstage coupling without the use of capacitors.

The circuit boards I used are glass epoxy. The oscillator board has foil on one side; the converter board has copper on both sides. Anyone can make the circuit boards, even if he has never etched boards before. Start by cutting the boards to size; clean the boards with steel wool and paint the copper. I used *Rustoleum* gray primer in a spray can — it protects the copper while in the etching solution, and it is easy to remove when you scribe the lines. After the painted boards have dried, using the pattern, trace out the lines; since there are no curved lines, scribing is very easy.

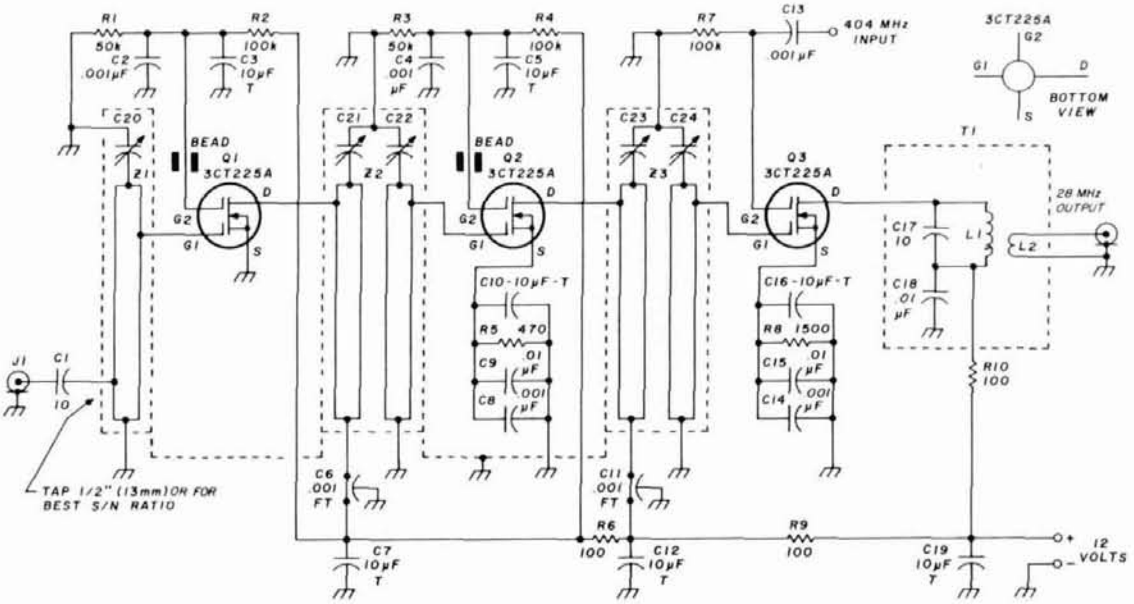
Use a small jeweler's screwdriver to scribe the lines. You can paint a spare piece of circuit board and practice getting the lines the right width; about 1/16 inch (1.5 mm) is best. Put the scribed boards in etching solution and agitate once in a while to speed up the etching process. After all the necessary copper has been etched off, thoroughly clean the boards with soap and water using steel wool. A final cleaning with abrasive household cleanser is recommended to remove any last traces of the etchant solution.



Finished components for one of the tuned stripline circuits. The button mica capacitor is soldered to the bottom of the line; the piston capacitor is inserted in the top and soldered in place (see text).

tuned-circuit construction

The enclosures for the striplines are made from brass strips 0.065 inch (1.5 mm) thick, 3/4 inch (19 mm) wide, and 12 inches (30 cm) long. These strips and the 1/4-inch (6.5 mm) brass tubes can be purchased in many hobby shops. The dimensions for the second rf and the mixer stages are 1-1/4 × 2-1/4 × 3/4 inches (32 × 57 × 19 mm); the first rf stage is 2-1/4 × 3/4 × 3/4 inches (57 × 19 × 19 mm). These are inside dimensions, so, when bending, allow for the bend angle. One way is to mark off the first bend, bend in a vise, then mark the second bend, and so on. One side must be longer than the other so it laps and can be soldered. The spacing between the tuned lines is 1/4 inch (6.5



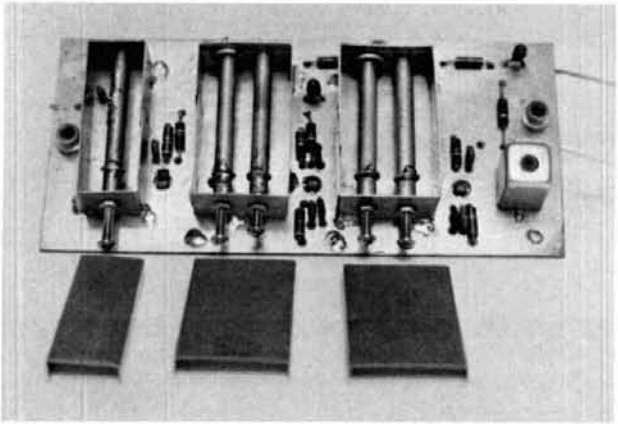
C20-C24 1.5-10 pF ceramic piston capacitors (Centralab 829-10)

T1 Hamtronics 7807 coil form and shield. L1 is 22 turns no. 26 (0.4 mm) close-wound; L2 is 4 turns no. 26 (0.4 mm) on bottom of L1

fig. 1. Schematic diagram of the high-performance converter for 432-435 MHz; companion local-oscillator chain is shown in fig. 2. Capacitors marked with FT are button mica feedthrough types; capacitors marked with T are tantalum; all other capacitors are disc ceramic. Z1, Z2, and Z3 designate the stripline circuits.

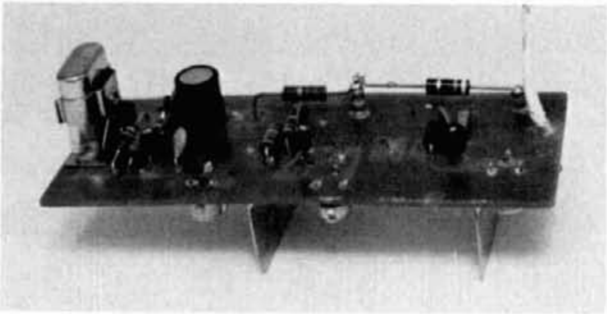
mm) and 1/4 inch (6.5 mm) from the side of the enclosure.

The tubes for the tuned lines are 2-1/8 inches (54 mm) long; this allows 1/8 inch (3 mm) clearance between the end of the tube and the enclosure where the trimmer is mounted. Solder tin the inside of the tube about 1/2 inch (13 mm) down. Remove the wire pigtail from the trimmer, and insert the trimmer in the tube; you may have to apply heat with the soldering iron to get it in. About 1/8 inch (3 mm) of the trimmer should protrude beyond the tube. Apply heat to the tube and sweat solder the trimmer in place.



Top view of the converter with the enclosure covers removed to show the placement of the tuned lines. Note the holes in the board for mounting the transistors.

Two of the tuned lines have dc blocking capacitors mounted at the ends; these are mounted like the trimmers, but the tubes will be shorter by the thickness of the capacitor. Form the piece of wire that connects the tuned lines to the PC board, but do not solder yet; this will be done later when the tuned lines are placed in the enclosures.

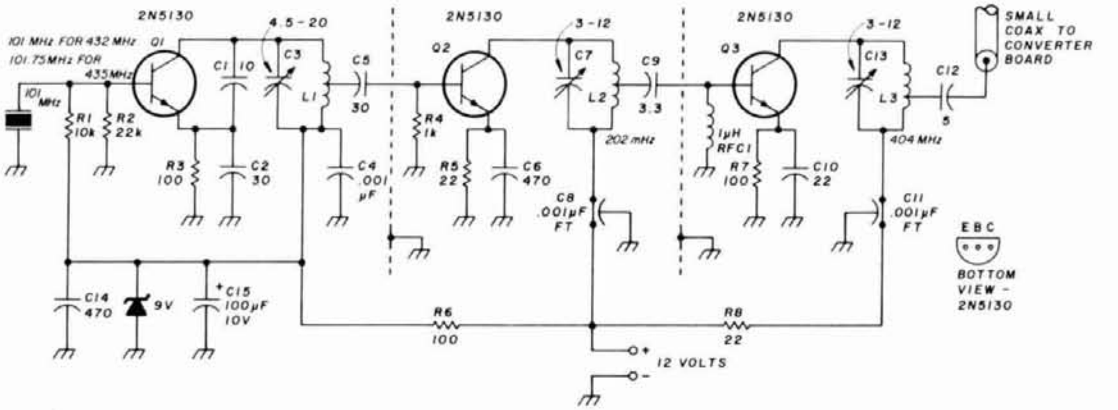


Construction of the 404-MHz local-oscillator chain. Printed-circuit layout is shown in fig. 4.

mounting and components

If you don't have a 404-MHz oscillator chain for injection, build the oscillator chain first. I built the oscillator board one circuit at a time, drilling the holes for the components as I went, testing each stage before starting the next. The schematics and photographs show all the details. After the oscillator board is complete and tested, lay it aside and start construction on the converter board.

Wiring the converter is straightforward; if you study the photographs and schematic you should have no problems. However, there's one precaution: Since the converter board has copper foil on both sides, the foil around the component holes on top of the board must be cleared to prevent shorts. To do this use a 1/8-inch (3-mm) drill and flatten the angle of the point so it will cut the copper from around the top hole but not touch the copper on the bottom of



L1 6 turns no. 16 (1.3-mm) wire, 1/4-inch (6.5 -mm) diameter, 3/4 inch (19 mm) long, tapped at one turn
L2 3 turns no. 16 (1.3-mm) wire, 1/4-inch (6.5-mm) diameter, 3/4 inch (19 mm) long, tapped at 1-1/2 turns
L3 1 turn no. 16 (1.3-mm) wire, 1/4-inch (6.5-mm) diameter, with 1/2-inch (13-mm) leads

fig. 2. 404-MHz local-oscillator chain for the 432-435 MHz converter. Capacitors marked with FT are feedthrough types; except for the electrolytic (C15), all other capacitors are disc ceramic.

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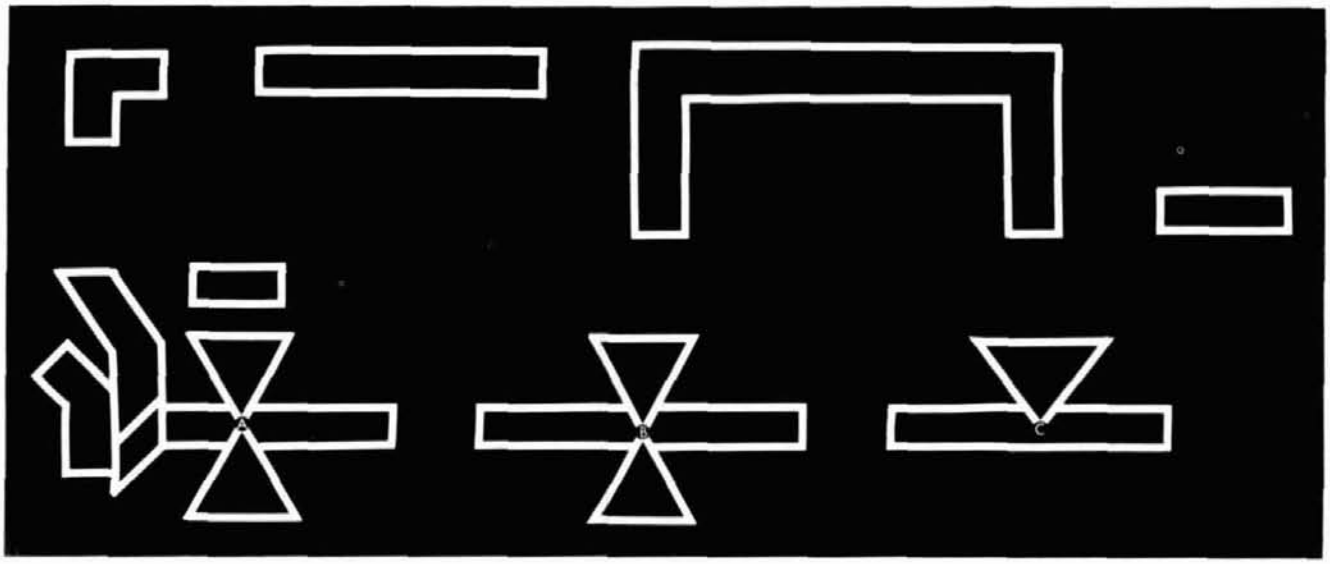


fig. 3. Full-size printed-circuit layout for the 432-435 MHz converter. Component placement for the converter is shown in the photograph. Transistor mounting holes marked A, B, and C are 5/16 inch (8 mm) diameter.

the board. Use an emery wheel to shape the drill bit for this task.

After the mixer wiring has been completed, connect the output of the oscillator to gate 2 of Q3 in the converter board with small coax cable; apply 12 Vdc and common to both boards. You should be able to receive the third harmonic of a two-meter transmitter that tuned to 144 MHz, or a weak-signal source if you have one. There are two excellent articles on weak-signal generators in past issues of *ham radio*.[1,2] When the mixer is working properly, proceed with the other two stages — testing as you complete each one.

You may be able to build the converter without using the 10-μF tantalum capacitors, but they do much to stabilize the circuits so they are well worth the extra cost. You may have noticed that on the schematic diagram ferrite beads are shown on gate 2 of each rf stage; the photograph shows only one on the first stage — my error, use the bead to improve stability.

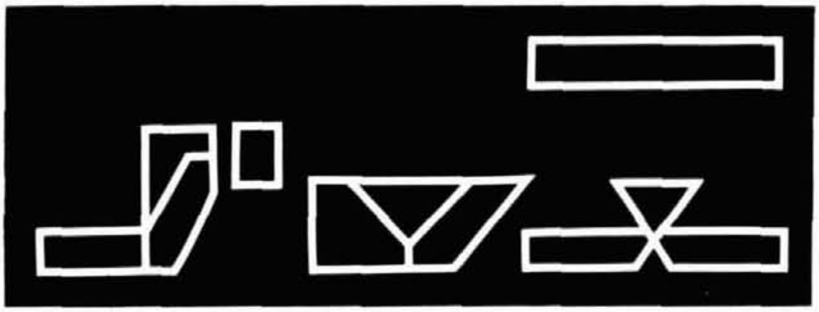
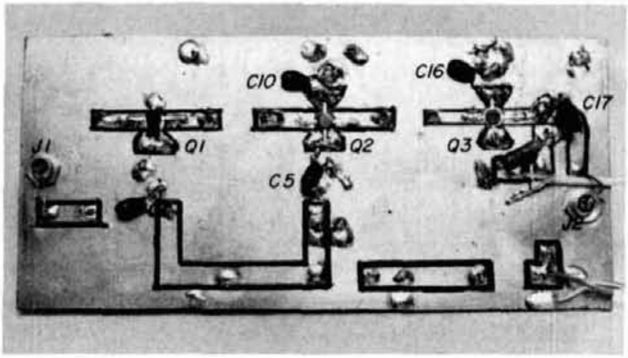


fig. 4. Printed-circuit layout for the 404-MHz local-oscillator chain. Component mounting and shield placement is shown in the photograph.

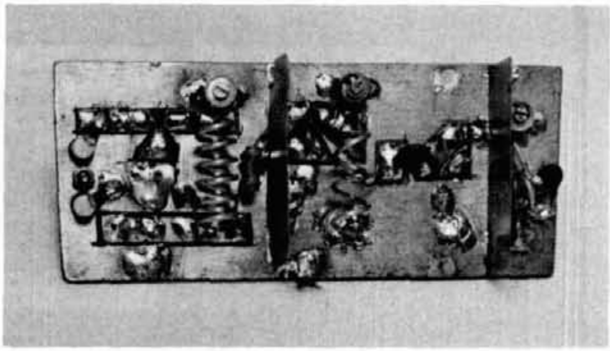
tune up

If a commercial signal generator is available, use it; if not, use a weak-signal source. You may find that the test signal will leak into the converter from places other than the antenna! Take the weak signal outside, as far away from the antenna as possible, or until you can't pick up the signal with the antenna disconnected. With the antenna connected you should receive the weak signal at S-5 or better. You can now peak everything up for optimum performance.

With my weak-signal source about 100 meters from the antenna I receive it at S-9 using a Kenwood TS-520. After the circuits have been tuned for maximum signal, tune away from the signal and check the noise level as indicated on the receiver's S-meter. The objective is to obtain the greatest signal-to-noise ratio. You can also adjust the converter for lowest noise figure if you have equipment for noise figure measurements.[3,4]



Component placement on the converter board (see fig.3). In this view the input is to the left, output to the right. Designated capacitors are mounted on the bottom of the board. Small coax cable near C17 couples 404-MHz injection to the converter.



Component placement on the circuit board for the 404-MHz local-oscillator chain.

When using this converter with a TS-520 transceiver and no 28-MHz preamp, I have consistently maintained schedules on 432 MHz with stations over 100 miles (160 km) away, with reports ranging from S-4 to 20 dB over S-9 depending upon conditions. During band openings many stations have been heard 700-800 miles (1100-1300 km) out with S-meter readings over S-9. I have also used this converter to copy Oscar 8 and hear the spacecraft signals around S-6 without any special antennas. I have built several 432-MHz converters from handbooks, magazine articles, and commercial kits, but this converter tops them all.

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ham radio

impedance measurements using an swr meter

The simple SWR meter and a handheld calculator can be used to measure complex relationships involving impedance in tuned circuits

An SWR measurement can indicate when an rf source is properly matched to a load, but often it's desirable to know the actual impedance values for the load. Once these values are determined, it's then possible to design matching circuits or transformers to translate the load impedance to the desired impedance.

For example, as more solid-state broadband transmitters come into use, a matched load impedance becomes more desirable. While the high SWR of an antenna may not damage transistors rated to withstand that SWR, the output power may be reduced because of an improper load impedance. The lack of tuning and load adjustments prohibits rematching the load impedance to the final output stage for the rated transmitter performance.

Another example of the requirement for knowing the load impedance can be seen for a typical installation with an antenna cut for the 75-meter phone band, but with an occasional trip down in frequency to check into an 80-meter CW net. When an antenna tuner isn't available it's possible to construct a matching circuit so the transmitter still sees a load impedance within its rated SWR, even though the actual antenna SWR is somewhat higher. Knowing the actual impedance values of the 75-meter antenna while using it at 80 meters will yield a starting point for construction of such a circuit.

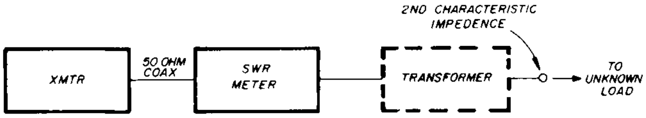

fig. 1. System for measuring characteristic impedances other than 50 ohms. The transformer can be a lumped-constant element, broadband transformer, or a transmission line such as a quarter-wavelength of 75-ohm coaxial cable at the frequency of interest.

measurement procedure

Many instruments are available to measure impedance values, but only a few fortunate Amateurs have them. An instrument that is available, or at least not too expensive or difficult to construct, is an SWR meter.

Once the SWR of an impedance is known, half the battle is over. If the reflection coefficient electrical angle, θ, is known, the impedance can be quickly determined by using a Smith chart. (This is discussed later.)

Information that's available is the SWR based on a characteristic impedance, Z_{01}, of 50 ohms. This means the source, usually a transmitter, looks like a 50-ohm source, and the SWR meter has a 50-ohm characteristic impedance. However, the use of a Smith chart isn't the only method of determining an impedance when the SWR and the reflection coefficient electrical angle are known. The impedance can be calculated outright by knowing the SWR, the reflection coefficient electrical angle, and the characteristic impedance of the measuring system.[1]

If another SWR measurement could be made with a different characteristic-impedance-measurement system, Z_{02}, the same impedance values would be found. Of course a different SWR reading would exist, as well as a different reflection coefficient electrical angle. But the impedance values of the load will not have changed.

Since the impedance is constant for the two measurements, the equations relating SWR, Z_{01}, and θ, are set equal to each other, yielding:

$$(\tan\theta_1)(\tan\theta_2) = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)\right](swr_1) - swr_2}{\left[\left(\frac{Z_{02}}{Z_{01}}\right)\right](swr_2) - swr_1} \qquad (1)$$

$$\frac{\tan\theta_1}{\tan\theta_2} = \frac{\left(\frac{Z_{02}}{Z_{01}} \cdot swr_1 \cdot swr_2\right) - 1}{(swr_1 \cdot swr_2) - \frac{Z_{02}}{Z_{01}}} \qquad (2)$$

The SWR is a measured value, and Z_{01} is known

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from the characteristic impedance of the test equipment.

*This means that a second characteristic-impedance-SWR-measurement system must be devised. Such a system is easily obtainable by using the exist*ing 50-ohm source and an SWR meter and transforming this impedance to some other impedance, as shown in **fig. 1**. The transformer can be a lumped-element, broadband transformer[2,3] or a transmission-line-type, such as a quarter-wavelength of 75-ohm coax at the frequency of interest. (This quarter-wavelength transformer results in a characteristic impedance of 112.5 ohms.)

The procedure is as follows. It's simpler to perform than to describe. Using the 50-ohm system, measure and record the SWR of an unknown impedance. This measurement yields swr_1 and $Z_{01} = 50\ ohms$. Then connect the transformer at the output, or antenna side, of the SWR meter. *Connect the unknown load* to the output port of the transformer. Again, measure the SWR. This measurement yields swr_2 and Z_{02}. These four values, swr_1, Z_{01}, swr_2 and Z_{02}, are substituted in **eqs. 1** and **2**, resulting in two equations and two unknowns. The two unknowns are $\tan\theta_1$, and $\tan\theta_2$, where θ_1 and θ_2, are the reflection-coefficient electrical angles.

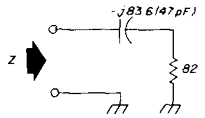


fig. 2. Impedance test circuit.

Unfortunately, more than one solution exists, giving more than one value for $\tan\theta_1$, and $\tan\theta_2$. But the resulting impedance calculations yield the same reactance magnitude using either values; whether the reactance is inductive or capacitive can't be determined directly. Another test must be conducted to make this determination; such tests are described later.

From the measurements, all the information necessary to calculate the actual impedance has been found. Substituting swr_1, $\tan\theta_1$, and Z_{01} into **eq. 3** gives the impedance. Likewise, substituting swr_2, $\tan\theta_2$, and Z_{01} into **eq. 4** gives the same impedance:

$$Z = Z_{01}\left[\frac{1-j(swr_1)\tan\theta_1}{swr_1-j\tan\theta_1}\right] \tag{3}$$

or

$$Z = Z_{02}\left[\frac{1-j\,(swr_2)\tan\theta_2}{swr_2-j\tan\theta_2}\right] \tag{4}$$

test results and example

To test this procedure, I constructed a test load using an 82-ohm resistor in series with a 47-pF capacitor. A randomly selected piece of 75-ohm cable was found to look like a quarter wavelength at 40.5 MHz. At this frequency the resistor and capacitor look like 82-j83.6 ohms (**fig. 2**).

The reactance of the 47-pF capacitor is found from:

$$X_C = \frac{1}{2\pi fC} = \frac{1}{2\pi(40.5\times 10^6)(47\times 10^{-12})}$$
$$= -j83.6$$

The one-quarter wavelength of 75-ohm coax transforms the 50-ohm source impedance to 112.5 ohms:

$$\frac{Z_L^2}{Z_1} = Z2 = \frac{(75)^2}{50} = 112.5\,ohms \tag{5}$$

Now the two SWR measurements are made. Using the 50-ohm system ($Z_{01} = 50\ ohms$), swr_1 of the capacitor and resistor in series is measured as 4:1. The quarter-wavelength, 75-ohm transformer is inserted between the SWR meter and the unknown impedance, and another SWR reading, swr_2, is measured as 2.5:1; Z_{02} is 112.5 ohms. These values are now substituted into **eqs. 1** and **2** to determine the values of $\tan\theta_1$ and $\tan\theta_2$. Note that calculating the values of θ_1 and θ_2 isn't necessary unless the reflection coefficient electrical angle is desired; only the $\tan\theta_1$ and θ_2 values are needed. These calculations yield:

$$(\tan\theta_1)(\tan\theta_2) = \frac{\left(\frac{112.5}{50}\right)(4)\ -2.5}{\left(\frac{112.5}{50}\right)(2.5)\ -4} = 4 \tag{6}$$

$$\frac{\tan\theta_1}{\tan\theta_2} = \frac{\left(\frac{112.5}{50}\bullet 4\bullet 2.5\right)-1}{(4\bullet 2.5)-\frac{112.5}{50}} = 2.77 \tag{7}$$

Eq. 7 yields: $\tan\theta_1 = 2.77\tan\theta_2$. Substituting into **eq. 6** yields:

$$2.77\tan^2\theta_2 = 4$$
$$\tan\theta_2 = \pm 1.2$$

Solving for $\tan\theta_1$ by substituting the value of $\tan\theta_2$ into **eq. 7** yields:

$$\tan\theta_1 = 2.77\,(\pm 1.2) = \pm 3.32$$

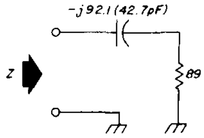
The positive value for $\tan\theta_1$, Z_{01}, and swr_1 are now substituted into **eq. 3** to give Z, the unknown impedance:

$$Z = 50\left[\frac{1-j\,(4)\,(3.32)}{4-j\,3.32}\right]$$
$$= 50\,[2.56\angle -46^\circ]$$
$$Z = 50\,[1.78-j\,1.84] = 89-j\,92.1$$

This gives the resistance as 89 ohms and the capacitive reactance as −j92.1 ohms, resulting in a capacitor value at 40.5 MHz of:

$$C = \frac{1}{2\pi f x_c} = \frac{1}{2\pi(40.5 \times 10^6)(92.1)} = 42.7\,pF$$

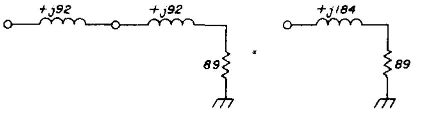
The measured circuit is then:



This compares favorably with the component values used in the circuit of **fig. 2**.

Two pertinent factors should be noted here. The values for Z_{02}, swr_2, and $\tan\theta_2$ could just as easily have been substituted into **eq. 4** to obtain the same impedance. The negative values for $\tan\theta$ could have been used, giving a positive value for the impedance reactive component. Often, observation of the circuit can indicate the proper sign of the reactive component. If this isn't possible a second measurement can be made at *another frequency, and the direction of* change for the reactive component can be observed. For instance, a higher frequency should yield a lower capacitive reactance. If the reactance is inductive, the higher frequency should yield a higher value of reactance.

Another method of determining the sign of the reactance would be to insert a known reactance value in series with the measured impedance and make another measurement. For the example given above, an inductor of approximately +j92 ohms at 40.5 MHz (0.36 μH) would cancel the −j92 ohm reactance, leaving only a real component of 80 ohms. If *the reactance happened to be positive in the first* place, the second measurement would yield:



Actually, a second calculation isn't necessary, since it can quickly be seen that 89 + j0 would yield a much lower SWR than 89 + j184.

theory

The derivation of this concept follows directly by equating **eqs. 3** and **4**. This is reproduced here for reference:

$$Z = Z_{01}\left[\frac{1-j(swr_1)\tan\theta_1}{swr_1 - j\tan\theta_1}\right] = Z_{02}\left[\frac{1-j(swr_2)\tan\theta_2}{swr_2 - j\tan\theta_2}\right]$$

$$(1 - j\,swr_1\tan\theta_1)(swr_2 - j\tan\theta_2) = \frac{Z_{02}}{Z_{01}}(1 - j\,swr_2\tan\theta_2)(swr_1 - j\tan\theta_1)$$

Equating the real and imaginary terms yields:

$$(A)\; swr_2 - swr_1\tan\theta_1\tan\theta_2 = \frac{Z_{02}}{Z_{01}}(swr_1 - swr_2\tan\theta_1\tan\theta_2)$$

$$(B)\; swr_1\,swr_2\tan\theta_1 + \tan\theta_2 = \frac{Z_{02}}{Z_{01}}(\tan\theta_1 + swr_1\,swr_2\tan\theta_2)$$

From A,

$$\left[\left(\frac{Z_{02}}{Z_{01}} \bullet swr_2\right) - swr_1\right]\tan\theta_1\tan\theta_2 = \left(\frac{Z_{02}}{Z_{01}} \bullet swr_1\right) - swr_2$$

$$\tan\theta_1\tan\theta_2 = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_1)\right] - swr_2}{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_2)\right] - swr_1}$$

From B,

$$\tan\theta_1\left[(swr_1 \bullet swr_2) - \left(\frac{Z_{02}}{Z_{01}}\right)\right] = \tan\theta_2\left[\left(\frac{Z_{02}}{Z_{01}} \bullet swr_1 \bullet swr_2\right) - 1\right]$$

$$\frac{\tan\theta_1}{\tan\theta_2} = \frac{\left[\left(\frac{Z_{02}}{Z_{01}}\right)(swr_1)(swr_2)\right] - 1}{\left[(swr_1)(swr_2)\right] - \frac{Z_{02}}{Z_{01}}}$$

Exactly what transpires can readily be seen by looking at the Smith chart shown in **fig. 3**. The first SWR measurement, $swr_1 = 4$, yields a reflection coefficient, ρ_1, of 0.6. The electrical angle of the reflection coefficient, θ_1, is found from $\tan\theta_1 = \pm 3.32$. Therefore, $\theta_1 = \pm 73.2°$. Using only the positive value and knowing that the total distance around the Smith chart is 180°, one-half wavelength, the ratio of 73.2° to 180° gives a reflection coefficient at 0.203λ. This is shown as Z_1 and corresponds to the normalized impedance of 1.78 − j1.84 for the 50-ohm measurement system. When the 112.5-ohm system is employed, swr_2 (2.5) results in a value of $\rho_2 = 0.43$. The reflection coefficient electrical angle, θ_2, is 50.2°, found from $\tan\theta_2 = 1.2$. For this system the reflection coefficient in wave-lengths is 0.14λ. These values are also plotted on the Smith chart, but it must be remembered that the center of the chart represents 112.5 ohms + j0. This normalized impe-

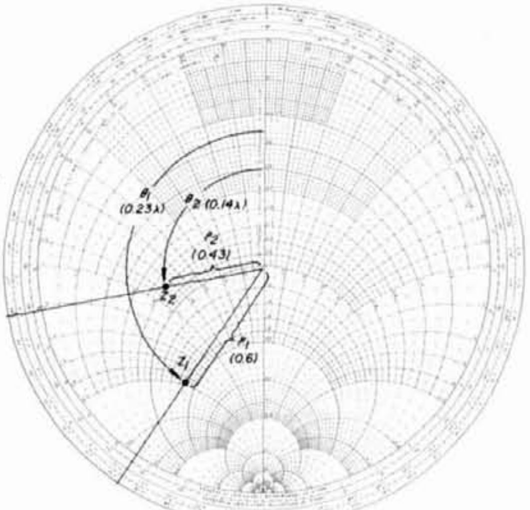


fig. 3. Test results shown on a Smith chart for the example described in the text.

dance is 0.79 − j0.82 and is shown as Z_2. Actually, both impedances, Z_1 and Z_2, are equal; only the characteristic impedance of the measurement system is different.

One other point should be mentioned. Suppose the load is actually 82 − j83.6, but a matching circuit is designed to match the source impedance to the measured value of 89 − j91.1. How suitable is this? A new reflection coefficient must be determined from:

$$\Gamma = \rho \angle \theta = \frac{Z - Z_0}{Z + Z_0} \tag{8}$$

where $Z_0 = 82 - j83.6,\ Z = 89 - j92.1$

Making these substitutions yields a reflection coefficient of $\rho = 0.045$, which results in an SWR of 1.09:1.

conclusion

While there are many methods of measuring impedance, the procedure described is straightforward and requires only one simple instrument, the SWR meter. Also, since so many handheld calculators are available, the calculations using the measured values can be accomplished very quickly. Other means of calculating can be employed, although more time is required. The results are acceptable for most practical applications.

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3. *The Radio Amateur's Handbook*, ARRL, Newington, Connecticut, April 1975, page 581.

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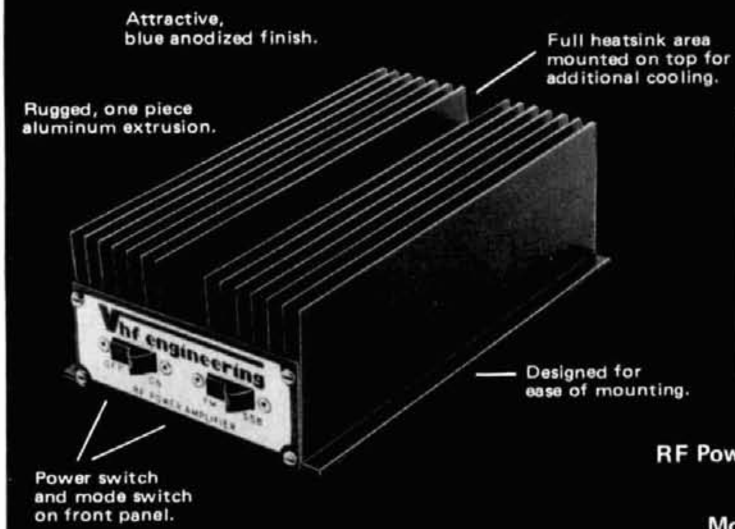

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DIGITAL TECHNIQUES

flip-flop internal structure

Previous parts of this series have shown the three basic flip-flops, the latch, and clocked flip-flops JK and D. Only the RS latch has been examined for timing. Timing is crucial in proper operation of JK and D flip-flops, so it's worthwhile to examine some typical internal structures for timing relationships.

Fig. 1 shows a NAND-gate equivalent of a master-slave flip-flop with waveforms and "1 and 0" state notation. The master-slave term comes from using one latch (G3 and G4) to control the other latch (G7 and G8). State feedback makes each latch dependent on the other. The 1 and 0 notation is useful for scratchpad state analysis, and the state is that of each gate after the clock edge has passed.

JK flip-flops are commonly found in master-slave form, and **fig. 1** has both J and K control inputs held high for a divide-by-two function. A negative clock edge will change output state, so the symbol would use an inversion bubble at the clock pin.

Initial conditions assume all inputs high and Q low. Master latch G3, G4 could be in either state but is chosen with G3 initially high. All gate states are stable in the left-hand 1 and 0 notation.

The first negative clock forces G1 high. G5 then goes low since both inputs are high (the *NAND RULE*) and forces G7 (Q output) high which, in turn, makes G8 ($\overline{Q}$ output) low. Waveform arrows show the sequential state change. The other four gates remain in the same state as long as the clock is low.

Returning the clock high doesn't affect output but will set up conditions for the next negative clock edge. G2 goes low and forces intermediate latch G3 and G4 to change state. G5 returns to a high, and output *latches* G7 *and* G8 *remain stable* (*both* G5 *and* G6 are now high).

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

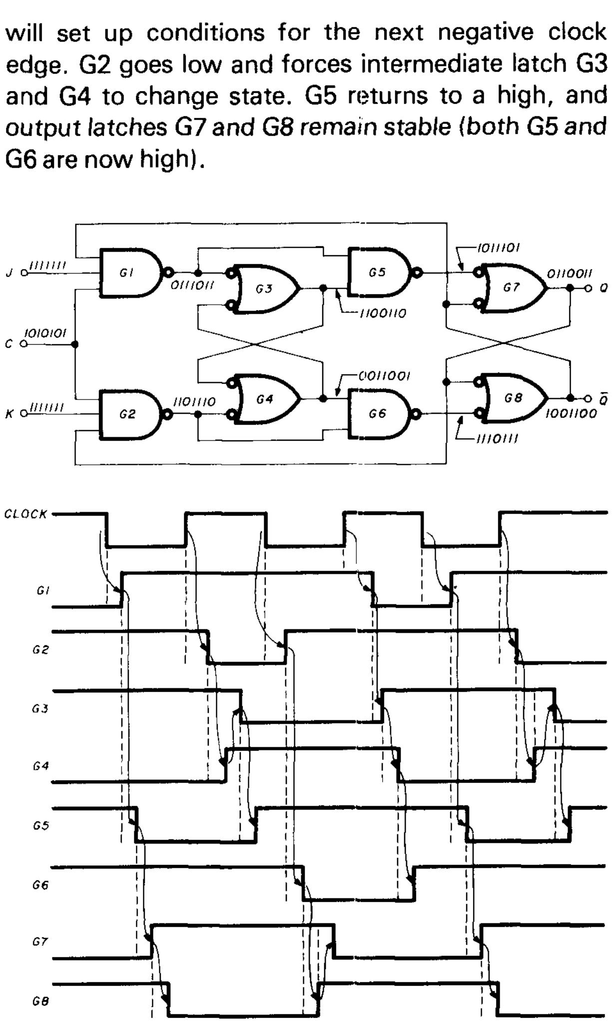

fig. 1. Master-slave flip-flop with J and K inputs held high.

A second negative clock will make G2 high. G6 now goes low and flips the output latch. Q output changes state. Returning the clock high will flip the intermediate latch and set up conditions for another output state change.

An important thing to note is that both clock edges affect internal states and that each edge causes a sequence of four gate state changes with attendant delays. A high clock level must persist long enough to set up conditions for an output change. A low clock level must persist for a time sufficient to ripple-through state changes to toggle the output. Each time will limit maximum clock frequency.

changing control inputs

Fig. 2 has the same circuit but input K is held low and only J is changed. Initial conditions have Q low and J low. Since K is low, G2 will always remain high. Note that if both J and K were held low, there would be no output change at all since input gates G1 and G2 would be held high constantly. The intermediate latch has G3 low.

The first low clock does nothing. Returning it high has no effect either, since J is held low. When J goes high with the clock high, the intermediate latch flips from the low state of G1. G6 goes high and the flip-flop is set up for an output change.

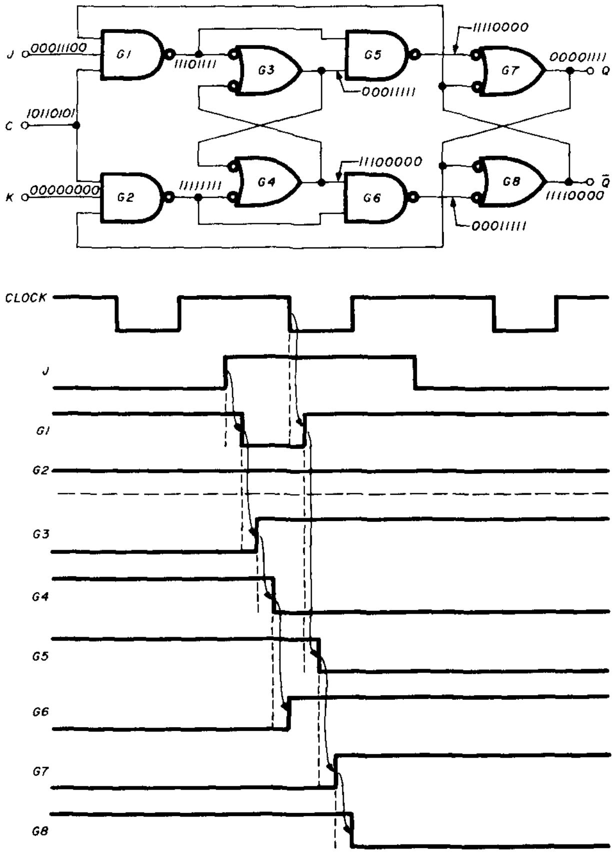

fig. 2. Master-slave flip-flop with K held low and J input switched.

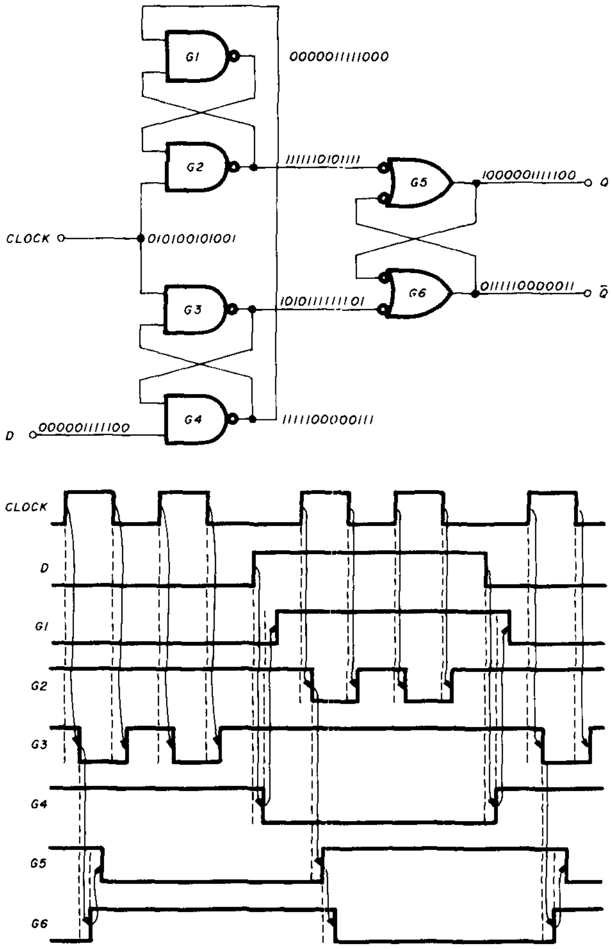

fig. 3. Type-D flip-flop.

The second low clock, while J is held high, will return G1 high and make G5 low. The output latch flips and Q goes high, matching the J input state. Again, four gate delays are required for both set-up and hold.

Once Q has toggled high, neither J nor the clock will have any effect. Why? The answer lies in state feedback from G8 (Q) to G1 input. The low G8 state will inhibit any inputs from changing G1 (the *NAND RULE* again).

Holding J low and changing only K will result in similar action, except that Q will eventually go low and remain in that state. Try this out on some scratchpaper.

type D flip-flop

A NAND gate equivalent is given in **fig. 3.** This version changes output on a *positive* clock edge. Gates G1 through G4 are "almost" latches with an *AND* symbol shape; their action is best observed by following input changes.

Initial state has Q high and D low. With both D input and clock low, G2 through G4 are held high. G1 is held low by state feedback from G4. Output latch

G5, G6 remains stable since all inputs (from G2 and G3) are high.

The first high clock will force G3 low; its other input from G4 is also high. G3 then forces G6 (Q) high to flip the output latch. Clock return to low will change only G3, as does a second clock input. G3 stays high on a low clock.

Changing D input high while the clock is low will force G4 low, which, in turn, forces G1 high. Input gates are now set up to change the output. A third positive edge clock will make G2 low and flip the output latch to a Q high state. As long as D remains high, further clocks will not change Q.

Returning D low while the clock is low will force G4 high, then G1 low to finish a setup. The cross connections of G1, G2, and G3, G4 appear to make two latches. These, plus G4 to G1 connection, make one large latch, not two, with G1 and G4 acting as inhibits for G2 and G3.

This structure is faster than the master-slave JK. Setup requires only two gate delays, output change only three.

direct set and reset

A direct set or reset will override the clock and any control inputs. The master-slave JK may be modified by increasing the latch gate inputs from two to three. A SET (active low) is connected to both G3 and G7. A RESET (again, active low) is connected to both G4 and G8.

The D flip-flop is also modifiable for active-low set or reset but is more complex. A RESET is made to G2, G4, and G6, a SET made to only G1 and G5. The number of gate inputs must be increased for either.

Actual devices may have direct set and reset (sometimes called preset and clear, respectively) either active low or active high. Check for inversion bubbles. Some dual devices have either or both common. Check the spec sheet. In any situation an unused set or reset must be made inoperative. An unused active low should be tied high; unused active high tied low.

Removing a set or reset will restore a JK or D to normal clocked operation. Some time is required for this restoration, similar to setup and hold times. The spec sheet for a particular device should be checked for this and proper time alloted in circuit operation.

The next part of this series will go into interfacing analog signals and present one-shots.

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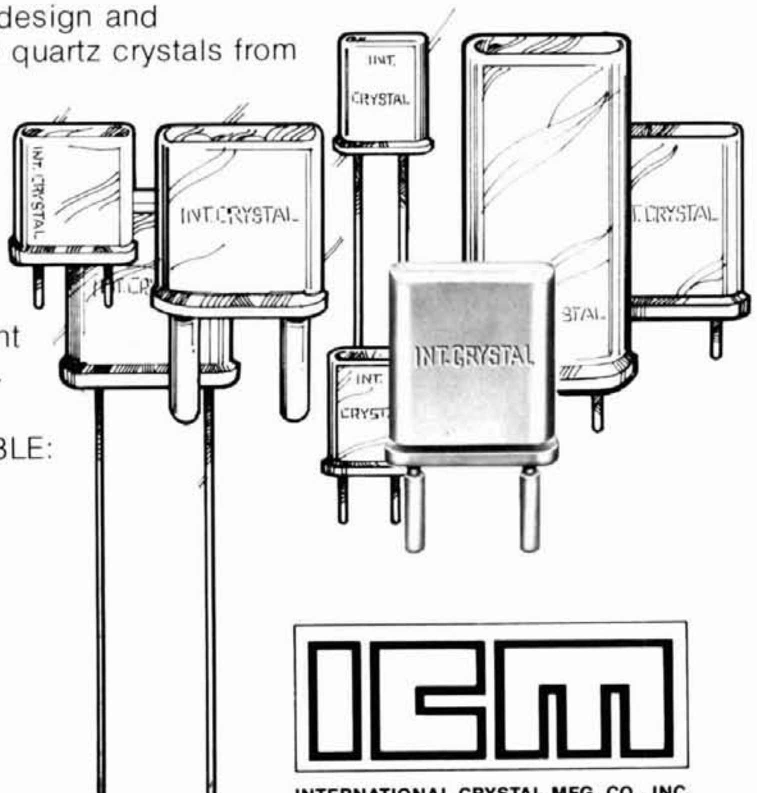

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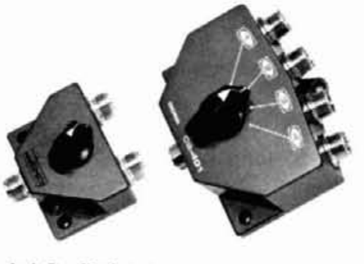

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external frequency programmer

The beauty of the IC-22S is that it offers precise frequency selection by programming with low-cost silicon diodes, rather than crystals with trimmers. This advantage though becomes overshadowed when channel selections need to be changed; out comes the soldering iron to redo the diode matrix. To solve this problem, I built an external programmer containing eight switches, eight diodes, and a zero-center microammeter for discriminator output readings.

External frequency programmer enclosed in a Ten-Tec style box. The connections to the transceiver are made via nine-conductor ribbon cable.

The internal speaker was disconnected from the external speaker jack and reconnected directly to the audio output. The discriminator output was then connected to the external speaker jack. The eight diode/cathode connections and the 9-volt common pad of channel 22 on the matrix board were connected to the nine pins of the accessory socket via a nine-conductor ribbon cable (see **fig. 1**).

The external programmer was built in a Ten-Tec cabinet. The discriminator meter was connected via a shielded cable and miniplug. Nine-conductor ribbon cable was used to connect the programmer to the accessory plug.

With the programmer, any frequency can be selected at the flip of a switch and at a considerable savings over the cost of a fully programmable 2-meter rig.

Hugh Pearl, WB9VWM

using the IC-22S below 146 MHz

As it turns out, the IC-22S is not restricted to 146.01 MHz and above. The IC-22S is, in fact, usable without modification from 145.350 MHz through 147.990 MHz. That's an additional 44 free channels! These additional channels do accommodate some of the new repeater frequencies as well as five Oscar frequencies. By keying the push-to-talk line you can also enjoy the excitement of Oscar 7 and Oscar 8.

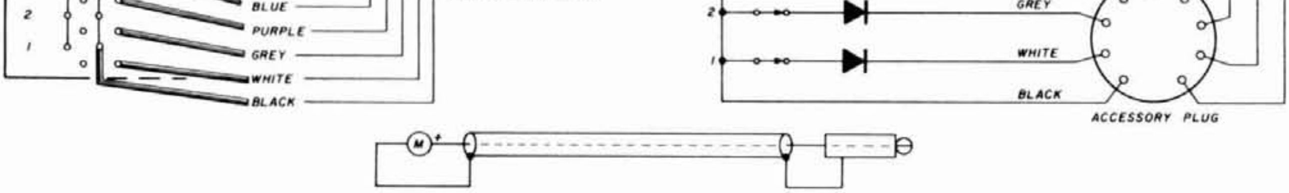

fig. 1. Diagram of the connections to the diode matrix board and switch connections for the external frequency programmer. The discriminator meter is connected to the rewired external speaker jack.

table 1. Diode programming matrix for frequencies below 146 MHz.

		diode insert positions							
frequency MHz	total N	128 D7	64 D6	32 D5	16 D4	8 D3	4 D2	2 D1	1 D0
145.350	64		•						
145.365	65		•						•
145.380	66		•					•	
145.395	67		•					•	•
145.410	68		•				•		
145.425	69		•				•		•
145.440	70		•				•	•	
145.455	71		•				•	•	•
145.470	72		•			•			
145.485	73		•			•			•
145.500	74		•			•		•	
145.515	75		•			•		•	•
145.530	76		•			•	•		
145.545	77		•			•	•		•
145.560	78		•			•	•	•	
145.575	79		•			•	•	•	•
145.590	80		•		•				
145.605	81		•		•				•
145.620	82		•		•			•	
145.635	83		•		•			•	•
145.650	84		•		•		•		
145.665	85		•		•		•		•
145.680	86		•		•		•	•	
145.695	87		•		•		•	•	•
145.710	88		•		•	•			
145.725	89		•		•	•			•
145.740	90		•		•	•		•	
145.755	91		•		•	•		•	•
145.770	92		•		•	•	•		
145.785	93		•		•	•	•		•
145.800	94		•		•	•	•	•	
145.815	95		•		•	•	•	•	•
145.830	96		•	•					
145.845	97		•	•					•
145.860	98		•	•				•	
145.875	99		•	•				•	•
145.890	100		•	•			•		
145.905	101		•	•			•		•
145.920	102		•	•			•	•	
145.935	103		•	•			•	•	•
145.950	104		•	•					
145.965	105		•	•					•
145.980	106		•	•				•	
145.995	107		•	•				•	•

The programming of the IC-22S diode matrix is governed by the equation:

$$programmed\ number\ (N) = \frac{desired\ frequency - 144.39}{.015}$$

The resulting number must be an integer (no fractional part). The decimal number computed from the equation is then translated into a binary number and programmed accordingly into the diode matrix. For convenience, **table 1** presents the additional frequencies and the corresponding diode positions in a format similar to that in the instruction manual. The simplex-offset switch functions the same as for other frequencies below 147 MHz.

Steven Holzman, W1IBI

75S ( ) CW sidetone

The CW sidetone provided by my 32S-1 transmitter is of sufficient amplitude to give an adequate monitoring level while using headphones. Even when low-impedance headphones (stereo types with the sections paralleled) were used, no problems were encountered here.

In an attempt to increase the sidetone level while using a speaker, the value of R99 (2.2 meg ohms) was altered. This change proved insufficient, and, if carried too far, vox operation on ssb was hampered. The best balance between speaker, headphones, overall receiver volume, and CW sidetone level was obtained in the following manner.

Remove the two wires connected to the SIDETONE jack on the receiver rear lip, solder them together, and insulate. Using a piece of shielded wire (RG-174/U in my case), connect one end to the SIDETONE jack and the other end to the center (wiper) terminal of the AF GAIN control. The cable should be neatly routed around the chassis following the path of the existent harness. Refer to **fig. 2** for the complete wiring changes. The result is a sufficient sidetone level for both speaker and phones.

Paul Pagel, N1FB

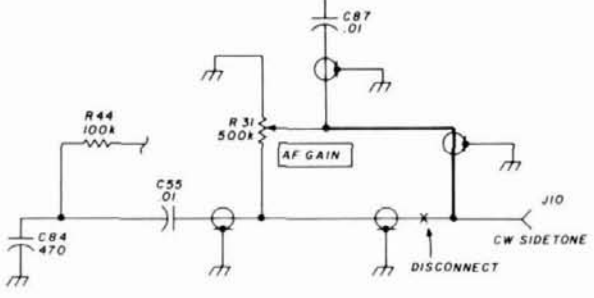

fig. 2. Sidetone modification for the 75S() series receivers. In this case, the level of the sidetone is controlled by the AF GAIN, providing a better balance between signal and sidetone levels. The heavy line indicates the added shielded cable.

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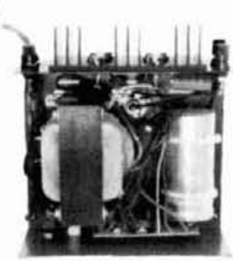

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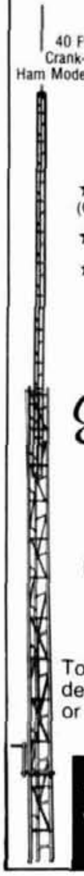

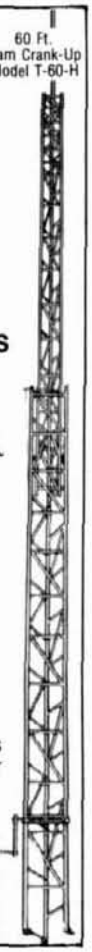

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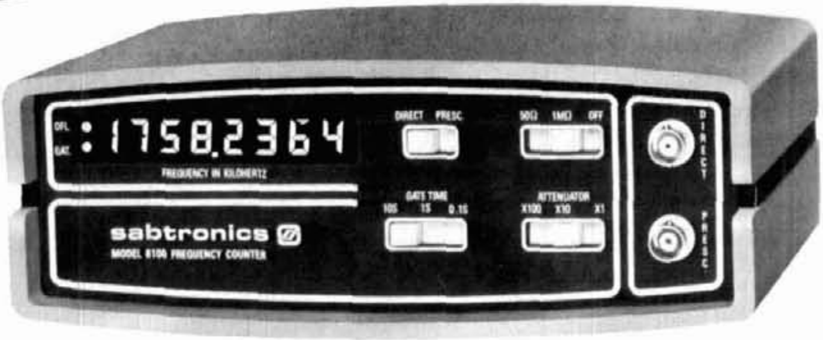

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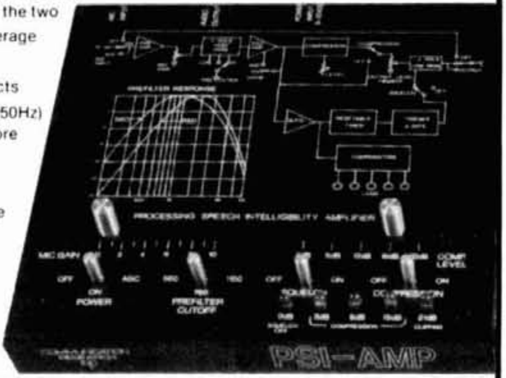

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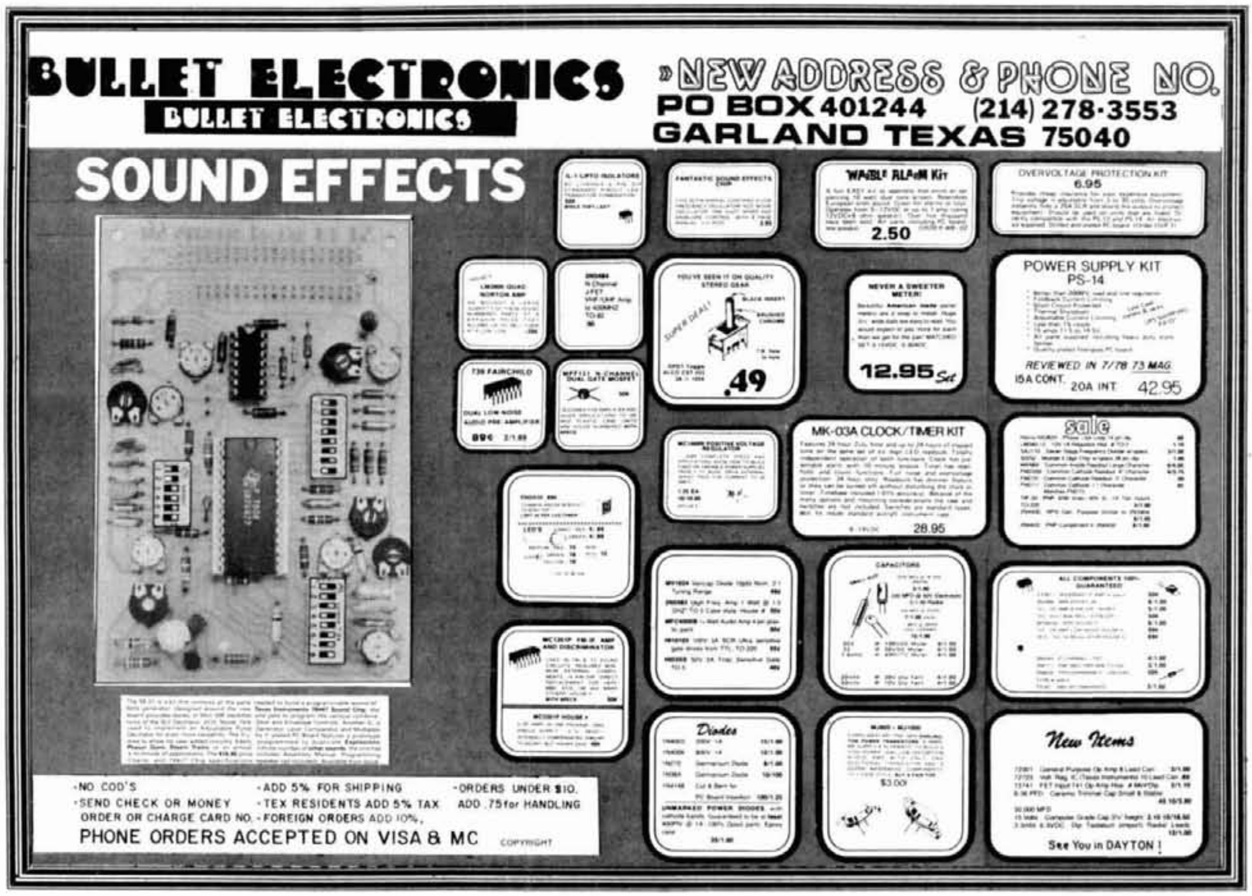

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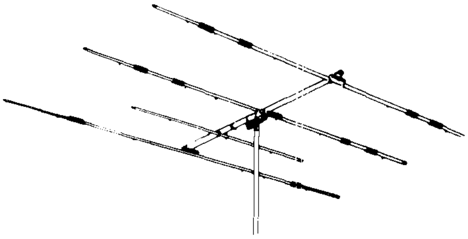

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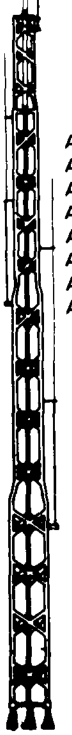

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Coming Events

ILLINOIS: 13th Annual Rock River Radio Club hamfest, Sunday, April 22, Lee County 4-H Center, South of Dixon. Advance, $1.50; Gate, $2.00. Indoor facilities, camping area, free coffee & donuts 7:30 to 8:30 AM. Food available, prizes. Talk-in 146.52 simplex or WR9ADG Repeater 146.37-146.97. Advance tickets: RRRC Hamfest, Chuck Randall, W9LDU, 1414 Ann Ave., Dixon, IL 61021.

NEW YORK STATE QSO party, sponsored by the University of Buffalo Amateur Radio Club, WA2NPQ, from 1700Z May 6 to 0500Z May 7 and 1200-2359Z May 7. Stations may be contacted once on phone and once on CW on each band. NY stations may work each other. Mobiles/portables changing counties may be reworked. Exchange consists of report, serial number starting with 001, and QTH (counties for NY, state/province for others). Suggested frequencies: CW, 1810 3560 7060 1460 2160 28060; Phone, 3900 7275 14285 21375 28550; Novice, 3725 7125 21125 28125. Score one point per QSO times the number of multipliers: states, provinces, countries and NY counties for NY stations; and NY counties for others (maximum of 62). This is the first time NY stations may include counties in the multiplier total. Number the first contact for each new multiplier. A check sheet is required for stations making more than 100 contacts. Awards. Logs must be received by June 16 to qualify and entrants desiring results please send #10 SASE. Please also include your name, address and county (if NY). Send all entries to Michael Bergman, WD2AJS, 45 Swartson Ct., Albany, NY 12209.

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

FLORIDA: The Daytona Beach Amateur Radio Association's First Hamfest, May 12 and 13, Holiday Inn Surfside, Daytona Beach. Advance registration: $3.00 per family, $3.50 at door. Indoor, air-conditioned facilities. "Drive-on" ocean beach, nearby Bellair Shopping Plaza. For info: David Rusler, WA4ZTT, 1725 Hope Drive, Ormond Beach, FL 32074, (904) 672-9536.

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

NEW YORK: MARAC — ICHN convention, May 4, 5, 6, 1979 in Albany, New York, and hosted by Russ Sawyer, W2UJ; Bruce Jacobs, K2QK; Cliff Miller, N2FN and Jerome Walsh, WA2LSU. The purpose of the convention is to meet club members face-to-face and to raise funds for donation to a national charity. For information, write: Jerry Walsh, WA2LSU, R.F.D. #1, P.O. Box 197B, Hudson, New York 12534.

SHIP ISLAND! The Magnolia DX Group with the support of the Mississippi Coas A.R.A. plans to operate WN5IJZ from Ship Island during the 1979 CQWW-WPX Contest. This is the first ever operation from Ship Island, 12 miles south of Gulfport, Mississippi in the Gulf of Mexico, and the home of historic Fort Massachusetts. Operation will be multi-single on 80 through 10 meters. Commemorative QSL to all stations worked. Manager: N5FG/WB5HVY, Floyd Gerald, 4706 Washington Avenue, Gulfport, Mississippi 39501. S.A.S.E. please!

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

INDIANA: Hamfest! Indiana's friendliest and largest hamfest. Wabash County Amateur Radio Club's 11th annual hamfest will be held Sunday, May 13, 1979, rain or shine, at the Wabash County 4-H Fairgrounds in Wabash. Large flea market (no set-up charge), technical forums, bingo, free overnight camping, plenty of free parking, good food at reasonable prices. Only one ticket to buy. Donation is $2.50 for advance tickets — $3.00 at the gate. Children under 12 years old free. For more information or advance tickets, write: Dave Nagel, WD9BDZ, 555 Valley Brook Ln., Wabash, IN 46992, S.A.S.E. Required.

Transistor Checker

– Completely Assembled –
– Battery Operated –

The ASI Transistor Checker is capable of checking a wide range of transistor types, either "in circuit" or out of circuit. To operate, simply plug the transistor to be checked into the front panel socket, or connect it with the alligator clip test leads provided. The unit safely and automatically identifies low, medium and high power PNP and NPN transistors. Size: 3¼" x 6¼" x 2".
"C" cell battery not included.

Trans-Check $29.95 ea.

Custom Cables & Jumpers

DB 25 Series Cables

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

Dip Jumpers

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

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

CONNECTORS

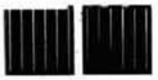
25 Pin-D Subminiature

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

PRINTED CIRCUIT EDGE-CARD

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

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

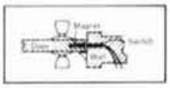


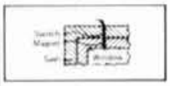
Solar Cells 2x2cm

- 0.4 volts
- 100mA
- 41 MW

Can be added in series for higher voltage or parallel for higher current.

#SC 2x2 **$1.95 ea. or 3/$5.00**





Magnetically Activated Switch

The 9250-0002 is a single pole normally closed switch. When the magnet is engaged, the circuit is open. This switch is only suitable for use in non-magnetic doors and windows.

#9250-0002 2/$1.00

AC Wall Transformer

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

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

REMOTE CONTROL TRANSMITTER & RECEIVER

- [illegible]
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- [illegible]
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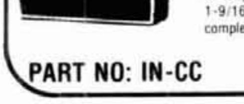
$19.95

INSTRUMENT/CLOCK CASE



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

PART NO: IN-CC $3.49 each

MICROPROCESSOR COMPONENTS

8080A/8080A SUPPORT DEVICES

8080A	CPU	$ 9.95
8212	8-Bit Input/Output	3.25
8214	Priority Interrupt Control	5.95
8216	Bi-Directional Bus Driver	3.49
8224	Clock Generator/Driver	3.95
8226	Bus Driver	3.49
8228	System Controller/Bus Driver	5.95
8238	System Controller	5.95
8251	Prog. Comm. I/O (USART)	7.95
8253	Prog. Interval Timer	14.95
8255	Prog. Periph. I/O (PPI)	9.95
8257	Prog. DMA Control	19.95
8259	Prog. Interrupt Control	19.95

6800/6800 SUPPORT DEVICES

MC6800	MPU	$14.95
MC6802CP	MPU with Clock and Ram	24.95
MC6810API	128X8 Static Ram	5.95
MC6821	Periph. Inter. Adapt (MC6820)	7.49
MC6828	Priority Interrupt Controller	12.95
MC6830L8	1024X8 Bit ROM (MC68A30-8)	14.95
MC6850	Asynchronous Comm. Adapter	7.95
MC6852	Synchronous Serial Data Adapt	9.95
MC6860	0-600 bps Digital MODEM	12.95
MC6862	2400 bps Modulator	14.95
MC6880A	Quad 3-State Bus. Trans. (MC8T26)	2.25

MICROPROCESSOR CHIPS—MISCELLANEOUS

Z80(780C)	CPU	$19.95
Z80A(780-1)	CPU	24.95
CDP1802	CPU	19.95
2650	MPU	19.95
8035	8-Bit MPU w/clock, RAM, I/O lines	19.95
P8085	CPU	19.95
TMS9900JL	16-Bit MPU w/hardware, multiply & divide	49.95

SHIFT REGISTERS

MM500H	Dual 25 Bit Dynamic	$ .50
MM503H	Dual 50 Bit Dynamic	.50
MM504H	Dual 16 Bit Static	.50
MM506H	Dual 100 Bit Static	.50
MM510H	Dual 64 Bit Accumulator	.50
MM5016H	500/512 Bit Dynamic	.89
2504T	1024 Dynamic	3.95
2518	Hex 32 Bit Static	4.95
2522	Dual 132 Bit Static	2.95
2524	512 Static	.99
2525	1024 Dynamic	2.95
2527	Dual 256 Bit Static	2.95
2528	Dual 250 Static	4.00
2529	Dual 240 Bit Static	4.00
2532	Quad 80 Bit Static	2.95
2533	1024 Static	2.95
3341	Fifo	6.95
74LS670	4X4 Register File (TriState)	1.95

UART'S

A-Y-5-1013	30K BAUD	5.95

MICROPROCESSOR MANUALS

M-Z80	User Manual	$7.50
M-CDP1802	User Manual	7.50
M-2650	User Manual	5.00

ROM'S

2513(2140)	Character Generator(upper case)	$9.95
2513(3021)	Character Generator(lower case)	9.95
2516	Character Generator	10.95
MM5230N	2048-Bit Read Only Memory	1.95

RAM'S

1101	256X1	Static	$1.49
1103	1024X1	Dynamic	.99
2101(8101)	256X4	Static	3.95
2102	1024X1	Static	1.75
21L02	1024X1	Static	1.95
2111(8111)	256X4	Static	3.95
2112	256X4	Static MOS	4.95
2114	1024X4	Static 450ns	9.95
2114L	1024X4	Static 450ns low power	10.95
2114-3	1024X4	Static 300ns	10.95
2114L-3	1024X4	Static 300ns low power	11.95
5101	256X4	Static	7.95
5280/2107	4096X1	Dynamic	4.95
7489	16X4	Static	1.75
74S200	256X1	Static Tristate	4.95
93421	256X1	Static	2.95
UPD414 (MK4027)	4K	Dynamic 16 pin	4.95
UPD416 (MK4116)	16K	Dynamic 16 pin	14.95
TMS4044-45NL	4K	Static	14.95
TMS4045	1024X4	Static	14.95
2117	16,384X1	Dynamic 350ns (house marked)	9.95
MM5262	2KX1	Dynamic	4/1.00

PROM'S

1702A	2048	FAMOS	$5.95
TMS2516 (2716)	16K*	EPROM(Intel 2716)	49.95
	*Requires single +5V power supply		
TMS2532	4KX8	EPROM	89.95
2708	8K	EPROM	10.95
2716 T.I.	16K**	EPROM	29.95
	**Requires 3 voltages, —5V, +5V, +12V		
5203	2048	FAMOS	14.95
6301-1(7611)	1024	Tristate Bipolar	3.49
6330-1(7602)	256	Open C Bipolar	2.95
82S23	32X8	Open Collector	3.95
82S115	4096	Bipolar	19.95
82S123	32X8	Tristate	3.95
74186	512	TTL Open Collector	9.95
74188	256	TTL Open Collector	3.95
74S287	1024	Static	2.95

CONTINENTAL SPECIALTIES
PROTO BOARDS

Proto Board 203

A total, ready-to-use power breadboard prototyping device with a built-in regulated, short-proof power supply. Just plug in and start building! 2 extra floating 5-way binding posts for external signals. Self-contained with power switch, indicator lamp and power fuse. 24 14-pin DIP capacity. Attractive hard-to-mark quality case. All metal construction. 9½" L x 6½" W x 2¾" H. 5 lbs.

PB 203 **$75.00**

Proto Board 203A

All the features of the PB-203 plus additional power supply flexibility: provides 1% Regulated 5VDC supply (same specifications as PB-203). Regulated separate +15VDC and -15VDC 0.5A supplies, each with internally and independently adjustable output voltage. Ripple and noise of + and - 15V supplies 10mV at 0.25A.

PB 203A **$124.95**

Model Number	L x W x H (Inches)	Price
PB-6	6.0 x 4.5 x 1.4	$15.95
PB-100	6.0 x 4.5 x 1.4	$19.95
PB-101	6.0 x 4.5 x 1.4	$22.95

Model Number	L x W x H (Inches)	Price
PB-102	7.0 x 4.5 x 1.4	$26.95
PB-103	9.0 x 6.0 x 1.4	$44.95
PB-104	9.8 x 8.0 x 1.4	$54.95

B+K PRECISION

3½-Digit Portable DMM

- Overload Protected
- .3" high LED Display
- Battery or AC operation
- Auto Zeroing
- 1mV, 1V≥, 0.1 ohm resolution
- Overrange reading
- 10 meg input impendence
- DC Accuracy 1% typical

Ranges: DC Voltage 0-1000V
AC Voltage 0-1000V
Freq. Response 50-400 Hz
DC-AC Current 0-100mA
Resistance 0-10 meg ohm
Size 6.4" x 4.4" x 2"

Model 2800
$99.95

Comes with test leads, operating manual and spare fuse.

Accessories:

AC Adapter BC-28	$9.00
Rechargeable Batteries BP-26	20.00
Carrying Case LC-28	7.50

CONTINENTAL SPECIALTIES CORPORATION

100 MHz 8-Digit Counter

- 20 Hz-100 MHz Range
- .6" LED Display
- Crystal-controlled timebase
- Fully Automatic
- Portable — completely self-contained
- Size — 1.75" x 7.38" x 5.63"
- Four power souces, i.e. batteries, 110 or 220V with charger, 12V with auto lighter adapter and external 7.2-10V power supply

MAX-100 $134.95

ACCESSORIES FOR MAX 100:

Mobile Charger Eliminator
use power from car battery — **Model 100 — CLA $3.95**

Charger/Eliminator
use 110 V AC — **Model 100 — CAI $9.95**

New

Mini-Max 6 Digit 50MHz Frequency Counter

- Guaranteed frequency range of 100 Hz to 50 MHz
- Full 6 digit display with antiglare window
- Fully automatic-range, polarity, slope, trigger, input level switching not required
- Lead-zero blanking—All zeros to the left of the first non-zero digit are blanked. Kilo Hertz and Mega Hertz decimal points automatically light up when the unit is turned on.
- Built in input overvoltage protection
- Use 9V Battery or 110/220V power
- Complete with mini antenna
- Lightweight — Only 8oz.

MINI-MAX $89.95

Accessories For Mini-Max

Part No.	Description	Price
MM-A4	Antenna	$ 3.95
MM-CS	Carrying case	5.95
MM-IPC	Input cable with clip leads	3.95
MM-AC2	110V adapter	9.95
MM-AC3	220V adapter	9.95

$10.00 Minimum Order — U.S. Funds Only
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The Incredible
"Pennywhistle 103"

$139.95 Kit Only

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

Data Transmission Method	Frequency-Shift Keying, full-duplex (half-duplex selectable)
Maximum Data Rate	300 Baud
Data Format	Asynchronous Serial (return to mark level required between each character)
Receive Channel Frequencies	2025 Hz for space, 2225 Hz for mark
Transmit Channel Frequencies	Switch selectable: Low (normal) – 1070 space, 1270 mark; High – 025 space, 2225 mark
Receive Sensitivity	-46 dbm acoustically coupled
Transmit Level	-15 dbm nominal. Adjustable from -6 dbm to -20 dbm
Receive Frequency Tolerance	Frequency reference automatically adjusts to allow for operation between 1800 Hz and 2400 Hz
Digital Data Interface	EIA RS-232C or 20 mA current loop (receiver is optoisolated and non-polar)
Power Requirements	120 VAC, single phase, 10 Watts
Physical	All components mount on a single 5" by 9" printed circuit board. All components included

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

TRS-80 16K Conversion Kit

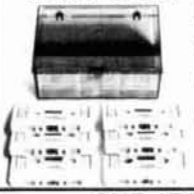
Expand your 4K TRS-80 System to 16K. Kit comes complete with:

- 8 each UPD416 (16K Dynamic Rams)
- Documentation for conversion

TRS-16K $115.00

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

COMPUTER CASSETTES



- 6 EACH 15 MINUTE HIGH QUALITY C-15 CASSETTES
- PLASTIC CASE INCLUDED 12 CASSETTE CAPACITY
- ADDITIONAL CASSETTES AVAILABLE #C-15-$2.50 ea

CAS-6
$14.95
(Case and 6 Cassettes)

SUP 'R' MOD II

UHF Channel 33 TV Interface Unit Kit

Wide Band B/W or Color System

- ★ Converts TV to Video Display for home computers, CCTV camera, Apple II, works with Cromeco Dazzler, SOL-20, IRS-80, Challenger, etc.
- ★ MOD II is pretuned to Channel 33 (UHF).
- ★ Includes coaxial cable and antenna transformer.

MOD II $29.95 Kit

RS-232 CONTROL CENTER

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

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

PART NO.
RS-232CC $89.95 kit only

CASSETTE CONTROLLER

Ideal for use with the TRS 80 and others.

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

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

#CC-100
$29.50

63-Key Unencoded Keyboard

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

Hexadecimal Unencoded Keypad

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

ILLINOIS: Kishwaukee Radio Club and DeKalb Co. Amateur Repeater Club's 21st annual indoor/outdoor hamfest, Sunday, May 6, 8:00 AM to 3:00 PM, Notre Dame School, Gurler Road, DeKalb. Tickets $1.50 advance, $2.00 door. Indoor tables available, own tables free, outdoor setup free. Talk-in 146.13/73 and 94 simplex. For info: SASE to Howard WA9TXW, P.O. Box 349, Sycamore, IL 60178.

ILLINOIS: The Moultrie Amateur Radio Klub's 18th annual hamfest, April 22, Moultrie County 4-H Center Fairgrounds, Cadwell Road, Mattoon. Indoor/outdoor flea market. Talk-in 146.94 and 146.055/.655. For info: M.A.R.K., Box 327, Mattoon, IL 61938.

INDIANA: The Cass County Amateur Radio Club's second annual hamfest, Sunday, May 6, from 7:00 AM to 4:00 PM, at the 4-H Fairgrounds. Rain or Shine. Go North to Logansport on Highway 25, turn right at road 100, follow QSY signs. Advance tickets $1.50, $2.00 at gate. Outside setup free, undercover $1.00. Bring your own tables. Free overnight camping, plenty of refreshments. Talk-in: 146.52 and Logansport Repeater: 147.78-18. Write to: K9DVL Dave Rothermel, RR 4, Box 146 G, Logansport, IN 46947.

CALIFORNIA: The 37th Annual Fresno hamfest, May 11-13, Sheraton Inn, Clinton and Hiway 99. Technical talks, swap tables, flea market, transmitter hunt on 2 meters (146.52), QLF contest, exhibits, prizes. Prime rib banquet. Registration with eligibility for pre-registration prize, $17.00 before April 27. $19.00 and no pre-reg prize after that date. Talk-in on 146.34/146.94. For info: Fresno Amateur Radio Club, P.O. Box 783, Dept. HF, Fresno, CA 93712.

KENTUCKY HAM-O-RAMA — Sunday, May 20, at Boone County Fairgrounds, Burlington, Kentucky. 10 miles South I-75 of Cincinnati, Ohio. Major and hourly door prizes included with $3.00 gate ticket. Info: N.K.A.R.C., Box 31, Ft. Mitchell, KY 41017.

MASSACHUSETTS: South Shore Repeater Association's 4th annual Ham Auction, Saturday, April 21, Central Junior High School, Broad Street, Weymouth. 12 Noon. Doors open 9 AM. Club share 10%, refreshments, door prizes. For info: South Shore Repeater Assn., Town Hall Annex, 402 Essex St. Weymouth, MA 02188.

MASSACHUSETTS: Central Mass. Amateur Radio Association's auction and flea market, April 27, 7:30 PM. Doors open 6:00 PM. Main South American Legion Post 341, Main Street at Webster Square, Worcester. Auction rates — 15% to CMARA, Flea Market tables $5.00. Door prizes, raffles , refreshments. Talk-in 146.37-146.97 and .52 direct. For info: Rene Brodeur, WA1LEA (617) 753-7480 and Dave Penttila, K1COW, (617) 885-4995.

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual auction, Saturday, April 7, 11:00 AM (doors open 10:00 AM). Wellesley High School Cafeteria, Rich Street, Wellesley. Talk-in on 96/36, 63/03, 04/64 and 52. Contact: Kevin P. Kelly, WA1YHV, 7 Lawnwood Place, Charlestown, MA 02129.

MASSACHUSETTS: The Hampden County Radio Association's annual flea market, Friday, May 4, Feeding Hills Congregational Church, intersection of Routes 57 and 187, west of Springfield. $2.00 per table, no admission fee. Doors open 7:00 PM. Refreshments available. For info: Andy Bouchard, WB1BZW — (413) 786-2301.

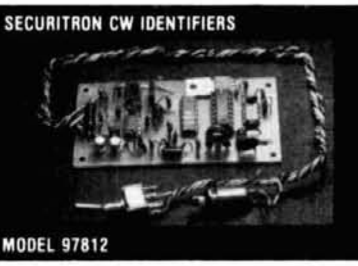
MICHIGAN: South Eastern Michigan Amateur Radio Association's 21st annual hamfest, April 8, 8:00 AM to 3:00 PM EST, South Lake High School, 21900 E. Nine Mile Road at Mack Avenue, St. Clair Shores. For info: SEMARA, Mark C. Wilke, WD8RDA, 171 Merriweather Rd., Grosse Pointe Farms, MI 48236.

MICHIGAN: The Wexaukee ARA's 19th annual Swap and Shop, Saturday, May 19. 9:00 AM - 4:00 PM, National Guard Armory, 415 Haynes St., Cadillac. Tickets $2.00, free parking. Talk-in on 146-37/97. Info from WD8RZL.

MISSISSIPPI: The Jackson Amateur Radio Club's annual hamfest, April 21 and 22, Manhattan Academy gymnasium, 5055 Manhattan Rd., Jackson. For info: JARC, Box 8371, Jackson, MS 39204 or call MS. Sideband Net daily, 3987.5, 2345 GMT.

MISSOURI: Central Missouri Radio Association's fourth annual hamfest, April 7, Cosmo Recreation Center, 1507 Business Loop 70 West, Columbia. FCC exams, forums, ladies' programs, flea market, prizes. Open for exhibitors April 6, 7-9 PM, April 7, 6:30 AM. Reserved commercial space, $15.00 per table. For info write: C.M.R.A., P.O. Box 283, Columbia, Missouri 65201.

MISSOURI: The P.H.D. Amateur Radio Association's tenth annual hamfest, Saturday and Sunday, April 21, 22, Kansas City Trade Mart, Kansas City Downtown Airport, free parking. Display booth space, $15.00 single, $25.00 double. 11:00 AM to 5:30 PM Saturday. 10:00 AM to 5:00 PM Sunday. Set up beginning 9:00 AM Saturday. PHD ARA, P.O. Box 11, Liberty, MO 64068.



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NEW ENGLAND: The Hosstraders Net will hold its Sixth Annual Tailgate Swapfest, Saturday, May 12, at the Deerfield, New Hampshire, Fairgrounds (covered buildings in case of rain). Admission: one dollar, no percentage or commission, commercial dealers welcome at same rate. Excess revenues benefit Boston Burns Unit of Shriners' Hospital for Crippled Children. Last year we donated $1140.50. Talk-in .52 and 146.40-147.00. Questions about New England's biggest flea market? SASE to Joe DeMaso K1RQG, Star Route Box 56, Bucksport, ME 04416, or Norm Blake, WA1IVB, P.O. Box 32, Cornish, ME 04020.

NORTH CAROLINA: The Raleigh Amateur Radio Society's 7th annual hamfest, April 22, 9:00 AM, Crabtree Valley Mall, US 70 West, Raleigh. $3.00 admission. Many prizes. First prize: Kenwood TS520S or ICOM 211 and others. Covered flea market, nearby motels and restaurants, hospitality room Saturday night. Talk-in Saturday and Sunday, WR4ACF (146.04/146.64) and WR4AOE (146.28/146.88). For info: RARS Hamfest, P.O. Box 17124, Raleigh, NC 27609.

NORTH CAROLINA: The Durham F.M. Association's annual Durhamfest, Saturday and Sunday, May 19 and 20, South Square Mall, Durham. Prizes, exhibits, programs. Shopping malls for XYLs. Sunday, ladies's bingo. Free tailgating spaces, covered flea market. $3.00 general ticket. Vendors and dealers included. Harmonics and unlicensed XYL's free. Power available. Talk-in 147.825-.225, 146.34-.94, 222.34-3.94. For info: DFMA, Box 8651, Durham, NC 27707.

OHIO: F.M. B*A*S*H, Dayton, April 27, 1979, Friday night of DAYTON HAMVENTION. Social evening for hams and friends, 8 PM til midnight, Dayton Convention Center, Main at Fifth Street. Admission free. Sandwiches, snacks and C.O.D. bar available. TV personality Rob Reider, WA8GFF, and his group will present a super show and there will be drawings for many fabulous prizes including a complete Drake UV-3 with 144, 220 and 440 MHz synthesized modules, power supply, encoder mike and antenna. For further information contact: Miami Valley F.M. Assn., P.O. Box 263, Dayton, Ohio 45401.

OHIO: The Dayton Amateur Radio Association's Hamvention, April 27, 28 and 29, Dayton Hara Arena and Exhibition Center. Prizes, contests, banquets. "Radio Amateur of the Year" award. Flea market hours: Saturday — 6 AM to 5 PM. Sunday — 6 AM to 4 PM. Own tables. Space: $10.00 ea. advance, $12.00 ea. door. Maximum four spaces. For additional information, reservations, accommodations: Dayton Hamvention, P.O. Box 44, Dayton, OH 45401. (513) 293-0459.

PENNSYLVANIA: Fifth Annual Northwest PA hamfest, Saturday, June 9, 1979, Crawford County Fairgrounds, Meadville, PA. Note date change. Gates open at 8:00 AM. Bring your own tables. $2.00 in; $1.00 out to display. $2.00 admission. Hourly door prizes, refreshments, commercial displays welcome. Talk-in 04/64, 81/21, 63/03. Details C.A.R.S., P.O. Box 653, Meadville, PA 16335. Attention: Hamfest Committee.

PENNSYLVANIA: The Annual Reading Radio Club hamfest, Sunday, May 27, 9:00 AM, Hamburg Field House, Hamburg. Door prizes, food, tailgate sales, dealer space available, rain or shine. Talk-in on 31/91 and 146.52. For info: Reading Radio Club, P.O. Box 124, Reading, PA 19603.

PENNSYLVANIA: The Penn Wireless Association's Tradefest '79, April 1, 8:00 AM. (7:00 AM for setup), National Guard Armory, Southampton Road and Roosevelt Blvd., Philadelphia. Display space $3.00 ea. includes one chance towards prizes.

PENNSYLVANIA: The Warminster Amateur Radio Club's Fifth Annual Ham-Mart Flea Market and Auction, Sunday, May 6, 9:00 AM - 4:00 PM, William Tennent Intermediate High School, Street Road (Route 132) Warminster, Bucks County. Registration: $1.00 per car (includes one door prize ticket). Tailgating: $2.00 add. Indoor tables: $3.00 ea. Talk-in on 146.16/76 and 146.52 MHz. For info: Horace Carter, K3KT, 38 Hickory Lane, Doylestown, PA 18901. (215) 345-6816.

PUERTO RICO: The Radio Club de Puerto Rico's annual convention and hamfest, April 28 and 29, Condado Holiday Inn Hotel, San Juan. For details: GPO, Box 693, San Juan, PR 00936.

WASHINGTON: The Skagit Amateur Radio Club's 27th annual convention, April 20, 21 and 22, Bryant Grange Center near Seattle. For info: Norman G. Ray, Program Chairman, 14005 - 132nd Avenue N.E., Kirkland, WA 98033.

WISCONSIN: 3-F Amateur Radio Club's annual Swapfest, Saturday, May 5, 8 AM to 3 PM, Neenah Labor Temple, 157 South Greenbay Rd., Neenah. Large parking, indoor swap areas, free auction. Refreshments available. Tickets: $1.50 advance, $2.00 door; tables $1.50 advance, $2.00 door. Reservations to: Mark Michel, W9OP, 339 Naymut St., Menasha, WI 54952.

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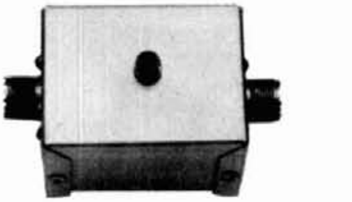

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(3.5 + 5.5 + 7.5 INCHES)

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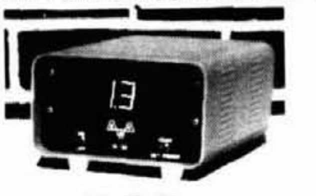

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THE MODEL SL-65* (20-2000 WATTS) AND THE QRP MODEL SL-65A* (0.2-20 WATTS) DIGITALLY INDICATE ANTENNA VSWR UNDER ANY TRANSMISSION MODE -- SSB, CW, RTTY, AM Etc. THERE IS NO CALIBRATION REQUIRED AND NO CROSSED METER NEEDLES TO INTERPRET. SIMPLY LOOK AT THE READOUT AND THAT IS THE VSWR. SPEAKING NORMALLY INTO A SSB TRANSMITTER MIC. INSTANTLY CAUSES THE VSWR TO BE DISPLAYED THROUGHOUT YOUR ENTIRE TRANSMISSION. REVERSING THE POSITION OF A FRONT PANEL TOGGLE SWITCH AND THE DISPLAY INDICATES THE NET POWER *(FORWARD LESS REFLECTED) THAT IS ACCEPTED BY THE ANTENNA. THE PEAK* OF THE NET PEP IS DETECTED AND DISPLAYED WITHOUT FLICKER FOR ANY MODULATION TYPE. DISPLAY UPDATE IS CONSTANT YET FLICKER FREE AS YOU MAY CHANGE THE POWER ACCORDING TO YOUR VOICE. THERE IS NOTHING LIKE THIS QUALITY INSTRUMENT AVAILABLE ANYWHERE ELSE. IT IS THE ONLY VSWR-NET POWER INDICATOR THAT LETS YOU KNOW THE STATE OF YOUR ANTENNAS AND TRANSMITTED POWER AT ALL TIMES WHILE TRANSMITTING. EITHER MODEL IS A SOPHISTICATED DEVICE CONTAINING FOUR CIRCUIT BOARDS AND THIRTEEN INTEGRATED CIRCUITS.

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SL-65
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- TWO DIGIT DISPLAY SHOWS VSWR *TO AN ACCURACY OF .1 FOR VALUES* FROM 1.0 AND 2.2. ACCURACY IS TO .2 FOR VALUES FROM 2.3 TO 3.4 AND TO .3 FROM 3.4 TO 4.0. FROM 4.1 TO 6.2 THE INDICATION MEANS THAT VSWR IS VERY HIGH.

- FOR VSWR VALUES NEAR 1.0, THE POWER RANGE FOR A VALID READING IS 20 - 2000 WATTS OUTPUT. FOR HIGHER VALUES THE UPPER POWER LIMIT FOR A FLICKER FREE VALID READING IS SOMEWHAT LESS (35 - 1000 WATTS FOR VSWR AT 2.0).

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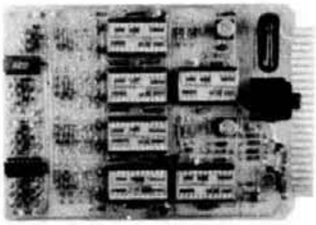

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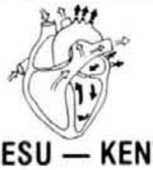

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FOX-TANGO CRYSTAL FILTER SPECIFICATIONS

		FILTER NO YF-	USED FOR	CENTER FREQUENCY kHz	BANDWIDTH — Hz		INSERTION LOSS dB	TERMINAL Z (IN-OUT) Ω/pF	CASE SIZE SEE BELOW	SEE NOTE
					−6 dB	−60 dB				
YAESU	FT-101 FR-101	31H250	CW-N	3179.3	250 ± 50	<750	<9	500Ω	A	1
		31H500	CW	3179.3	500 ± 50	<1200	<7	500Ω	A	2
		31H1.8	SSB-N	3180	1800 ± 100	<3500	<6	500Ω	B	3
		31H2.4	SSB-F	3180	2400 ± 100	<4200	<6	500Ω	B	4, 6
		31H6.0	AM	3180	6000 ± 500	<11K	<6	500Ω	A	5
	FT-301 FT-7	91H250	CW-N	8999.3	250 ± 50	<750	<10	500Ω	C	1
		91H500	CW	8999.3	500 ± 50	<1400	<8	500Ω	C	2
		90H1.8	SSB-N	9000	1800 ± 100	<3500	<6	500Ω	C	3
		90H2.4	SSB-F	9000	2400 ± 100	<4200	<6	500Ω	C	6
	FT-901 FT-101Z	89H250	CW-N	8988.3	250 ± 50	<750	<10	500Ω	C	1
		89H500	CW	8988.3	500 ± 50	<1400	<8	500Ω	C	2
KENWOOD	TS-520 R-599	33H250	CW-N	3395	250 ± 50	<750	<9	4.7K/33pF	B	1
		33H400	CW	3395	400 ± 50	<1200	<8	4.7K/33pF	B	7
		33H1.8	SSB-N	3395	1800 ± 100	<3500	<6	4.7K/33pF	B	3
	TS-820	88H250	CW-N	8830.7	250 ± 50	<750	<10	470/5pF	D	1
		88H400	CW	8830.7	400 ± 50	<1400	<9	470/5pF	D	7
		88H1.8	SSB-N	8830	1800 ± 100	<3500	<6	470/5pF	D	3
HEATH	All Except SB/HW 104	33H250	CW-N	3395.4	250 ± 50	<750	<9	2KΩ	E	1
		33H400	CW	3395.4	400 ± 50	<1200	<8	2KΩ	E	8
		33H2.1	SSB	3395	2100 ± 100	<3500	<6	2KΩ	E	9
DRAKE	R-4C	56H8.0	CW/SSB	5645 ± 150	8000 ± 500	<13K	<3	500Ω	F	10, 11
		56H800	CW	5645 ± 150	800 ± 100	<1800	<5	500Ω	C	10, 12
		56H125	CW-N	5695	125 ± 50	<350	<13	50Ω	G	10
$	All filters, except Drake, are $55 each. For Drake prices, see ad page 43. Prices include airmail postpaid to U.S., Canada and Mexico. For overseas airmail, add $3									

CASE SIZES
(LWH in mm)

A	45 × 23 × 28
B	50 × 25 × 25
C	40 × 20 × 21
D	50 × 18 × 18
E	57 × 24 × 25
F	57 × 24 × 18
G	57 × 24 × 21

NOTES

1. Sharp CW Filter for DX and Contest work
2. Use instead of optional 600 Hz (6 pole) unit
3. For narrow SSB to reduce QRM
4. Use instead of XF-30A (6 pole) in early units
5. Superior to XF-30B (6 pole) AM unit
6. Used in G3LLL RF Speech Processors
7. Use instead of optional 500 Hz (6 pole) unit
8. Use instead of optional 400 Hz (4 pole) unit
9. Superior replacement for standard SSB unit
10. For detailed description and prices see ad page 43
11. Also known as GUF-1
12. Also known as GUF-2S

Dealer Inquiries Welcomed

VISA & MASTER CHARGE Accepted

Florida residents add 4% (sales tax)
See our other ad page 43 this issue

FOX-TANGO CORP.
Box 15944, W. Palm Beach, FL 33406

Advertisers ✓ check-off

...for literature, in a hurry — we'll rush your name to the companies whose names you **"check-off"**

Place your check mark in the space between name and number. **Ex: Ham Radio ✓ 234**

INDEX

AED ____ 710
Abletronics ____ 770
Alliance ____ 700
Aluma ____ 589
Antek ____ 733
Astron ____ 734
Atlas ____ 198
Barrett *
Barry *
Bauman ____ 017
Budwig ____ 233
Bullet ____ 328
Cal Crystal ____ 709
CoaxProbe ____ 726
Communications Center ____ 534
Comm. Research ____ 753
Comm. Spec. ____ 330
Creative Elec. ____ 751
Crystal Banking ____ 573
Curtis Electro ____ 034
Cushcraft *
D&V Radio *
DCO ____ 324
DSI ____ 656
Dames Comm. ____ 551
Data Signal ____ 270
Drake *
E.T.O. *
Electrocom ____ 663
Electronic Farm *
Elec. Research Virginia *
Erickson *
Fox-Tango ____ 657
G & C Comm. ____ 754
GLB ____ 552
GL Enterprises *
Gilfer ____ 207
Gray ____ 055
Gregory *
Group III ____ 701
Gull ____ 635
Hal *
Hal-Tronix ____ 254
H. R. B. ____ 150
Heath ____ 060
Henry ____ 062
Hy-Gain ____ 064
Icom *
Integ. Circuits ____ 518
Int. Crystal ____ 066
Jameco ____ 333
Jan ____ 067
Jones ____ 626
Kantronics *
Kenwood *
Long's ____ 468
Lunar ____ 577
MFJ ____ 082
Madison *
Microwave Filter ____ 637
J. W. Miller ____ 745
Milo ____ 736
Monroe ____ 715
Oak Hill Acad. A.R.S. *
P.C. Elec. ____ 766
Palomar Eng. *
Port. Comm. Sup. ____ 769
Callbook ____ 100
Radio Elec. ____ 767
Radio Sys. Tech. ____ 422
Radio World *
Ramsey ____ 442
Rockwell Collins ____ 258
S-F A. R. S. ____ 640
SST ____ 375
Sabtronics *
Howard W. Sams ____ 104
Schober Organ ____ 762
Securitron ____ 461
Sherwood ____ 435
Slep ____ 232
Space ____ 107
Spec. Int. ____ 108
Sunset Elec. ____ 768
Swan ____ 111
TPL ____ 240
Telrex ____ 377
Ten-Tec *
Texas RF Distr. ____ 763
The Communication Center *
Thomas Comm. ____ 730
Tri-Ex ____ 116
VHF Eng. ____ 121
Van Gorden ____ 737
Vanguard Labs ____ 716
Varian ____ 043
Webster Assoc. ____ 423
Western *
Whitehouse *
Willcomp ____ 764
Wilson ____ 123
Yaesu ____ 127

*Please contact this advertiser directly.

Limit 15 inquiries per request.

April, 1979

Please use before May 31, 1979

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HAM RADIO MAGAZINE — "check off"
Greenville, N. H. 03048

NAME ....

CALL ....

STREET ....

CITY ....

STATE .... **ZIP** ....

Advertisers index

AED Electronics 116
Abletronics 117
Alliance Mfg. Company 63
Aluma Tower Co. 96
Antek, Inc. 98
Astron Corporation 96
Atlas Radio 31
Barrett Electronics 110
Barry Electronics 90
R.H. Bauman Sales Co. 102
Budwig Mfg. Co. 115
Bullet 102
Cal Crystal Lab, Inc. 98
CoaxProbe Co. 114
Communications Center 105
Communication Research, Inc. 98
Communications Specialists 10, 11
Creative Electronics 100
Crystal Banking Service 100
Curtis Electro Devices 104
Cushcraft 64, 65
D & V Radio Parts 115
DCO, Inc. 102, 116
DSI Instruments 54, 55
Dames Communications Systems 108
Data Signal, Inc. 88
Drake Co., R.L. 67
Ehrhorn Technological Operations 119
Electrocom Industries 94
The Electronic Farm 114
Electronic Research Corp. of Virginia 115
Erickson Communications 108
Fox-Tango Corp. 43, 117
G & C Communications 100, 115
GLB 98
GL Enterprises 49
Gilfer Associates 84
Gray Electronics 100
Gregory Electronics 52
Group III Sales Company 110
Gull Electronics 103
Hal Communications Corp. 22
Hal-Tronix 25
Ham Radio's Bookstore 117
Heath Company 59
Henry Radio Stores Cover II
Hy-Gain Electronics 1
Icom 5
Integrated Circuits Unlimited 111
International Crystal 89
Jameco Electronics 107
Jan Crystals 96
Jones, Marlin P. & Assoc. 106
Kantronics 72
Trio-Kenwood Communications, Inc. 7, 60, 61
Long's Electronics 120
Lunar Electronics 79
MFJ Enterprises 2
Madison Electronic Supply 43, 52, 94, 118
Microwave Filter, Inc. 116
J.W. Miller Division, Bell Industries 90
Milo Associates 110
Monroe Electronics, Inc. 90
Oak Hill Academy Amateur Radio Session 106
P.C. Electronics 106
Palomar Engineers 84
Portable Communications Supply 116
Radio Amateur Callbook 104
Radio Electronics 101
Radio Systems Technology, Inc. 114
Radio World 114
Ramsey Electronics 91
Rockwell International, Collins Division 77
S-F Amateur Radio Service 116
SST Electronics 94
Sabtronics International, Inc. 97
Howard W. Sams & Co., Inc. 109
Schober Organ Corp. 104
Securitron 108
Sherwood Engineering 114
Slep Electronics 110
Space Electronics 115
Spectrum International 83
Sunset Electronics 114
Swan Electronics 95
TPL Communications 49
Telrex Laboratories 96, 102
Ten-Tec 9
Texas RF Distributors 99
The Communication Center 114
Thomas Communications 44, 45
Tri-Ex Tower Corporation 73
VHF Engineering, Div. of Brownian 85
Van Gorden Engineering 110
Vanguard Labs 90
Varian, Eimac Division Cover IV
Webster Associates 84
Western Electronics 117
Whitehouse, G.R. & Co. 100
Willcomp, Inc. 115
Wilson Electronics 53
Yaesu Electronics Corp. Cover III

The Sangre de Cristo Mountains - ETO'S backyard

WHY WISH YOU'D BOUGHT ALPHA?

IS SOMETHING ELSE "JUST AS GOOD?"

New **ALPHA** owners often tell us, *"I wish I'd saved my time and money and bought an* **ALPHA** *in the first place."* Why not benefit from their experiences? Compare first!

TRY TO GET ANY OTHER MANUFACTURER TO TELL YOU - IN WRITING - THAT IT'S SAFE TO OPERATE HIS DESK TOP LINEAR AT A FULL D-C KILOWATT . . . SAY FOR 24 HOURS KEY-DOWN. OR, ASK HIM FOR A FULL YEAR WRITTEN WARRANTY. LOTS OF LUCK!

YOUR NEW *ALPHA* WILL HAPPILY AND COOLY RUN THAT KILOWATT KEY-DOWN . . . FOR 24 DAYS IF YOU WISH. AND YOU'LL BE PROTECTED BY ETO'S UNMATCHED WARRANTY FOR TWO YEARS. WE PUT IT IN WRITING ALL THE TIME. IT'S THE WAY WE BUILD AND WARRANT EVERY *ALPHA*!

The new **ALPHA's** are the best we've ever built. Nothing else even approaches an **ALPHA's** combination of power, convenience, quality, and owner protection. The ETO/**ALPHA** two year limited warranty offers you eight times as much protection as the industry-*standard 90 day warranty.*

The new **ALPHA 374A** adds NO-TUNE-UP operation to all the other traditional ALPHA qualities and capabilities. You can hop instantly from one HF band to another, with full maximum legal power and with little or no amplifier tune-up at all! (If new amateur bands are added, you can manually adjust your **ALPHA** to work them, too.)

In 1974 the original **ALPHA 374** set a standard of high power convenience that has remained unmatched since. Despite its small size, not even one '374 owner ever burned out a power transformer. Impressive? The new '374A has an even huskier power supply. And it *has ETO's ducted-air system* with acoustically-isolated centrifugal blower to insure cool, whisper-quiet operation.

Before you get serious about any other brand of linear, compare its convenience and quality, its transformer heft, its cooling system efficiency and noise level - and its warranty - with the **ALPHA's.** Be sure to ask around about its reputation.

Call or write for detailed literature and thoroughly check out all the great new **ALPHA's** . . . so you don't make a mistake.

ETO

EHRHORN TECHNOLOGICAL OPERATIONS, INC.
BOX 708, CAÑON CITY, CO 81212 (303) 275-1613

MOBILE VALUES FROM LONG'S

1-800-633-3410

KENWOOD TR-7625 25 watt 2m transceiver

Featuring: memory channel, mode switch for simplex or switching transmit freq. up or down, full 4 MHz coverage, 800 channels, 5 kHz offset switch, MHz selector switch, LED, unlock indicator.

425.00 Call for quote.

ICOM IC-211 2m transceiver

• 144 to 148 MHz coverage • Modes: SSB, CW, FM • LSI synthesizer PLL • 4-digit LED readout • Pulse-type noise blanker • VOX w/adjustable gain • SWR bridge • CW monitor • Automatic power control • AC/DC power supplies • Antenna impedance: 50 ohms unbalanced • TX output: 10W PEP.

899.00 Call for quote.

YAESU CPU-2500RK 2m FM *transceiver*

800 PLL channels, auto. scan over entire 2m band, 4 memories, tone burst, 3/25 watts, fixed ± 600 KHz programmable offsets, 13.8 VDC at 8 amps continuous. *Comes with keyboard mic for remote* input, scanning control, aux. repeater split, and 2 tone input for auto patch or control link.

585.00 Call for quote.

YAESU FT-227RA The "Memorizer" transceiver

One knob channel selection for 800 channels, frequency 144-148 MHz, 4-digit LED readout, fully synthesized frequency control, selectable 10 watt Hi/1 watt Low output, 4 memories, touch control on mic for scanning, scan selectable for clear or busy channels.

399.00 Call for quote.

KENWOOD TS-700SP all mode 2m transceiver

Covers the entire 2 meter band. The features are SSB, CW, FM, AM and semi-break-in CW. Also side tone monitor digital freq. readout, receiver preamp and 600 KHz repeater offset op (144.5-145.5 Hz) 10 watts RF output, AC/DC pow. supply.

799.00 Call for quote.

TUCH COM 1215 encoding mic

500 ohm low impedance mic, needing no battery or *modification to use with the* ICOM 22S, 245, 211, 215, and 280.

49.00 Call for yours today!

CUSHCRAFT AMS-147 FM mobile 2m mag *mount antenna*

A ⅝ wave length antenna with 3 dBi gain, low SWR operation over 146-148 MHz FM band. Move center freq. ± 1.5 MHz. Matched *to 50 ohms.* Built in connector takes PL-259.

34.95 Call for yours today.

ICOM IC-280 remotable 2m FM mobile

Frequency 143.90-148.11 MHz • Power: 10 watts Hi, 1 watt adj. Low • Power requirements: 13.8 VDC at 2.5 amps • Main PLL control head may be detached and remotely mounted • With microprocessor, stores 3 frequencies • Easy to read LED's.

480.00 Call for quote.

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MAIL ORDERS: P.O. BOX 11347 BIRMINGHAM, AL 35202 • STREET ADDRESS: 2808 7TH AVENUE SOUTH BIRMINGHAM, ALABAMA 35233

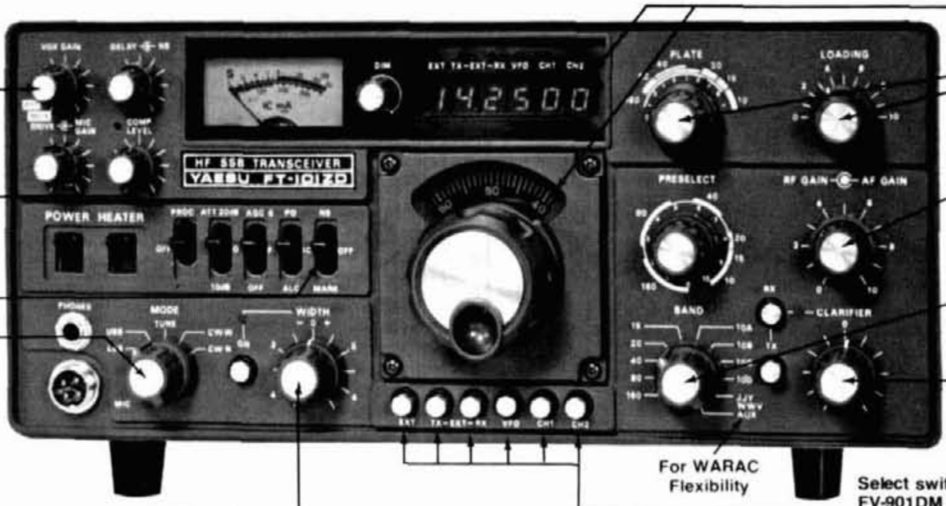
Remember, you can Call Toll Free: **1-800-633-3410** in the U.S.A. or call **1-800-292-8668** in Alabama for our low price quote. Store hours: 9:00 AM til 5:30 PM, Monday thru Friday.

ALL NEW

FT-101ZD

HIGH-PERFORMANCE HF TRANSCEIVER

Today's technology, backed by a proud tradition, is yours to enjoy in the all-new FT-101ZD transceiver from YAESU. A host of new features are teamed with the FT-101 heritage to bring you a top-dollar value. See your dealer today for a "hands on" demonstration of the performance-packed FT-101ZD.



ast front panel, plus heavy
case

lt-in, fully adjustable, VOX
uitry

lt-in RF speech processor
nore "talk power" when you
d it

lt-in, threshold adjustable,
se blanker

ipped for SSB and CW
ration. Choice of wide or
row bandwidth for CW (with
ional CW filter installed)

Continuously variable IF bandwidth: 300 Hz to 2.4 KHz

For WARAC Flexibility

Select switches for use with FV-901DM synthesized scanning VFO (option). FV-901DM provides scanners plus 40 frequency memory bank.

Digital plus analog frequency readout. Digital display resolution to 100 Hz

Rugged 6146B final amplifier tubes with RF negative feedback

RF and AF gain controls located on concentric shafts for operator convenience

Full band coverage: 160 through 10 meters, plus WWV/JJY (receive only)

TX, RX, or transceive frequency offset from main dial frequency

SPECIFICATIONS

TRANSMITTER

PA Input Power:
180 watts DC
Carrier Suppression:
Better than 40 dB
Unwanted Sideband Suppression:
Better than 40 dB @ 1000 Hz, 14 MHz
Spurious Radiation:
Better than 40 dB below rated output
Third Order Distortion Products:
Better than −31 dB
Transmitter Frequency Response:
300-2700 Hz (−6 dB)
Stability:
Less than 300 Hz in first 30 minutes after 10 min. warmup; less than 100 Hz after 30 minutes over any 30 min. period
Negative Feedback: 6 dB @ 14 MHz
Antenna Output Impedance:
50-75 ohms, unbalanced

GENERAL

Frequency Coverage:
Amateur bands from 1.8-29.9 MHz, plus WWV/JJY (receive only)
Operating Modes:
LSB, USB, CW
Power Requirements:
100/110/117/200/220/234 volts AC, 50/60 Hz; 13.5 volts DC (with optional DC-DC converter)
Power Consumption:
AC 117V: 75 VA receive (65 VA HEATER OFF) 285 VA transmit; DC 13.5V: 5.5 amps receive (1.1 amps HEATER OFF), 21 amps transmit
Size:
345 (W) × 157 (H) × 326 (D) mm
Weight:
Approximately 15 kg.

COMPATIBLE WITH FT-901DM ACCESSORIES

RECEIVER

Sensitivity:
0.25 uV for S/N 10 dB
Selectivity:
2.4 KHz at 6 dB down, 4.0 KHz at 60 dB down (1.66 shape factor); Continuously variable between 300 and 2400 Hz (−6 dB); CW (with optional CW filter installed): 600 Hz at 6 dB down, 1.2 KHz at 60 dB down (2:1 shape factor)
Image Rejection:
Better than 60 dB (160-15 meters); Better than 50 dB (10 meters)
IF Rejection:
Better than 70 dB (160, 80, 20-10 m); Better than 60 dB (40 m)
Audio Output Impedance:
4-16 ohms
Audio Output Power:
3 watts @10% THD (into 4 ohms)

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arma
amateur radio manufacturer's association

Price And Specifications Subject To Change Without Notice Or Obligation

YAESU
The radio.

YAESU

379X

YAESU ELECTRONICS CORP., 15954 Downey Ave., Paramount, CA 90723 ● (213) 633-4007
YAESU ELECTRONICS Eastern Service Ctr., 9812 Princeton-Glendale Rd., Cincinnati, OH 45246

Two EIMAC 3-500Zs provide the punch in Kenwood's new amplifier.

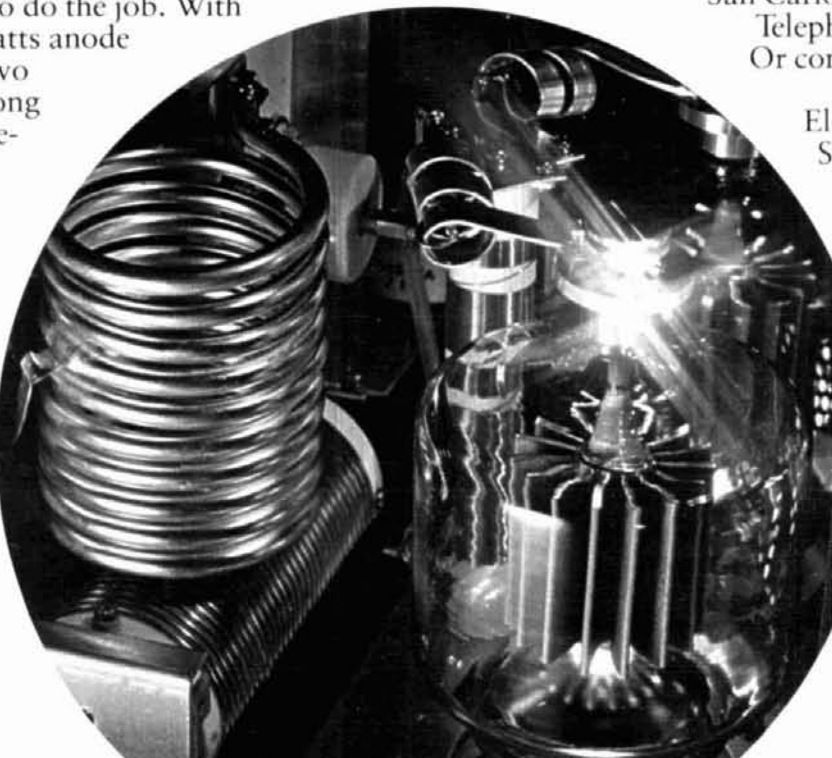
Kenwood chooses EIMAC for trouble-free service.

The new heavy-duty Kenwood TL-922A linear amplifier provides 2 kW PEP input for SSB service and 1 kW input for CW, RTTY, and SSTV operation.

Kenwood chose two EIMAC 3-500Z high-mu triodes to do the job. With a total of 1000 watts anode dissipation, the two 3-500Zs coast along to provide trouble-free, long-life service.

For more information

Send for the EIMAC Quick Reference catalog covering the complete line of EIMAC products and for the 3-500Z Data Sheet. Learn why the important manufacturers of communication equipment choose EIMAC. Varian, EIMAC Division, 301 Industrial Way, San Carlos, California 94070. Telephone (415) 592-1221. Or contact any of the more than 30 Varian Electron Device Group Sales Offices throughout the world.



What's your pleasure?

DX chasing? Traffic nets? RTTY? Rag chewing? SSTV? The EIMAC 3-500Z provides the power when you need it, with ample safety margin. Value wise amateurs always look for the EIMAC power tube for reliability. And equipment manufacturers, such as Kenwood, choose EIMAC for leadership in power tube technology.

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