one dollar

ham radio

magazine

AUGUST 1976

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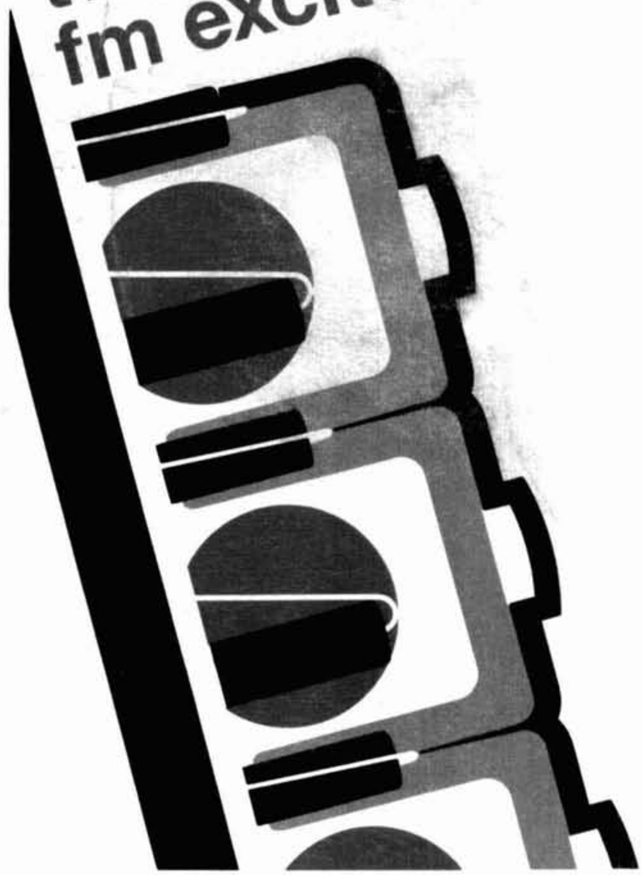
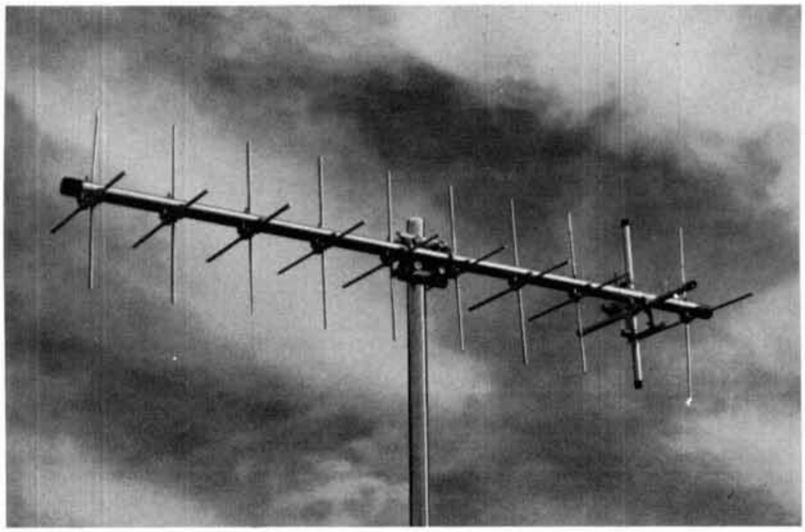
- and much more . . .

high-performance two-meter fm exciter



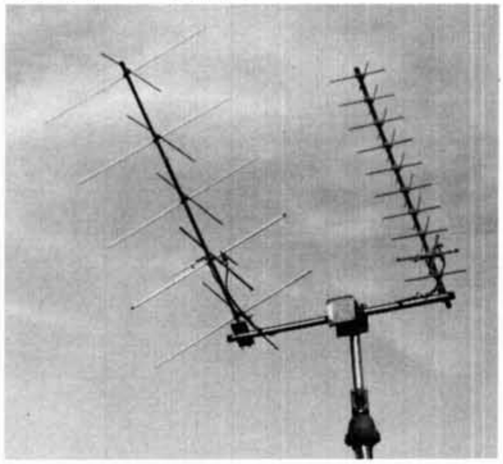
2 METERS...



. . . Alive with activity at *both* ends of the band! Be a part of the total 2 Meter picture with the Cush Craft Twist Antenna. Actually two, easily assembled, 10 element yagis in one — the vertical elements are cut for the high end, the horizontal elements for the low end, and separate feed lines are used. The A147-20T is tailored to meet the demands of the operator who enjoys the best of both worlds — FM and SSB/CW.

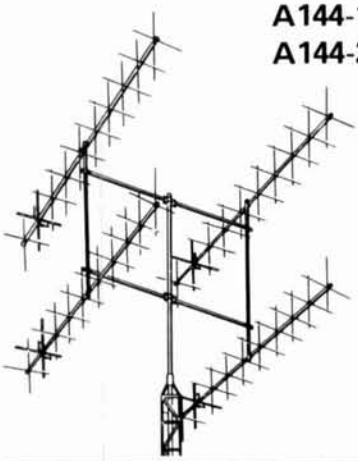
A147-20T $47.50

CQ OSCAR .... ..



. . . Extend your horizon — Explore the exciting new world of amateur satellite communications using low power with our multi-polarized Twist Antennas. All models include phasing harness for selectable linear or right/left circular polarization. Two of these Twists may be mounted on the A14T-MB mounting boom which is complete with a pre-drilled plate for a readily available mast-through rotator. *Face this challenging frontier—Become a Specialist!*

A144-10T $32.50 A432-20T $45.00
A144-20T $47.50 A14T-MB $13.95



PERFORMANCE ARRAYS . . .

Enjoy fade-free contacts on VHF/UHF with Twist Antennas and Arrays. Excellent for scatter and other long-haul techniques. Double your effective radiated power by stacking two Twists, or quadruple ERP by stacking four Twists. Arrays are easily assembled for your special communications requirement. Write for stacking and phasing harness details concerning amateur and commercial frequencies.

Dependable communications — *Now Yours!*

SPECIFICATIONS

Model	A147-20T	A144-10T	A144-20T	A432-20T
Center Freq. (MHz)	144.5/146.5	145.9	145.9	432
No. Elements	10/10	10	20	20
Weight (lbs.)	6	3.5	6	3.5
Wind Surf. Area (ft.²)	1.42	.74	1.42	.37
Mounting	Center	Rear	Center	Rear
Dimensions (Inches)	40x40x140	40x40x70	40x40x140	14x14x57
Front-to-Back Ratio (dB)	22	22	22	22
Forward Gain (dBd) circular	—	10.8	13.6	13.6
linear	12.4	9.6	12.4	12.4

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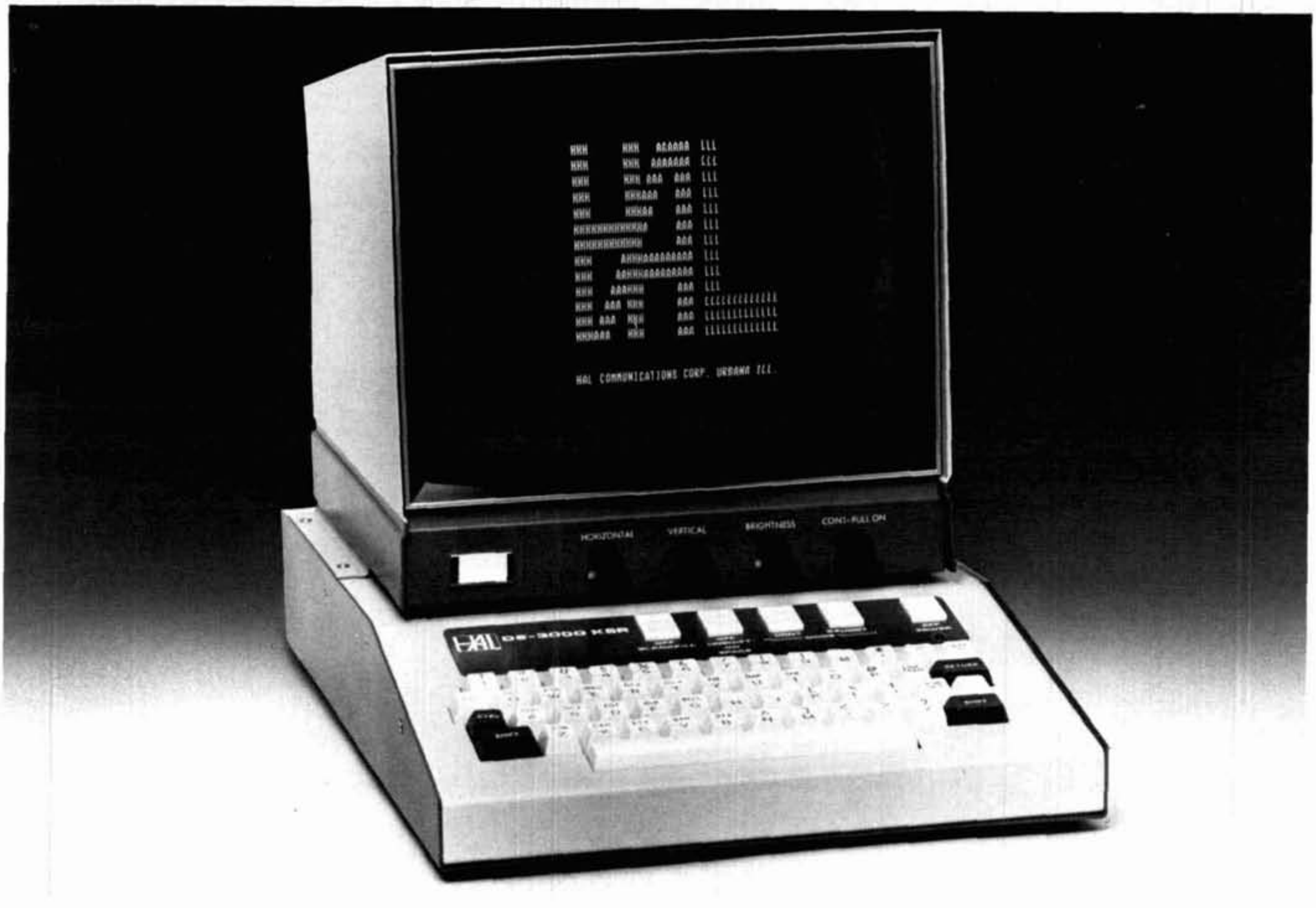
More Details? CHECK—OFF Page 126

Stay tuned for future programs.



The HAL ST-6000 demodulator /keyer and the DS-3000 and DS-4000 KSR/RO series of communications terminals are designed to give you superlative TTY performance today —and in the future. DS series terminals, for example, are re-programmable, assuring you freedom from obsolescence. Sophisticated systems all, these HAL products are attractively priced—for industry, government and serious amateur radio operators.

The HAL ST-6000 operates at standard shifts of 850, 425, and 170 Hz. The tone keyer is crystal-controlled. Loop supply is internal. Active filters allow flexibility in establishing different tone pairs. You can select AM or hard-limiting FM modes of operation to accommodate different operating conditions. An internal monitor scope (shown on model above) allows fast, accurate tuning. The ST-6000 has an outstandingly high dynamic range of operation. Data I/O can be RS-232C, MIL-188C or current loop.

The DS-3000 and DS-4000 series of KSR and RO terminals provide silent, reliable, all-electronic TTY transmission and reception, or read-only (RO) operation of different combinations of codes, including Baudot, ASCII and Morse. The powerful, programmable 8080A microprocessor is included in the circuitry to assure maximum flexibility for your present needs —and for the future. The KSR models offer you full editing capability. The video display is a convenient 16-line format, of 72 characters per line.

These are some of the highlights. The full range of features and specifications for the ST-6000 and the DS series of KSR and RO terminals is covered in comprehensive data sheets available on request. Write for them now—and tune in to the most sophisticated TTY operation you can have today... or in the future.

HAL

HAL Communications Corp., Box 365, 807 E. Green Street
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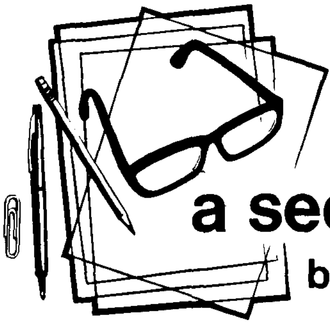

a second look

by Jim Fisk

> "At this very minute, with almost absolute certainty, radio waves sent forth by other intelligent civilizations are falling on earth. A telescope can be built that, pointed in the right place, and tuned to the right frequency, could discover these waves. Someday, from somewhere out among the stars, will come the answers to many of the oldest, most important, and most exciting questions mankind has asked."
>
> **Frank D. Drake**
> *Intelligent Life in Space*

In the late 1950s, soon after the United States became involved in a concentrated effort to place an astronaut into orbit around the earth, several scientific groups began to think seriously about using radio telescopes to search for extraterrestial intelligence. Inspired by articles in such prestigious magazines as *Scientific American* and *Sky and Telescope*, amateur radio astronomers began taking part in Project Ozma, pointing their antennas toward the heavens, looking for radio signals from intelligent beings in outer space. They didn't know where to look, nor what frequency to tune, so it's not surprising that when nothing was found for several years, enthusiasm began to wane. Then an English radio astronomer discovered the first pulsar, many scientists postulated that it might be a controlled radio beam from intelligent life in another galaxy, and a whole new search for extraterrestial communications was soon underway.

Unfortunately, there was no organized follow-up to Project Ozma, so it's not known today how many individuals or groups are still listening, nor who and where they are. Nick Marshall, W6OLO, a member of the very first Oscar group back in 1960, became interested in this problem and announced the formation of a "Reviva Ozma" committee at the 1975 Project Oscar meeting at Foothill College in California. One of the goals of Reviva Ozma (now known as Starquest) would be to locate and communicate with individuals and groups who were still listening, and to help them publish their findings. Another goal was to assemble a state-of-the-art listening post which would be dedicated to continuing the search for radio signals of extraterrestial origin.

In late 1975 Marshall contacted Dr. John Billingham at the NASA-Ames Research Center for further information on the Project Cyclops follow-on, a NASA study of a system for detecting intelligent extraterrestial life. It became immediately apparent that Starquest could serve an interim purpose by disseminating correct information about Project Cyclops and, more importantly, assist in trying to keep various listening frequencies clear of occupancy by commercial, industrial, and government radio services.

Marshall also made contact with members of Operation Serendip, a study being conducted by a small group from the University of California at Berkeley. This group has been using an 82-foot (25m) dish near Mt. Lassen to listen and record extraterrestial signals on 1420 MHz and process the data on a pdp-8 computer.

Foothill College graciously offered the Starquest group an antenna location near their observatory, and work is now underway on a spherical dish which is patterned after the 1000-foot (305m) dish at Arecibo, Puerto Rico, although on a much smaller scale. The initial Starquest dish will have a 35- to 50-foot (10.7m-15.2m) center section which can be expended later to about 100 feet (30m) in diameter. Starquest members are also formulating plans for a state-of-the-art receiving and data-processing system which will be used in the station. In the near future the group is expected to start publishing a quarterly *Starquest Bulletin* which will be mailed to members and affiliates.

In addition to their other activities, Starquest has established a world-wide amateur-radio net on 20 meters for gathering and disseminating information relating to the search for extraterrestial signals. This net, which meets on 14.280 MHz (±5 kHz) at 1900 GMT, the first Sunday of every month, is now operational. It is hoped that interested amateurs and experimenters will listen in, and help Starquest locate other interested groups in their own local areas who are involved in the search for intelligence in outer space. Amateurs living in Northern California who would like to participate directly in the construction of the Starquest listening station are invited to contact W6OLO for more information.

Jim Fisk, W1DTY
editor-in-chief

Now It's Crystal Clear

Yes, now ICOM helps you steer clear of all the hassles of channel crystals. The new **IC-22S** is the same surprising radio you've come to know and love as the **IC-22A**, except that it is totally crystal independent. **Zero crystals.** Solid state engineering enables you to program 23 channels of your choice without waiting. Now the ICOM performance you've demanded comes with the convenience you've wanted, with your new **IC-22S.**

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More Details? CHECK—OFF Page 126

presstop

FCC'S FIRST ACTION ON RESTRUCTURING was taken in mid June, and Novices and Technicians became the principal beneficiaries. This first Report and Order on Docket 20282 liberalizes both license classes' privileges but appears to reject the "Dual Ladder" approach to restructuring that was such an integral part of the Docket's original philosophy.

Technician Class Licensees receive the most significant benefit in the form of full Novice privileges — CW on 80, 40, 15 and 10 meters. No other Technician privileges were changed at this time, so it's still possible that 10-meter phone and/or expansion of Technician frequencies on the bottom of two or six meters may be part of expected later actions on 20282. Technician exams will no longer be "mail order," either — the Report and Order makes administration of Tech exams an FCC Field Office job.

Novices Receive two benefits under the just announced Rules changes. Power limits for Novices have been raised to 250 watts, and a Novice who has failed to upgrade during his two-year license term will be permitted to continue as a Novice by taking another Novice exam — the requirement for a one-year delay before going for another Novice license has been dropped. The new 250 watt Novice power limit actually applies to Novice frequencies, by the way — any station operating in the Novice sub bands will be limited to 250 watts, regardless of license class.

All "Mail Order" Amateur Exams except Novice are effectively being eliminated. The one exception that will remain is for an applicant who cannot travel because of medical reasons and who can back it up with a physician's statement.

Deletion Of Conditional Amateur Licenses won't shut the door on Americans overseas such as servicemen and missionaries who can't get to an FCC Field Office for their exams. Procedures to take care of such people are being developed and they will be taken care of on a case-by-case basis. One idea being considered is a "Conditional" license good only while the holder remains out of the country — he'd have to be reexamined by the FCC on return if he wished to continue his Amateur activities.

All Present Conditional and Technician (C) license holders will be "grandfathered" into straight Technician and General Class licenses as their present licenses come up for renewal. No re-examinations will be required, which should make at least a few present license holders breathe easier!

AMATEUR LICENSING TIMES have definitely dropped since our last report, with several recent renewals returned 37 to 45 days after they went into the mail. As of early July the FCC was handling requests for volunteer exams in about two weeks, down from six weeks in early June.

The Next Big Hurdle is the phase-in of WD calls, where computer programming problems have developed. Until they're overcome, significant delays in getting new licenses out are likely to begin to occur despite efforts to head the problems off.

AMATEUR RADIO IS VERY PROMINENT in the Smithsonian Institution's newest displays. OSCAR 1 is on display in the Communications Satellite area of the new Hall of Satellites, and NN3SI, the Institution's new Amateur station located in the Museum of History and Technology, was dedicated on June 8th. First contact from NN3SI was on CW with trustee W4KFC working W1AW with the same key General Sarnoff used in 1912 when he participated in the Titanic disaster.

NN3SI Is the Institution's special events call — WB3APS is its regular call — and NN3SI has been authorized for a year with operation on all bands and OSCAR planned.

Smithsonian's New Amateur Station can be a bit tricky to get to, warns K8NHR. It's almost directly under the antennas, and can be found by bearing hard left after entering the building's main entrance.

UNLICENSED 27 MHZ OPERATORS are protected by a legal loophole — the Communications Act of 1934 gives FCC jurisdiction only over licensees. Justice Department is the agency with the power to go after unlicensed operators, but it has neither the tools (as FCC does) nor time to chase down all such violators.

This "Catch 22" May Change shortly, however — FCC Chairman Wiley said at the CES show in Chicago in mid June that there is legislation pending before both houses of Congress that would give the FCC power to prosecute unlicensed operators and confiscate their equipment, and he expects it to become law very quickly.

TWO-LETTER CALLSIGN availability chart has been prepared by the ARRL from current FCC records and can be had for the asking. It covers all ten districts and includes instructions on the procedure for requesting a two-letter call. Send an SASE with 24¢ postage to ARRL for a copy.

MOONBOUNCE ENTHUSIASTS are going to get a crack at Alaska thanks to K6YNB/KL7. Wayne plans to open up from Ketchikan, August 11th on 144 and switch to 432 about the 18th. He also plans to be on 50 and 220 MHz, but not with the same big gun antennas that he'll use on the other bands. Meteor scatter (the Perseids shower occurs during that period) as well as possible tropo and Aurora contacts are also expected during the trip.

KENWOOD'S TS-520

...worth waiting for!

Why wait any longer for a rig that offers top performance, dependability and versatility . . . the TS-520 has proven itself in the shacks of thousands of discriminating amateurs, in field day sites, in DX and contest stations, and in countless mobile installations.

Superb craftsmanship is evident throughout . . . in its engineering concepts as well as its construction and styling . . . craftsmanship that is a Kenwood hallmark.

Maybe the Kenwood TS-520 is the one you have been waiting for.

Kenwood offers accessories guaranteed to add to the pleasure of owning the TS-520. The TV-502 transverter puts you on 2-meters the easy way. (It's completely compatibile with the TS-520.) Simply plug it in and you're on the air. Two more units designed to match the TS-520 are the VFO-520 external VFO and the model SP-520 external speaker. All with Kenwood quality built in.

TS-520 Specifications

MODES: USB, LSB, CW
POWER: 200 watts PEP input on SSB, 160 watts DC input on CW
ANTENNA IMPEDANCE: 50-75 Ohms, unbalanced
CARRIER SUPPRESSION: Better than −45 dB
UNWANTED SIDEBAND SUPPRESSION: Better than −40 dB
HARMONIC RADIATION: Better than −40 dB
AF RESPONSE: 400 to 2600 Hz (−6 dB)
AUDIO INPUT SENSITIVITY: 0.25μV for 10 dB (S+N)/N
SELECTIVITY: SSB 2.4 kHz (−6 dB), 4.4 kHz (−60 dB). CW 0.5 kHz (−6 dB), 1.5 kHz (−60 dB) (with accessory filter)
FREQUENCY STABILITY: 100 Hz per 30 minutes after warmup
IMAGE RATIO: Better than 50 dB
IF REJECTION: Better than 50 dB
TUBE & SEMICONDUCTOR COMPLEMENT: 3 tubes (2 x 6146B, 12BY7A), 1 IC, 18 FET, 44 transistors, 84 diodes
DIMENSIONS: 13.1" W x 5.9" H x 13.2" D
WEIGHT: 35.2 lbs.
SUGGESTED PRICE: $629.00

VFO-520

Provides high stability with precision gearing. Function switch provides any combination with the TS-520. Both are equipped with VFO indicators showing at a glance which VFO is being used. Connects with a single cable and obtains its power from the TS-520. Suggested price: $115.00.

SP-520

Although the TS-520 has a built-in speaker, the addition of the SP-520 provides improved tonal quality. A perfect match in both design and performance. Suggested price: $22.95.

TV-502

TRANSMITTING/RECEIVING FREQUENCY: 144-145.7 MHz, 145.0-146.0 MHz (option).
INPUT/OUTPUT IF FREQUENCY: 28.0-29.7 MHz
TYPE OF EMISSION: SSB (A3J), CW (A1)
RATED OUTPUT: 8W (AC operation)
ANTENNA INPUT/OUTPUT IMPEDANCE: 50Ω
UNWANTED RADIATION: Less than −60 dB
RECEIVING SENSITIVITY: More than 1μV at S/N 10 dB
IMAGE RATIO: More than 60 dB
IF REJECTION: More than 60 dB
FREQUENCY STABILITY: Less than ±2.5 kHz during 1-60 min after power switch is ON and within 150 Hz (per 30 min) thereafter.
POWER CONSUMPTION: AC 220/120V, Transmission 50W max., Reception 12W max. DC 13.8V, Transmission 2A max., Reception 0.4A max.
POWER REQUIREMENT: AC 220/120V, DC 12-16V (standard voltage 13.8V)
SEMI-CONDUCTOR: FET 5, Transistor 15, Diode 10.
DIMENSIONS: 6⅝" W x 6" H x 13¼" D
WEIGHT: 11.5 lbs.
SUGGESTED PRICE: $249.00

CW-520
500 Hz CW Crystal Filter: $45.00.

Prices subject to change without notice

TRIO-KENWOOD COMMUNICATIONS INC

KENWOOD GIVES YOU A CHOICE FOR 2-METER SSB

KENWOOD'S TV-502 TRANSVERTER PUTS YOUR TS-520 OR TS-820 ON THE 2-METER BAND...SSB AND CW. SIMPLY PLUG IT IN AND YOU'RE ON THE AIR

OR GO ALL THE WAY WITH THE BEST...THE TS-700A

Ever tried 2 meter SSB or CW? How about the OSCAR satellite? Tune the band with a VFO instead of fixed channel crystals and experience DX-ing on VHF. In fact, there's a VHF QSO party coming up on September 11 thru 13. FMers improve your scores...beginners try it for the first time. You don't need a big antenna to do it either...anything from a coat hanger to - - - ? The OSCAR satellites (6 & 7) are waiting for you too! Or go exotic with meteor scatter or tropospheric ducting. The "Sky is the limit" on VHF SSB and CW.

116 EAST ALONDRA/GARDENA, CA 90248

KENWOOD

...pacesetter in amateur radio

high-performance two-meter fm exciter

Eleven channels, superior modulation and complete kit availability are just a few features of this little jewel

After reviewing the limited number of two-meter fm transmitter construction articles available to the home-brew enthusiast, I decided it was time to break away from the Sonobuoy-type design and try to generate some interest in building a more conventional commercial-type exciter. This article is the result of the overwhelming response to an earlier construction article for an fm receiver[1] of the type of design I am encouraging.

Before dismissing the Sonobuoy-type exciter completely, I'd like to mention that these designs, which have appeared in the amateur literature for the past few years, deserve a great deal of credit. They were easy to build and were the first solid-state transmitters to gain wide popularity and to be constructed in quantity. However, they had some disadvantages that I've attempted to correct:

1. The Sonobuoy designs used direct frequency modulation of the oscillator with a varicap diode. Although sometimes described as a feature such modulation often resulted in unsymmetrical modulation because of improper dc biasing. For some reason, this flaw has been perpetuated in several spinoffs of the Sonobuoy transmitter that I've seen.

2. Direct fm circuits made crystal switching difficult. Such circuits could not be used with a frequency synthesizer.

3. Audio circuits weren't really optimized for voice operation with a variety of microphones; not surprising since Sonobuoys were designed for a different purpose.

4. Tuned circuits were unshielded and construction, in general, was intended to be of the "disposable" type

By Jerry Vogt, WA2GCF, 182 Belmont Road, Rochester, New York 14162

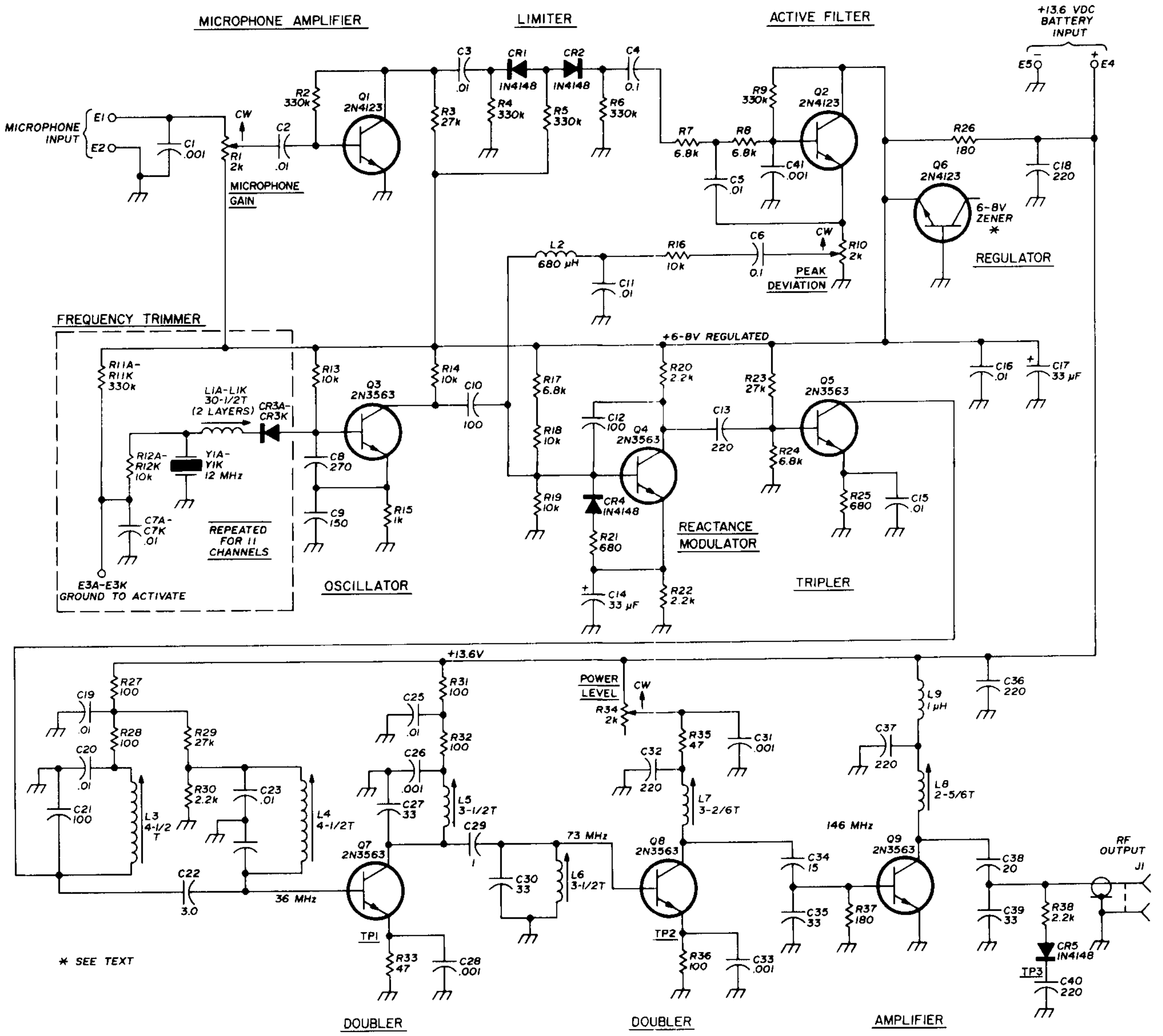


fig. 1. Schematic of the high-performance 2-meter fm exciter. A complete parts kit is available together with other accessories at modest cost.

because of the original intended design function. You can readily see the difference when such a unit sits next to a commercially designed two-way radio.

design goals

After sorting through the many available circuits for a two-meter fm transmitter, design objectives were based on the best features of some of these circuits that could be implemented with readily available parts and simple construction.* The design goals were:

> Here's an example of *ham radio's* promise to bring you construction articles based on available kits so that parts scrounging will be a thing of the past. This two-meter fm exciter has been designed with the homebrewer in mind — easy to build and easy to align and get working. You'll find many features that are included in the latest commercial designs. Best of all, the parts kit price won't fracture your pocketbook. **Editor**

1. Superior modulation — quality to be proud of in either a repeater or home station.

2. Inclusion of both deviation and microphone gain controls.

3. Effective lowpass filter following the modulation limiter to eliminate raspy voice signals.

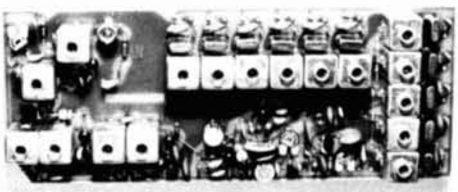
4. Compatibility with either carbon or transistor-amplified dynamic microphones.

*A list of kits and accessories for the transmitter will be found at the end of this article.

5. Phase modulation suitable for use with multichannel operation and frequency synthesizers.

6. Shielded coils and sufficient tuned circuits between multiplier stages to reject harmonics and spurious signals.

7. A maximum number of electronically switched and individually adjustable channels, consistent with the selected board size along with exciter and multiplier stages.



Top view of the completed 2-meter fm exciter showing clean layout of parts without overcrowding.

8. An oscillator using common 12-MHz series-resonant crystals.

9. Voltage regulator for oscillator, modulator, and audio stages to minimize effects of line-voltage variation and noise.

10. Adjustable output level.

11. Sufficient output power (150-200 milliwatts) to drive the new rf power modules now available.

12. A design easy to build, align, and test.

features and technical characteristics

The exciter measures 3 x 7½ x 1 inches (7.6 x 19 x 2.5cm) and weighs 5 ounces (142gm). Operating power is +13 volts at 70 mA. Sufficient power is provided for use as a control link, or the exciter can be used as a low-power transmitter. It can be adapted easily to 50 or 220 MHz by changing the multiplier tuned circuits, or it can be adapted to 450 MHz by adding tripler/driver stages on a separate board.

Eleven oscillator channels are used. Multilayer oscillator frequency-netting coils provide vernier adjustment on each channel. A simple ground-*on* diode channel-switching scheme allows easy adaptation to trunk mounting (important in these days of ripoffs) or other remote-control application. CW operation is possible by keying the B+ line to the multiplier and output-amplifier stages.

The exciter can also be used as an inexpensive multichannel frequency standard or fm signal generator. A convenient output-level control, used with an external fixed attenuator, provides variable output level.

The exciter schematic is shown in **fig. 1**. Audio signals from a carbon or transistor-amplified dynamic microphone are applied to microphone amplifier Q1 through microphone gain control R1. This control, normally not provided in most transmitters, allows the audio level to the limiter stage to be adjusted independently of deviation level, so that sufficient audio punch is obtained without excessive clipping. Limiter CR1, CR2 consists of back-to-back diodes forward biased a small amount. When audio peaks exceeding the bias level are applied, clipping action limits the audio transferred to the next stage. Q2 is a lowpass filter with a cutoff frequency just above the normal voice range to remove the audio harmonic components generated by clipping the audio signal. This stage, together with its separate microphone gain and deviation controls, is primarily responsible for the professional-sounding modulation of this exciter.

The 12-MHz injection to the modulator is provided by Clapp oscillator Q3. Eleven channels are diode switched by grounding the appropriate control lines. The diodes are reverse biased except the one having its control line activated to cause dc conduction, thereby completing the path to one crystal circuit from the base of Q3. The variable coil allows the crystal load reactance to be varied for frequency netting.

The phase-modulated 12-MHz signal from reactance-modulator stage Q4 is multiplied in tripler stage Q5, doubler Q7, and doubler Q8. Double-tuned circuits between multiplier stages provide rejection of spurious frequencies. (Multipliers create more than one harmonic, of course; and if tuned circuits of sufficiently high Q are not used, undesired frequencies will be passed through the multiplier chains to cause spurs in the transmitter output).

The B+ voltage to doubler Q8 is adjustable with series potentiometer R34 to provide output power level control. This control is normally set fully clockwise but may be used to reduce output level if desired. (For example, to limit drive to a power amplifier or provide variable output when used as an alignment generator). Amplifier Q9 provides 150-200 mW output to a 50-ohm load (2-3 volts). This level was chosen to drive an rf power module and is also suitable for several other applications. If desired, a simple one-watt PA stage could be added as well as many other types of amplifiers, although the rf power module is simplest by far.

To allow for easy alignment with only a vtvm, three test points provide dc signals as a function of rf levels at several stages. TP1 and TP2 provide indications of emitter current in the two doubler stages. TP3, in conjunction with rf detector CR5, C40, provides an indication of rf output to the antenna or power amplifier.

power amplifiers

There is only one word to describe the new rf power modules by TRW and Motorola: "fantastic." You have probably seen them in advertisements for various radios.

The photo of a power amplifier using one of the rf power modules shows how simple it is to make a PA of moderate power level today. The exciter in this article was designed to drive these power-module PAs.

The rf power modules, or "bricks" as they're sometimes called, are magic compared to the alternative. Each PA brick is an integrated circuit containing several power amplifier stages with decoupling and tuned circuits to provide many watts output for 150-200 mW input. All that's required externally is a means of connection (a PC board) and low-frequency decoupling components.

Operated at normal 13.6 volts from a battery supply, the PA bricks provide 20 watts of rf at 145 MHz, or 13 watts at 432-450 MHz. (A tripler/driver module kit, model T-20, is available for 450-MHz operation). At this power level, no damage will occur even when operating into a load of infinite vswr. Usually, the bricks can be driven to 25 watts at 2 meters or 15 watts at 450 MHz or higher if loaded properly and operated with sufficient drive and B+ supply. Of course, the greatest feature is that absolutely no tuning is ever required! Current requirement is 2-4 amperes, depending on rf level. Efficiency is 30 to 50 per cent.

No doubt it's more expensive, watt-for-watt, to use an rf power module instead of an amplifier with discrete components. However, considering the difficulties you can have with discrete PAs and the careful tuning and special packaging required, the brick is a bargain and certainly adds a lot of pleasure to homebrewing transmitters.

construction

The exciter is assembled on a single-sided 3 x 7½ inch (7.6x19cm) PC board. Construction details are shown in **fig. 2.** The following details of coil assembly and other suggestions are given to facilitate assembly.

Plastic coil forms of ¼ inch (6mm) OD are used with ½ inch (13mm) square shields and vhf tuning slugs. The coils (**fig. 1**) are wound in a clockwise direction as viewed from the top, using the solderable wire supplied with the kit. *All turns are close spaced.* The diagram of **fig. 3** is exaggerated for clarity. However, all leads should be pulled tight, no fancy bends are required, and no Q dope is needed. Holes in the base of the form are numbered as shown. Fractions of turns relate to the number of funnels on the form; for example, 2/6 turns beyond a full number (as in L7, **fig. 1**) means ending two funnels *beyond* the starting funnel.

Oscillator trimmer coils are wound with no. 26 (0.3mm) wire in two layers, with just over half the turns on the first layer and the remaining turns on the second layer. Wind from bottom to top, going directly back to the bottom and winding up again to the top on the second layer. All other coils are wound in one layer with no. 24 (0.5mm) wire. Coils with capacitors will have capacitor leads inserted through coil-form funnels after the coils are wound. The capacitors should be seated down onto the board as much as possible, as should all parts. This is important at vhf!

Don't be over concerned with coil winding. Neatness isn't important; turns can overlap and windings need not

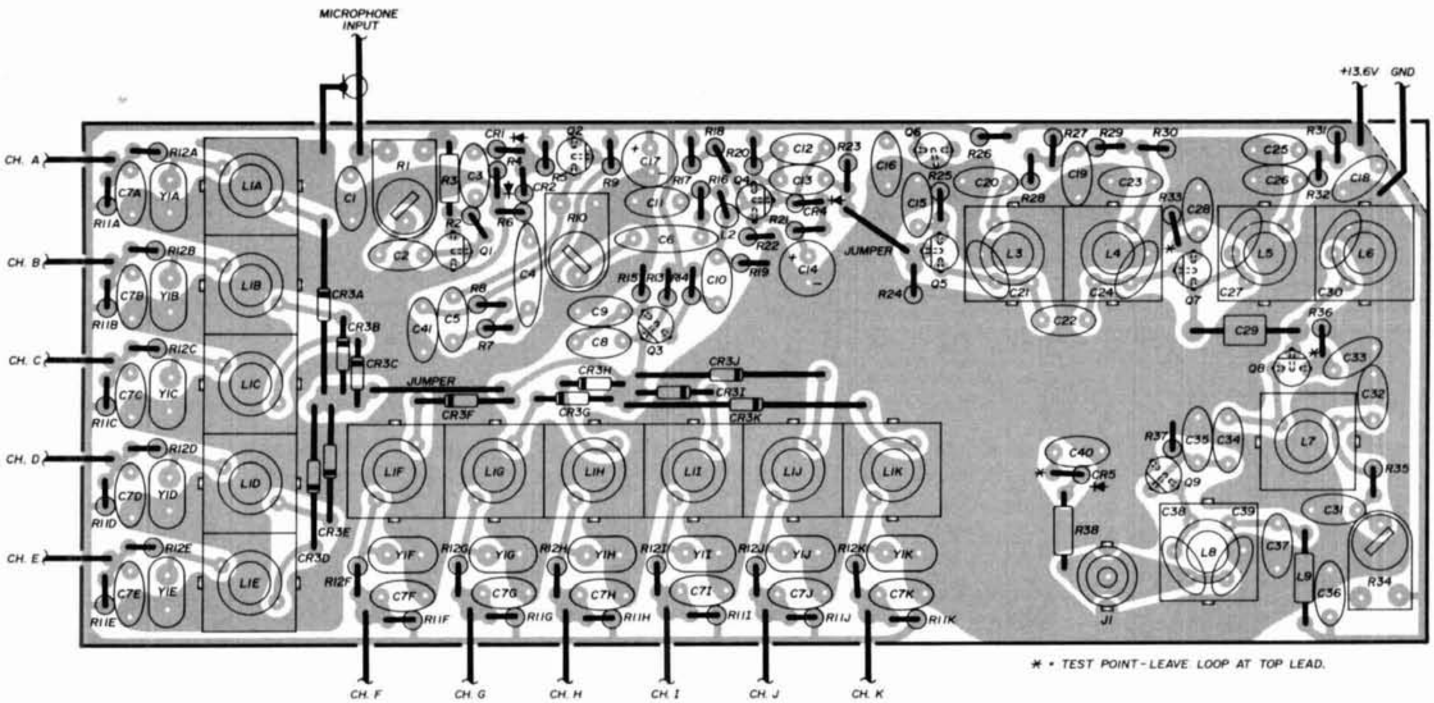
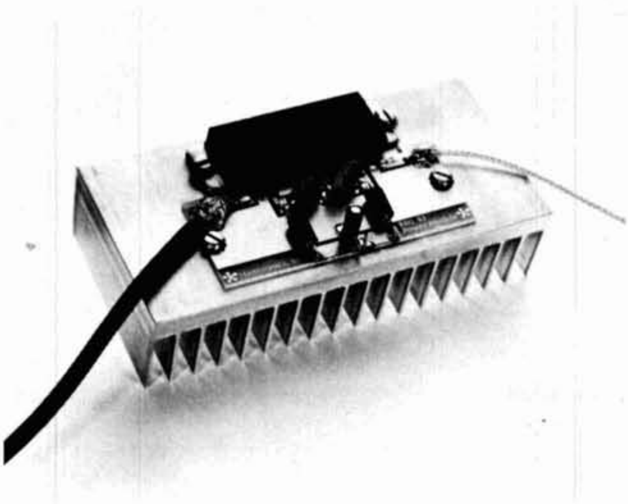


fig. 2. PC-board component layout for the 2-meter fm exciter. Eleven channels are included plus microphone gain and deviation-level controls all on a 3 x 7½ inch (7.6x19cm) printed-circuit board.

be uniform. Start coil leads through board holes while the coil form is above the board, then seat the coil into place onto the board. Don't attempt to insert capacitor leads with the coil form tight against board. After coils are installed, apply heat from a very hot soldering iron



The 2-meter fm exciter married to one of the "instant PAs" for increased power output.

for 10-15 seconds with solder applied. This will automatically strip the wire and allow solder bonding to occur. If you prefer, the leads may be stripped in the conventional way before installation in the board. Don't solder-strip the leads unless the coil is mounted on the board, or the leads will migrate in the warm plastic.

Shield-can tabs should be soldered to the board for proper grounding and mechanical support. It is unnecessary to bend the tabs. Diodes and electrolytic capacitors must be installed with proper polarity. The top leads of R33, R36, and CR5 should be installed with loops accessable for probe connection. The extra length allows the top of the component to serve as a test point. Crystal sockets may be used with the popular HC-25/U type crystal, or crystals may be soldered directly to the board.

External leads should be connected by soldering them into large pads on the PC board. The microphone shield may be easily terminated by wrapping a few turns of bus wire around the shield and soldering. The channel switch should ground the desired channel line for activation. It's sometimes possible to use a common switch for both transmit- and receive-channel selection if the receive switching scheme also uses a ground.

mounting

The T40 PC board is designed to slide into a vertical groove in a companion cabinet just forward of the rf power module mounted on the rear panel. The components should face forward, and the board should be oriented upward as shown in the component location diagram. The upper right-hand corner may be cut off as indicated in **fig. 2** to allow cables to be routed past the board in the cabinet. If the companion cabinet is not used, appropriate holes may be drilled in ground areas of the PC board to allow standoff mounting to your own chassis or panel.

crystals

The exciter uses HC-25/U series-resonant crystals. The crystal frequency, in the 12-MHz region, is determined by dividing the 2-meter channel frequency by 12. When adapted for other bands, the divisor changes accordingly; e.g., 36 for 450 MHz, 18 for 220 MHz, and 4 for 52 MHz.

alignment

After constructing and visually checking the PC board for proper assembly and soldering, you are ready to apply power and perform alignment and testing. Caution: Use a proper tuning tool; a loosely fitting tool may crack the powdered-iron tuning slugs.

1. Install one crystal at the approximate center of the desired frequency range, and ground the corresponding channel control line.

2. Set the audio controls R1 and R10 to full counter clockwise, and set power control R34 fully clockwise.

3. Preset all tuning slugs to half range.

4. Connect 50-ohm load to output connector J1.

5. Apply 13.6 Vdc (battery power, etc.). *Observe polarity.*

6. Check regulated voltage. It should be approximately +6 to +9 Vdc.

7. Connect vtvm, set to 0.5 Vdc range, to TP1 (loop at top of R33 at first doubler Q7).

8. Peak L3; then peak L4. Dip L5. Dc voltage should be roughly +0.3 to +0.4 volt.

9. Connect vtvm, set to 1.5 Vdc range, to TP2 (top of R36 at second doubler Q8).

10. Peak L6; then peak L5. Dip L7. Dc voltage should be roughly +1 to +2 volts.

11. Connect vtvm, set to 1.5 Vdc range, to TP3 (top of CR5 at final amplifier Q9).

12. Peak L8; then peak L7. Dc voltage should be roughly +1 to +2 volts.

13. Repeat all above steps to eliminate effects of interaction and to check tuning. When you're finished, the dc current drain should be about 70 mA. If you have an accurate rf probe arrangement, the rf voltage at the 50-ohm load should be about 2-3 volts (150-200mW), which is the level required to drive the rf power module. Note that output will be somewhat less if the power supply voltage is below 13.6 Vdc. *Note*: Do not attempt to tune in any manner other than described. In particular, multiplier stages should *not* be repeaked for maximum at the antenna connector.

14. One, by one, ground each channel control line with crystals installed, and adjust the corresponding oscillator trimmer coil to net each channel to the proper frequency.

troubleshooting

The usual troubleshooting techniques of checking dc voltages at transistor elements and tracing ac signals, with a voltmeter and an rf probe where applicable, apply in this case. Don't overlook the possibility that parts may be installed incorrectly.

For convenience, the regulated voltage is obtained for the low-level stages by using a 2N4123 as a zener diode (Q6 in **fig. 1**). Since transistors are not calibrated for this parameter, the zener voltage should be checked the first time the board is fired up to ensure that a zener voltage in the range of 6 to 9 Vdc occurs. If lower, you may wish to substitute another 2N4123 to find one with a useful zener voltage. The exact voltage is unimportant; it is only necessary that the voltage be held stable under varying line conditions.

If trouble is encountered in netting one or more channels, check the number of turns on the corresponding oscillator trimmer coils. Make sure the coil is wound in two layers, as described earlier. If a channel won't oscillate at all, check the corresponding diode and other components in the control-line circuit. The following typical test voltages will serve as a *rough guide* to proper transistor operation, based on 13.6 Vdc input.

transistor	emitter	base	collector
Q1	0	0.6	2
Q2	2.2	2.9	7*
Q3	3.8	4.4	6*
Q4	2.6	3.2	4.5*
Q5	1.2	1.8	13.4
Q6	7*	0	-------
Q7	0.35†	0.6†	13.4
Q8	1.7†	1.7†	13.6

*Assumes 7 volts from a regulated supply, but supply ranges between 6-9 V in actual units.

†Rough indication of drive level.

Base and collector voltages of Q9 cannot be measured with drive applied because of rf effects on meter.

microphone and audio adjustments

The exciter is designed to operate with either a carbon microphone or a transistorized dynamic microphone. The microphone should be connected with shielded cable to avoid rf pickup. To adjust deviation level, preset R1 and R10 both fully clockwise. Key the exciter, and make sure that the carrier is adjusted properly to frequency. Speak into the microphone and observe the deviation meter on the receiver, or listen to the audio with the squelch set tight. Reduce deviation control R10 setting until all noticeable effects of overdeviation are removed; e.g., distortion, meter swing on peaks, squelch pumping.

The setting of microphone gain control R1 is a refinement not found on most transmitters. It should be set to provide sufficient audio for full modulation on voice peaks but low enough to remove background noise and obvious clipping effects, which normally result from overdriving a clipper.

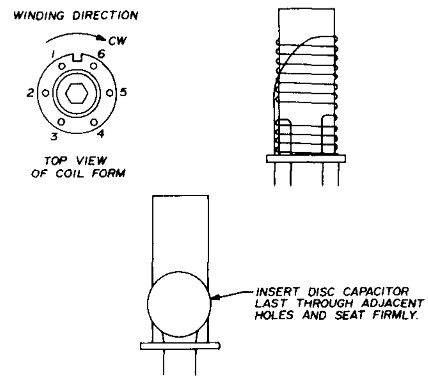


fig. 3 Diagram for winding coils (exaggerated for clarity). The number of turns for all coils is shown in the schematic. All turns are close spaced; winding direction is extremely important.

Since the amount of deviation depends on the frequency band to which the exciter signal is multiplied, resistor R16 is provided to allow the amount of deviation at full control setting to be altered. The value of R16 may be reduced to obtain wider deviation if required, and it may be increased to reduce the range of the deviation control. (R16 is also a part of the lowpass filter following the limiter).

The following kits and accessories are being made available in conjunction with this article.

part number	description	price
T40	Two-meter fm exciter module kit	$39.95
T80-150	Two-meter rf power module PA	$79.95
T80-450	432-450 MHz rf power module PA	$79.95
	note Rf power module PAs are wired and tested and are complete with heat sink. No construction or tuning is required, ever.	
T20	Two-meter to 450 MHz tripler/driver module kit	$19.95
A25	Cycolac cabinet, 7 x 7¾ x 3½ inches (17.8x19.7x8.9cm) with aluminum panels, to house exciter, rf power module PA, and other modules	$24.95
	Crystals for any desired channel frequencies	$ 5.50

When ordering, please add $1.00 for UPS or parcel post shipping. New York State residents, please add 4% sales tax. Other kits offered include fm receivers, converters, preamps, scanner and multifrequency adapters. Send a self-addressed, stamped envelope for a complete catalog to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

reference

1. G. Francis Vogt, WA2GCF, "High Performance VHF FM Receiver," *ham radio*, November, 1975, page 8.

ham radio

update of the phase-locked loop RTTY demodulator

Here are the answers to your questions about this terminal unit plus a modified circuit for upward shift

Since the NS-1 phase-locked loop RTTY demodulator first appeared in the *RTTY Journal,* October, 1974, and later in *ham radio,*[1] I have received numerous inquiries about its operation. I hope this article will answer some of the questions and also provide some added tips.

First, as an explanation, the NS-1 was developed primarily for fsk downward shift on the high-frequency bands, using low tones to take advantage of a narrow receiver passband. The 741 op-amp limiter drops off around 2000 Hz, so the tones must be within this limit. On fsk the tones can be varied by receiver tuning, so it's easy to obtain these low tones. Shift reversal is accomplished by changing sidebands if receiving in the ssb mode or by moving the bfo to the other side of zero beat if receiving in the CW mode.

meter adjustment

I suppose the most-often asked question concerned the value of the zero-center, meter-adjustment pot. This value depends on the meter movement; i.e., whether it's 50 μA, 200 μa, 1 mA, etc. The easiest way to determine this value is to put a 500k pot in series with the meter and gradually reduce the resistance until full scale (+) is obtained on the mark/hold signal. This resistance can then be measured and a fixed resistor substituted.

If you don't have a zero-center meter, **fig. 1** shows such a circuit using a regular 0-1 mA meter. Before connecting the 18k resistor to the *M* terminal, set the 25k pot about midway, ground the 18k resistor to the chassis, and adjust the pot until the meter reads center scale (0.5mA). Now, when the 18k resistor is connected to the *M* terminal, the meter will act like a zero-center meter with plus to the right and minus to the left.

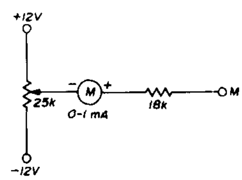


fig. 1. Using a 0-1 mA meter for zero-center indication. Connected as shown, the meter will indicate plus to the right and minus to the left. See text for adjustment instructions.

cross pattern

Several readers asked where to connect a scope to receive a cross pattern. A scope cross pattern requires tuned filters to distinguish between the mark and space tones, one displayed vertically and the other horizontally. Since the NS-1 has no filters, there is no place to connect the scope. However, if you have a scope with tuned filters that *will* produce a cross pattern, connect the scope ahead of the NS-1 or at the receiver output. Tune in a signal that gives a good cross pattern, then adjust R1, the 5k vco pot, until you get good copy. Thereafter you can use the cross pattern for tuning.

Some readers complained that they were unable to

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get *wide-shift copy*. The usual 2125/2975 Hz tones will not work, as the 741 limiter will not accept frequencies much above 2000 Hz, as stated previously. So tones, say, around 2125/1275 Hz should be used. A compromise setting of R1 can be found that will permit copy of both wide and narrow shifts.

other tips

The purpose of the switch is to put the teletype machine in a "hold" condition. Random noise and signals will produce garble when tuning. Also the switch should be closed when transmitting, as interaction between the receiver and TU will also produce garble.

Remember that the NS-1 will act very much like any other fm demodulator. If a stronger signal is near the one you're trying to copy, the stronger signal will take over. I've found that a receiver T-notch filter is quite effective at times. Also, use the narrowest selectivity you have on your receiver that will permit copy of the tones; this reduces the effects of interference. One amateur wrote that he installed an active audio filter (2100-2300 Hz) ahead of the NS-1 and adjusted the unit for 2295/2125 Hz. He said this arrangement compared very favorably with the ST-type terminal units.

modified unit for upward shift

A few people have tried to use the NS-1 on afsk on the six- and two-meter bands and found it would not work. This is because afsk is usually upward shift (mark low, space high), and the NS-1 will not copy this way. Afsk tones are fixed. Nothing can be done at the receiving end to reverse the shift, and there are no provisions for it on the TU.

To achieve copy with upward shift, it's necessary to reverse the polarity of the voltage coming out of the 741 comparator on the mark/hold signal. Several circuit changes were tried including reversal of the two inputs to the 741. This change worked, but best results were obtained after installing a transistor just ahead of the 2N5655. A 2N706 was used. It switches the keying transistor off and on and has the effect of reversing the voltage from the 741 output, which permits smooth upward-shift copy.

The 741 limiter was eliminated since it is restricted to around 2000 Hz; thus high tones such as 2125/2975 Hz can be used. Two reversed diodes were placed ahead of the 565 PLL. These give good limiting and prevent PLL overload. The 565 is very sensitive, so no amplifier is needed. This modified arrangement works equally well on afsk and fsk, wide or narrow shift. The revised circuit diagram is shown in **fig. 2** and has been designated the NS-1A.* The same adjustment procedure as used on the NS-1 should be used for the NS-1A. The zero-center tuning meter will now show full scale minus (left) reading on the mark/hold signal. The tuning meter may not even be needed on afsk.

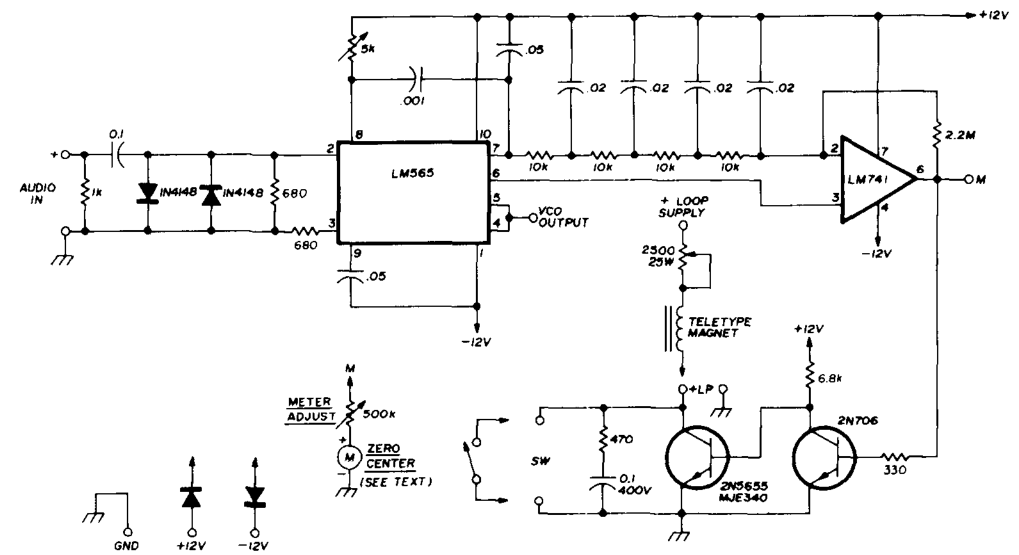

fig. 2. Schematic of the NS-1A phase-locked loop demodulator for copying afsk/fsk (upward shift). The 2N706 switches the 2N5655 on and off, which reverses the polarity of the voltage from the LM741 comparator on mark/hold. This permits smooth upward-shift copy.

acknowledgements

I wish to thank Ron, W8BBB; Buck, WAØLEM; and Rick, WB5FHU for their tests and evaluation of the NS-1A on afsk. The added feature of afsk copy should encourage more activity on two- and six-meter RTTY with this simple terminal unit.

*At present no modified boards are available; however, wired and tested NS-1A units are available for $29.95 postpaid; the circuit board is $4.75 postpaid. For more information, contact the author.

reference

1. Nat Stinnette, W4AYV, "Phase-Locked Loop RTTY Terminal Unit," *ham radio*, February, 1975, page 36.

ham radio

the hand-held electronic calculator: its function and use

First of a four-part series on how to apply the simple four-function machine to math problems encountered in radio work

This is the first of four articles on the hand-held electronic calculator — how it works and how it can work for you to solve even the most complex problems encountered in radio work. The operating principles of the four-function machine are discussed first, with examples illustrating how the basic arithmetic functions are performed in response to input-key manipulations. Discussed next are some of the more important operations that can be performed with larger and more expensive machines: chain operation, the use of constants, and the use of various memories. Finally, some suggestions are given to help you choose the best machine to fit your needs.

In the second article, which will be published in the next issue of *ham radio,* the power of the simple four-function calculator is expanded for solving problems in radio work. Examples are given on using approximations as substitutes for special calculator functions, the use of problem-organization technique to overcome the limitations of simple machines, and the use of the "scratch pad" as a substitute for calculator memory. Examples are given on using the simple four-function machine to solve problems involving transcendental functions and numerical integration. The following articles will cover transmission-line calculations and use of the new programmable calculators.

These articles should help dispel the mystery of how to use the four-function electronic calculator in solving electronics problems by providing basic information on the logic of operations within the machine and providing the approximations and tools available.

basic calculator principles

Addition. In digital electronics, a *register* is a place to *store* numbers. Actually, these don't have to be electronic devices; a chain of relays, a toothed wheel, or even knots in a piece of string can serve as a register. But in the small calculator registers are always electronic and are almost always made using field-effect transistors.

To be useful there must be a way of putting numbers into the register, of taking them out, and, for at least some registers, of telling what number is stored. Hand-held calculators have standardized on a set of push-buttons or *keys* for *input*, and a stylized *display* for presenting the content or number stored.

Suppose we wish to perform the simple arithmetic operation of addition. We could use three registers, one for the *addend,* or first number, a second for the *augend,* or second number, and a third for the result, or *sum.* Actually it's easier to do this with only two registers by a technique called "add to storage." In this, the addend, when received, is first placed in the sum register, then the augend is added to it to get the final sum. The two

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registers involved can be called the keyboard register and the answer register.

To make a small calculator, we need the two registers, the keyboard, and the display. We also need some control circuitry to tell the register how to work. For convenience, the control can also be arranged to switch the display from the keyboard to the answer register. In block diagram form, these elements are shown in **fig. 1.**

We can see these registers in operation on any calculator by considering only the ten number input keys 0 through 9, and the *add* key. This key is usually marked +/= in small calculators, but may be marked +. When the calculator is first turned on, a *0* will appear at the extreme right of the display (a decimal point may appear also, but ignore this for now). Suppose we wish the addend to be the number *123*. Pressing the *1* key causes the *0* to change to a *1*. Pressing the *2* key causes the *1* to jump, or shift, to the second position from the right and a *2* to appear at the rightmost position. Similarly, pressing the *3* key causes the numbers displayed to shift again, the display now reading *123*. We have now *entered* the desired number, *123*.

To transfer this number to the answer register, we now press the add key (+/= or +). All that seems to happen is a blink of the display, but the display is now presenting the number in the answer register. We can show this as follows. Press the *0* key, which causes the display to change to a zero. Now press the add key again: the display changes to *123*, the number stored in the answer register. This display will continue to show the number in the answer register until another operation is started.

In this test we are actually performing an addition, *123 + 0 = 123*. We can continue addition by entering another number then pressing the add key. Each time, the display switches to the keyboard register then back to the answer register as the add key is pressed. The answer register always shows the sum of all numbers entered.

When pressing the number keys it's possible to make a mistake, say by pressing *132* instead of the desired *123*. This is easily seen, since the display shows the number actually entered; this is the reason for switching the display back and forth.

To correct a mistake in entry, it's necessary to wipe out the entire number entered, or *clear* the keyboard register, and start over. On most small calculators this is done by pressing a special control key marked *CE* for *Clear Entry*. In this example, pressing *CE* causes the display (and the keyboard register) to change from *132* to *0*. The correct number is then entered.

Let's leave addition and go on to other operations. To get ready for these, the number stored in the answer register must be removed lest it make an error in the next calculation. To do this, another control key is pressed, marked *C* for *Clear*, which sets the internal register to zero. The same result could be obtained by switching off the calculator; the internal circuits in most calculators are arranged to clear the registers when the power is first switched on.

In some calculators *Clear Entry* and *Clear* are combined into a single key, usually marked *C* or *CLR*. Pressing this key once clears the entry, and pressing it twice in succession clears all registers.

Subtraction. As far as the user is concerned, the operation of subtraction is almost identical to that of addition, the one difference being that a different operation key, marked – or –/= is used. Suppose we wish to subtract *456* from *123*. The number *123* is first entered into the keyboard register then transferred to the answer register by pressing the + key. The number *456* is entered into the keyboard register then transferred to the answer register with the proper sign by pressing = (equals). The answer, *–333* appears immediately. In some calculators the minus sign appears next to the left number; in others it's at the extreme left. The number is called a *"signed number"* to indicate that the sign relates to the number rather than to the operation (subtract).

When working with signed numbers the order of entering the numbers isn't important providing the signs associated with the numbers are entered properly. For example, *123 – 456* may be handled as follows:

enter	press	display
	C	*0*
	+	*0*
123	–	*123*
456	=	*–333*

Alternatively, the operation may be:

enter	press	display
	C	*0*
	–	*0*
456	+	*–456*
123	=	*–333*

Multiplication. Recall that multiplication is just a method of successive addition. Suppose we wish to multiply *456* by *123*. The answer is equal to *456 + 456 + 456*, plus *4560 + 4560* plus *45600*. This can be confirmed by making the addition, the sum being *56088*.

To solve this problem in a calculator, the internal controls cause the multiplicand to be added to itself three times, and the result is stored in the answer register. The multiplicand is then multiplied by *10*, and this value is added to the answer register twice. Finally, the multiplicand is again multiplied by ten, and this value is added to the answer register to get the result, *56088*. This process requires a place to store the multiplicand, a third register, as shown in block form in **fig. 2.** In most small calculators the contents of this register can't be displayed so it's often called a *hidden register.*

To the user the operations for multiplication are not much different than for addition. The first number, the *multiplicand* is first entered, followed by x. This operation stores the number in the hidden register. The second number, the *multiplier,* is then entered. Pressing the = or +/= key gives the answer, the *product*.

In very small calculators, it's not possible to multiply negative numbers directly. The easiest way to handle this is to remember that, from algebra, the product is positive if *both multiplier and multiplicand have the same sign*; if not, the product is negative.

Overflow. Recall that the number of digits in the product is either equal to the sum of digits of the multiplier and multiplicand, or is one less than the sum. When two large numbers are multiplied, the product may have more digits than the register (and display) capacity. This condition is called *overflow*. It can also occur in addition, since the number of digits in the sum can be one greater than the largest number of digits of the two numbers being added.

Most calculators have some form of indication of overflow. It may be a special symbol at the left of the display, a glowing dot, or a blink. Some small calculators eliminate the need for this by making the answer register twice the size of the keyboard register: pressing a special key marked with a right-pointing arrow causes the additional digits to be displayed. Also overflow can be avoided by a technique called *multiple precision*, described in the next article.

Division: underflow. Just as the calculator handles multiplication by successive addition, it handles division by successive subtraction. For example, to divide *56088* by *123*, the calculator first subtracts *123* from the left 3 digits, *560*, leaving *437*. This is a positive number, so *123* is subtracted again, to get *314*, then again to get *191*, and stil again to get *68*. Another subtraction of *123* causes the sign to change to minus, producing a reading of *−55*, a signal that subtraction has proceeded too far. Accordingly, *123* is added, to get *68* again. This number is shifted one position to the left, bringing in the next digit, for an internal reading of *688*. The net number of subtractions, 4, is recorded as the first digit of the *quotient*. The process of subtract, test, adjust when necessary, then shift continues until the answer is obtained, *456* in this case.

Suppose we wish to divide a very small number by a larger one, say, *3* divided by *987654*. On the small calculators the number appearing as an answer is a zero. This answer is obviously wrong, since the actual answer must be greater than zero. This condition is called *underflow*. It is handled in one of several ways, depending on calculator design. In some, an underflow signal is presented; in others, a special conversion may be used as described below.

Decimals and calculator notation. So far we have ignored the decimal point and have looked at whole number problems. A few types of calculators do the same thing. It's up to the user to keep track of the decimal. This is not difficult if three simple rules are remembered:

1. In addition and subtraction, zeros must be added to the right as needed to make the number of places to the right of the decimal the same for all numbers. Example: 7.6 + 0.25 must be entered as 760 + 25. The indicated answer, 785, is read as 7.85.

2. In multiplication, the number of places to the right of the decimal in the product is equal to the sum of the number of places of the multiplier and multiplicand.

3. In division, a simple method is to use zeros to give the same number of places in divisor and dividend. The answer displayed is the part of the quotient to the right of the decimal point. On most no-decimal calculators, the decimal part of the quotient can then be displayed by pressing a key marked →.

By far the largest number of calculators on the market have decimal provision, indicated by a key labeled with a decimal dot. There are, however, several ways of handling the decimal. Some calculators allow entry of the decimal point at any place up to a limit – often 2, 4, or 5 places. The answer is given to the same number of places (fixed decimal).

Some assume that the number to be entered has two decimal places, as in dollars and cents. Answers show two places (adding machine entry).

Some allow the decimal to occur at any place on the display for both entry and answer. The answer display starts with the first digit, or with a decimal, as necessary (floating decimal).

Some express the answer as a number multiplied by a power of ten, such as 1.2345×10^{-3} (scientific notation).

Some restrict the power of ten to 10^0, 10^3, 10^6 (engineering notation).

In complex machines, there may be manual or automatic change from one notation to the other. The method of handling the decimal is very important, and time should be spent with the instructions and in practice until the mode of operation is thoroughly understood.

simple and extended calculators

So far we've looked at the elements basic to all calculators. As seen with respect to the keyboard, these are:

numericals: *0* to *9*, and . (decimal point)
operations: +, −, x, ÷, =
instructions: C, CE

Depending on the design, these can be placed on 15 to 17 keys, 16 being the most common. In simple calculators only these keys are found. There are however, a number of other instructions and operations that can be added and which will be found in various combinations in the larger and more expensive machines.

Chain calculations. In the simplest calculator the number in the answer register is available only to the display As a result, this register must be cleared at the end of each calculation before starting another to prevent error.

More advanced calculators, by internal switching, make this number available to the keyboard or to the hidden register, to serve as the input for a new series of calculations. This technique is called *chain*, since it allows linking or chaining of successive operations.

Where this method of construction really pays off is in mixed calculations. For example, the problem $\frac{2 \times 3}{4} + 6$ is solved by the successive entries of *2*, x, *3*, ÷, *4*, +/=, *6*, +/=, (calculators with no separate = key), for an answer of *7.5*. Without chain, the entries would be *2*, x, *3*, +/=, C, *6*, ÷, *4*, +/=, C, *1.5*, +, *6*, +/= again for an answer of *7.5* but with the necessity of remembering the result of each step long enough to input it for the next step.

There are some things to be aware of with chain operation. For example if you wish to solve $2^3 = 8$ and you press the keys *2*, x, x, x the display will show 16. In the multiply, divide, and add operation above, omission of the intermediate equal sign gives the erroneous result of *6* instead of *7.5*. More important, the order of operations must be correct. In most simple calculators, multiply/divide can be mixed in any order as can add/subtract; but multiply/divide must precede add/subtract in a chain.

Percent. Many small calculators have a key marked with the symbol for percent, %. This is used exactly like the = key to obtain an answer, but it has the effect of shifting the decimal point two places to the right on divide and two places to the left on multiply For example, *2*,÷, *3*, % gives *66.6666*, and *2*, x, *3*, % gives *0.06*. On divide, the result answers the question, "What percent of the second number is the first?" and on multiply, the question, "What is the second number percentage of the first number?"

Values of constants. When a register is cleared, it is actually set to contain zeros. It is no great problem to design it to be set to some other number.

In the small calculator, such "set to a value" capability is rarely provided. Even the larger ones include only a single value, that of π, or *3.141592654*. An exception is the family of special calculators designed for metric conversion. In these, the multiplying factors for feet to meters, pounds to kilograms, etc., are built in and come into play as the appropriately marked key is pressed.

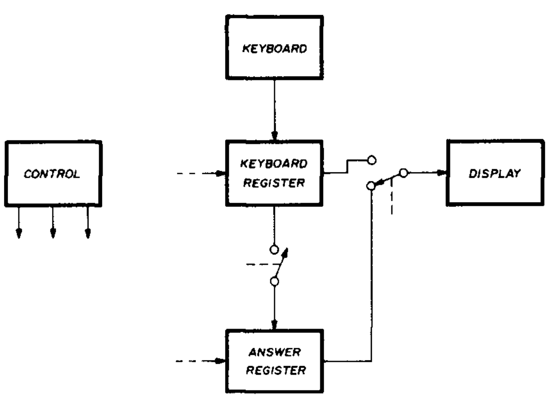

fig. 1. Elements of the basic hand-held electronic calculator. The registers store information input from the keyboard, and the display provides a readout of calculation results based on instructions to the registers from the control circuits.

Change sign. In many small calculators, operations such as division by a negative number are not directly possible, since the − is an instruction to subtract. However, some avoid the problem by providing a special key, usually marked +/−, i.e., *change sign*. This operates only on the keyboard register, causing its sign to change + to − or − to +.

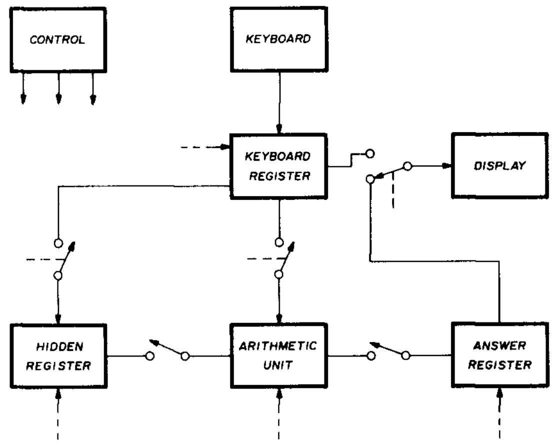

fig. 2. Calculator with add, subtract, divide, and multiply functions. A third register is used to store the multiplicand during problems involving multiplication. Its contents can't be displayed, hence the name "hidden register."

Exchange. In **fig. 2**, the number in the third, or hidden, register can't be seen; it can't be checked; nor is it available except for specifically intended operations. To increase flexibility, a key may be provided that causes the contents of the keyboard and hidden registers to be exchanged. Usually the key is marked *EX*; i.e., *exchange.* Pressing the key once allows the contents of the hidden register to be examined; a second action returns the contents to the initial condition. Alternatively, after the first actuation, an operation can be performed, say, addition.

Memory 1: constant. One function of the registers shown in **figs. 1** and **2** is to remember the numbers placed in them. It seems obvious that calculators would be more flexible in use and would have more calculating power if more memory capability were provided, just as the calculator of **fig. 2** has more capability than the simpler unit of **fig. 1.** In practice this is true.

The simplest type of added memory is called *constant.* When provided, it is usually made evident by a slide switch, one position being marked *K*. When this key is depressed, a number can be loaded into this memory by pressing an arithmetic operation key. Thereafter, entering another number followed by = causes the operation to be performed between two numbers. For example, suppose *constant* is on, and has been loaded by pressing *3.14159*, x. Pressing *120*, =, gives the answer, *376.9908*, or 120π. Now pressing *2*, = gives *6.28318*. The constant is retained until replaced by keying a number and some operation other than =.

Quite a range of variation is possible in the method of operation. Some calculators require two numbers to be loaded and place the second in the constant memory. In others, the constant memory holds the first entry. Another variation is *automatic constant,* which seems to be coupled with the chain. It operates the same as the switchable constant described above but is always available without switching. In other variations, *constant* can be used with multiply and divide but not with add and subtract. However, despite its exact form, this simple form of memory adds greatly to the calculator power.

Memory 2: four-function memory. A further increase in calculation power can be secured by making the memory more flexible so that it can be loaded or recalled at any point in the calculation cycle. Typically, this is done by providing four keys, *+M, -M, RM, CM,* indicating: add number displayed to the number in memory, subtract number displayed from the number in memory, read memory (transfer number in memory to the keyboard register, leaving the number in memory intact), and clear memory (set the memory number to zero). Some calculators include a flag to show when memory is loaded. Some may omit one or more keys, usually the *-M* key.

The major use of this form of memory is to store intermediate results while making a continuing calculation. For example, to solve $(A \cdot B) - (Y \cdot Z)$, the key strokes are A, x, B, =, +M, Y, x, Z, =, -M, RM. Another use is storage for constants, such as π, or the rate of sales tax, for example.

Memory 3: stacks. As problems become more complex, it is found that several numbers must be "remembered" at the same time. One way of accomplishing this is to provide several memory registers arranged so that each successive call for memory input causes a number already in memory to be displaced into another register. These registers are said to form a *stack.*

Stack memory appears in two forms. One is based on conventional algebraic notation and usually appears on the keyboard as a pair of keys marked [(and)], standing for inner and outer parentheses. Successive keying of the left parenthesis causes the stack shift. The other stack arrangement is based on a special notation called Polish reverse notation (no one can pronounce the mathematician's name!).* The two keys for this are usually marked *ENTER↑*, and *RCL* for recall. A key marked *R↓* for *roll* may be provided to allow reading the memories successively. This type has no equals key; the notation arrangement makes it unnecessary Readout of both of these memories is automatic, occurring as the logic of the problem demands.

Memory 4: addressable memory: programs. As the amount of memory provided is increased, the stack arrangement becomes too limiting. To avoid this problem each memory can be assigned to a symbol, so a number can be placed into any memory cell, or the number in any cell can be read out without calculation. This arrangement is called *addressable memory.*

*The Polish logician, Jan Lukasiewicz.

Addressable memory may be found combined with a special memory, one which remembers the successive steps needed to solve a problem. The assembly of steps is called a *program.* In some calculators the program is stored internally and is lost when the power is shut off. Other calculators store the program on a card, usually similar to magnetic tape. The power of these addressable memory and programmable calculators approaches the capability of computers.

Functions. A function of a quantity bears a definable relationship to the quantity. If X represents the quantity, the quantity $Y = X \cdot X$ or $Y = X^2$ is a function of X.

Small calculators have no built-in provisions for these specific functions. Slightly larger ones may provide X^2, $\sqrt{X}$ and $1/X$, by special keys. Still more powerful ones may provide $\sin X$, $\cos X$, $\sin^{-1} X$, e^x, X^y, etc. These additional functions are generated by special circuits in the control section of the calculator. They are calculated each time they are needed, using the number present in the keyboard register. Usually the calculation takes a noticeable time period, during which the display is blank, or flashing.

The next article in this series describes methods of approximating these functions on the smaller calculators.

which calculator?

Which calculator to buy should be based on expected use, problems to be solved, and cost. A simple calculator is best for simple problems; there is much less chance of error. If more complex problems are to be solved or much repetition is involved, the calculator power should be increased accordingly. However, if complex problems are encountered only at rare intervals, it's best to stay with the simpler calculators, using scratch pads and function tables as substitutes for calculator power (discussed in the next article). Not only is the cost lower, but the time wasted is less – it's easy to forget the tricks of the larger calculators.

The four-function, 16-key calculator with constant is about the simplest usable machine for routine radio calculations. The type with four-function memory (usually 20 keys) is appreciably better. Of course, the regular user should consider the advanced calculator. Regardless of the type used, practice is in order. A good technique for refreshing your memory of problems, and at the same time becoming acquainted with the calculator, is to solve the example problems in *The Radio Amateurs Handbook,* or *The Radio Handbook.* In running these problems, it's a good idea to practice approximate mental solution at the same time – the best way to avoid error.

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syllabic vox system

for Drake equipment

Elimination of conventional vox delays is featured in this novel system adapted to the popular Drake T-4XB and R-4B

Radiotelephone conversations without speech delays caused by conventional vox systems is truly a delightful experience. The system described here has been designed for the Drake T-4XB and R-4B transmitter and receiver. It eliminates poor speech habits induced by vox relay delay and also eliminates vox relay contact-bounce transients. In addition, true break-in CW keying is possible with this system.

Dr. Hildreth, WØIP, described his system for instantaneous voice interruption (IVI) in *QST.*[1] While IVI clearly demonstrated the feasibility of rapid-switching syllabic voice response in the control of rf transmission, I experienced difficulties when trying to use conventional vox with IVI.

conventional vox systems

Vox, anti-vox, and delay circuits in most popular transmitters are similar in design principles. Therefore, these techniques might be called "conventional vox." Conventional vox depends on passive integrals developed from the audio signal by rectified dc charges to a control capacitor. Output from the capacitor appears at the grid of a relay-switching tube; or, in solid-state circuits, as current to the base of a relay switching transistor. When a positive charge provides vox activation during transmission, anti-vox during reception requires a negative charge to inhibit operation of the vox relay.

A significant period of time is required for both vox and anti-vox functions because the control capacitor can't be charged or discharged instantaneously. Time, dependent on source and load resistance, together with the value of the control capacitor, is extended accordingly. Likewise, time is equally important to the anti-vox input, which requires similar processing of opposite polarity. A delay of many milliseconds, or even seconds, may be involved, which varies with the amplitudes of vox input and anti-vox receiver output. Usually, these functions are mutually dependent. Such factors contribute to annoying and compromised operating adjustments, often leading to abandonment of vox in favor of PTT.

In the use of conventional vox systems, the operating-point bias to the final amplifier tubes is active throughout the vox relay *on* time. This, of course, means that final-amplifier bias current is present between speech syllables, words, and code elements. Power used in this manner, without rf transmission, is wasted. Also, a T-R switch loses merit, because final-amplifier tube noise appears at the switch and receiver input.

Both these undesirable conditions are eliminated by the syllabic vox method. Final-amplifier operational bias is applied only during rf transmission. At all other times, the final amplifier tubes are cut off completely and no plate current flows. Significant improvement in transmission efficiency results.

adaptation to Drake equipment

Using this system with the Drake R-4B and T-4XB requires no hole drilling or component changes. The only outboard unit required is a small circuit board containing the 5-volt power supply, which is mounted in the Drake MS-4 speaker cabinet. During operation, all controls and transmission modes are in accordance with the Drake instruction manual.

All components except the T-R switch and power supply are mounted on a 3.5 by 4.75-inch (89 by 121 mm) Vector board. The components are self supporting on the board and no heatsinks are required. The circuit board is mounted in the upper right-hand corner of the T-4XB transmitter, where adequate space is available without crowding. Two small L-brackets secure the board to the chassis. Existing chassis screws secure the board. No critical or difficult wiring problems were encountered. Total component cost is about $50.00.

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Referring to **fig. 1**, all logic components are in the 7400-series ICs, U2, U3, and U4. The retriggerable monostable, U3, is the most important single component in this circuit. With the RC values shown, a single positive pulse at U3A pin 3 would deliver a *low* output at U3A pin 6 for 8.4 milliseconds, plus or minus the tolerance of the RC combination. However, the device offers retriggering capability at all audio frequencies between 120 and 4000 Hz in this application, which means that pin 6 of U3A or U3C will remain low during voice syllables. Upon cessation of voice signals, U3A-6 or U3C-6 will go high after 8.4 milliseconds, which is the time required for one CW dot at 140 wpm. Obviously, this cutoff time is insignificant with respect to voice syllable duration.

U1, an LM3900 quad operation amplifier, was chosen for front-end application to both vox and anti-vox channels. In each case, the first stage offers negligible loading to a high-impedance source and delivers approximately unity gain throughout the range of the source. The second stage produces the required signal gain and stabilization. Use of this amplifier group assures circuit reproducibility with no concern for the varying parameters of discrete devices.

Output from each two-stage amplifier is introduced directly to Schmitt triggers U2A and U2B for square-wave shaping necessary to following logic. The Schmitt triggers have hysteresis values of nearly one volt between upper and lower trigger points. This characteristic is quite important to noise immunity when setting vox and anti-vox gain during operation. Signal inversion occurs in the vox amplifier channel, but no inversion is present in the anti-vox amplifier.

These conditions are essential to the positive CW start pulse and the negative CW stop pulse but they have no influence on radiotelephone transmission. In this way, activation and inhibition of CW vox occurs at the start and stop points of all code elements. Thus true break-in follows with only the inherent delay of the receiver mute circuit.

One section of the 7400 NAND gate, U4, provides a low at either U4A pin 1 or U4A pin 2, which will deliver a high at U4A pin 3. This condition is essential to the solid transmission *on* requirement for *tune,* and *PTT,* while retaining the syllabic voice response capability of normal vox transmission.

U4A drives U4B, an inverter, which activates Q1 into the open-collector cutoff condition during transmission and into saturation during reception. Under the open-collector transmission mode, 50 mA flows through CR5 and the 2.5 mH rf choke in the T-R switch. This action places the node at the junction of C11, CR5, C12, and the rfc at virtual ground during transmission. In the receiving mode, this nodal point offers no loading to the received signal other than the desired receiver input. Diodes CR4, CR6, and CR7 are protective devices in the event forward bias fails.

Output of CR9 from Q1 also drives the base input of Q4 to saturation during transmission. This condition allows Q5 to operate in the open-collector mode and permits full muting to the receiver through the mute line. While receiving, Q4 is cut off. Now, current flows through the 430-ohm collector resistor and CR15 to saturate Q5. The mute line is brought to ground in this manner for normal reception by the R-4B.

Transistor Q4 also drives Q6, which is emitter-biased to +1.5 volts by CR17 and CR18. When Q4 is saturated during transmission, Q6's base goes negative with respect to the emitter and saturates. When saturated, the collector potential of Q6 is approximately +1.5 V with respect to the collector of Q7. This action results in saturation of Q7, which removes cutoff bias to a kW power amplifier, if one is used.

Returning to the output, U4A pin 3, this point also triggers U3B pin 3. The output at U3B pin 6 now goes low for a constant period of 100 microseconds. With U3B pin 6 low, U4D pin 13 is held low and U4D pin 11 remains high during the 100-microsecond delay. The high at U4D pin 11 holds Q2 at saturation with a collector potential approaching ground. This condition holds cathode switch Q8 at cutoff for the delay period. The 100-microsecond delay for the T-4XB final amplifier is not significant, operationally. However, it provides full assurance that the T-4XB rf output cannot appear at the T-R switch until 100 microseconds *after* the T-R switch, mute line, and kW final-amplifier bias have been activated for transmission.

Both U4D pin 12 and U4D pin 13 must be high to force a low at U4D pin 11 during transmission. During reception U4B pin 6, driving inverter U4C, delivers a low at U4C pin 8. As long as U4C pin 8 remains low, no change of state at U4D pin 13 can alter the high at U4D pin 11. However, when both U4D pin 12 and U4D pin 13 are high, a low will hold at U4D pin 11. Propagation time for a change of state through U3B is somewhat greater than that of U4B and U4C. So, a small delay in the change from low to high at U4D pin 12 is desired. This delay is accomplished by means of a 270-ohm resistor and C15 (0.01 μF), which are shown at the output, U4C pin 8. This step permits propagation time for U3B and eliminates the possibility of a short, unwanted low transient at U4D pin 11.

The anti-vox signal is picked up through an extension cable from a point near the anti-vox gain pot in the T-4XB, and routed to the circuit board. This signal performs the inhibit function to U3A. If the output at U3C pin 6 goes low, any low existing at U3A pin 6 will promptly go high, having been inhibited by U3C. Also, when U3C pin 6 is held low, no activation of U3A can occur. This is an important and useful feature of the retriggerable monostable. Transmission, then, must be initiated between syllables and words. This is a normal occurrence in landline telephone conversations. The anti-vox circuit is quite similar to the audio and sidetone circuit, making further description of anti-vox unnecessary.

Note that no variable adjustments by pots and no component tailoring are required in this design. However, the normal T-4XB vox gain and anti-vox gain controls are used for noncritical settings of each function. The vox delay pot in the T-4XB is not used. In operation of the new vox method, no time delays are

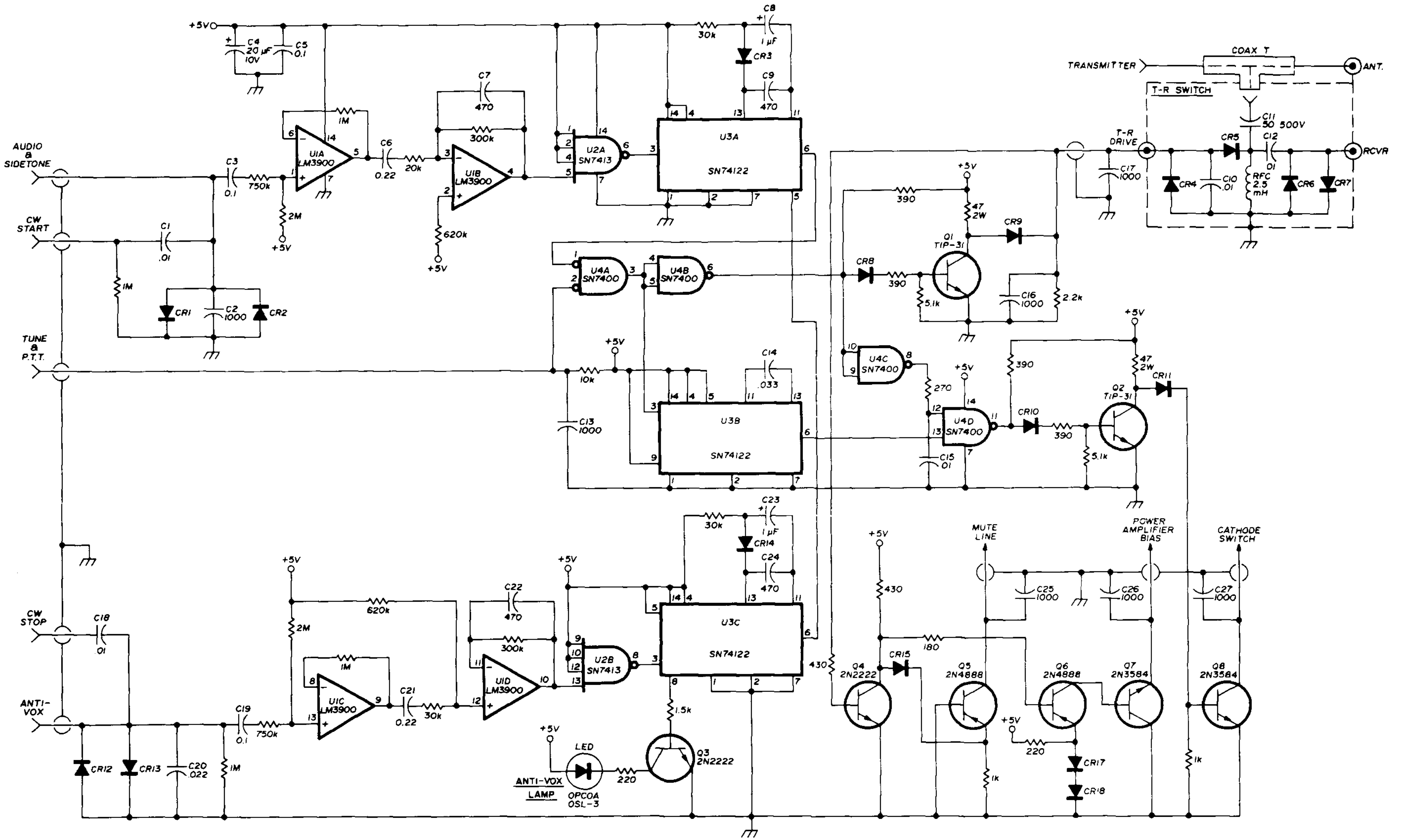

fig. 1. Syllabic vox and CW break-in system schematic for use with the Drake T-4XB and R-4B transmitter and receiver. Circuit provides instantaneous speech communication without conventional vox relays. Diodes CR4, CR5, CR6, CR7, CR9 and CR11 are 1N4004 or equivalent; all other diodes are 1N914 or equivalent. The rf choke is a 2.5 mH transmitting type (100 mA). All resistors are ½ watt, 10% unless otherwise specified.

present that can be observed by unaided human senses. Also no vox/anti-vox interaction occurs. During the initial setup, the threshold for vox activation is established by the vox gain adjustment. After this setting, an increase in voice level does not alter the vox response in any way. Similarly, the anti-vox signal threshold assumes a constant level for activation after adjustment of anti-vox gain.

setup procedure

The setup procedure is no more difficult and is perhaps simpler than that of conventional vox.

Step 1. With the receiver and transmitter in the ssb mode, the anti-vox gain, vox gain, and receiver audio gain pots should be set to minimum gain. With the voice at normal audio level and the microphone in its customary position, bring up the vox gain sensitivity only enough to observe vox enabling, which will be shown by a fluttering action of the S-meter between syllables and words. A sustained audio note will hold the S-meter at maximum indication. Do not change this setting of the vox gain pot in the next step.

Step 2. With receiver band noise only, turn up the receiver audio gain until the vox is triggered occasionally, which is indicated on the S-meter as in the first step. This audio gain level should be high relative to that used in normal signal reception. Continuing with the band noise conditions, increase the anti-vox gain until the anti-vox LED glows intermittently, while the S-meter remains at zero.

The anti-vox LED is useful when communicating with a very weak signal under high-level noise conditions. These conditions combine to present a "worst-case" operational example. In this situation, it is advisable to reduce receiver audio gain until the anti-vox LED does not glow intermittently due to adjacent frequency audio interference.

CW transmission

The CW start pulse in the T-4XB is used for fast vox initiation upon key or keyer contact closure at the start of a dot or dash code element. After starting, the sidetone signal holds the vox system *on* throughout the time duration of a dot or dash. However, the end of a dot or dash must inhibit the vox circuit promptly to prevent holding the mute line open for 8.4 milliseconds after the end of a dot or dash. This requirement is accomplished by negative pulse differentiation upon restoration of the grid bias line in the T-4XB. Then, the negative-going CW stop pulse is routed to the circuit board anti-vox circuit, which inhibits U3A.

power supply

The T-4XB normally uses a 6EV7 for vox gain and vox relay driving. Neither the vox relay nor the 6EV7 are used with the new vox system. So a simple heater wiring change is made that eliminates the 6EV7 heater load of nearly 4 watts. The new circuit board requires a separate, regulated power supply that delivers 5 volts dc at 425 mA. Elimination of the 6EV7 suggested use of the Drake AC-4 power supply for the 5 Vdc requirement without adding an additional load to the AC-4 unit. Consequently, a 5-Vdc power supply was designed for use with the 12.6 Vac transformer winding of the AC-4 (**fig. 2**).

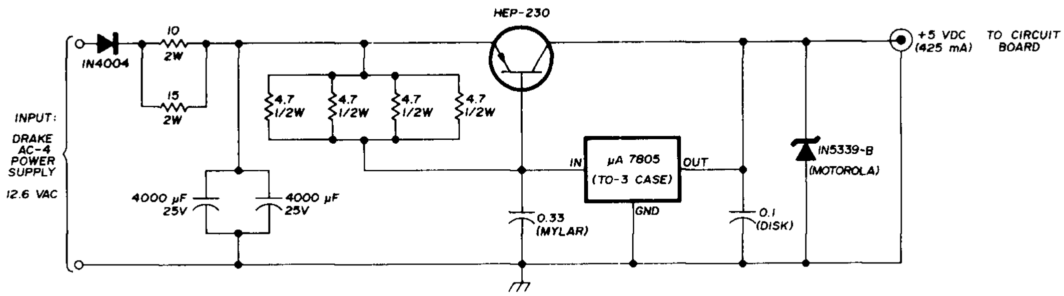

fig. 2. Simple 5-volt power supply for the syllabic system. Input power is supplied by the Drake AC-4 power supply. Unit is installed in the Drake MS-4 speaker cabinet.

A small circuit board was prepared to support all components of the 5-Vdc power supply. Then 12.6 Vac and ground leads were soldered directly to these transformer terminals inside the AC-4 unit. The new leads were brought out of the AC-4 case through ventilation holes and connected to the input of the 5-Vdc power supply This new supply was then located between the AC-4 power supply and the speaker in the Drake MS-4 speaker cabinet. This location offers more than adequate space for the 5-Vdc power supply, and there is no temperature problem. Incidentally, this method of connecting the available 12.6 Vac power automatically places the 5 Vdc power under the control of the power *on/off* switch on the T-4XB panel.

T-R switch

The T-R switch used in this system is a compromise with respect to electromechanical relays. The titles "T-R switch" or "diode switch" are erroneous in technical fact. But these titles are used since correct and definitive nomenclature is not immediately available. Casual observation of the circuit may lead to the false conclusion that the T-R switch is a simple device for receiver protection. However, a rigorous, quantitative analysis of this device is beyond the scope of this article. The interested reader is invited to bypass the T-R switch with a continuous 50-ohm line from antenna to receiver for comparison testing. Such tests conducted at my station displayed no discernible difference of signal strength on the S-meter.

Like other electronic T-R switches, this switch must be handled properly to avoid signal suckout. In this case, the switch enclosure is mounted directly to the T-4XB output connector. Rigid coaxial hardware is used and provides secure mechanical support for the small T-R switch can. No signal suckout will occur with this installation of the switch unless the receiver and transmitter are tuned in different bands. When operating normally in a given band, improvement in signal-to-noise ratio may be observed. This is because the received signal looks into the output tank of the transmitter before appearing at the receiver.

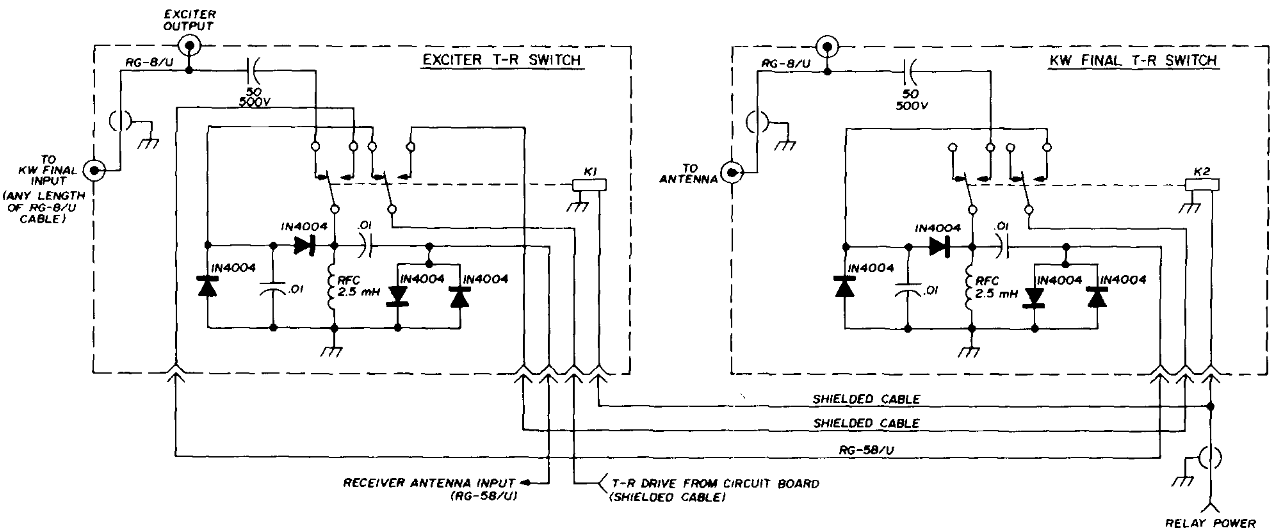

fig. 3. T-R switch schematic for use with kilowatt final amplifiers. Suggested relays are Potter & Brumfield type KT-11D. Each T-R enclosure should be mounted to each output by rigid coaxial connectors. A bud type CU-234 utility box is suggested for each T-R switch.

T-R switch with linear amplifier

When a final amplifier is used, the dual T-R switch method should be used, and the second T-R switch should be mounted rigidly to the output of the kW amplifier. See **fig. 3** for a schematic and component details. In this case, two changes must be made in the control of the kW amplifier:

First, the power amplifier bias control must be removed from the antenna relay section that normally switches cutoff bias control to ground. Then, the bias control lead should be a direct, isolated conductor from the bias control point in the linear amplifier to the power amplifier bias terminal on the new circuit board in the T-4XB. It is assumed that the bias control point in the final amplifier has a negative potential not exceeding -150 Vdc and current loading not greater than 150 mA when grounded. Positive cathode bias is not considered in this article.

Second, the antenna relay in the kW amplifier must be wired so that *on/off* control of this relay is performed by the *power on* switch on the kW amplifier panel *only.*

These changes were made to an NCL-2000 amplifier in less than one hour. The end result was a pleasing lack of clacking noise from the cumbersome relay in dynamic operation.

T-4XB changes

The following steps define installation of all extension leads, minor changes, and cabling required to the T-4XB chassis. No holes are drilled, no switches added and no defacement of components or chassis are required. (The Drake T-4XB manual is referenced).

1. Connect a jumper wire between socket terminals 4 and 5 of V10.

2. Disconnect lead from pin 5 of V9 to pin 3 of V11.

3. Disconnect lead from pin 9 of V9 to ground.

4. Connect new lead from pin 9 of V9 to pin 3 of V11.

5. At the vox relay *ant* section to J7, connect a jumper wire between the normally open contact and the movable contact. Then disconnect the lead to *rcvr* J6 from the normally closed contact.

6. At the vox relay mute section to J8, disconnect the movable contact from ground. Connect a board extension lead labelled "mute line" to the normally closed contact. Then connect another extension lead labelled "cathode switch" to the normally open contact using no. 22 AWG (0.6 mm) or larger shielded wire.

7. The third section of the vox relay is not specified but is used to control the grounding of kW-amplifier negative bias in this application. Connect an extension lead labelled "power amplifier bias" to the normally open contact of this section. The 2N3584 will easily handle a 150-mA load at -150 Vdc if required.

8. Locate the PTT contact on microphone jack J3 and connect an extension lead labelled "tune and PTT" to this contact.

9. Locate diode CR7 on the small circuit board at the right rear section of the T-4XB chassis. Apply soldering iron and pull out the cathode end of CR7. Then connect an extension lead labelled "CW start" to the exposed cathode of CR7.

10. The anode of CR7 connects to capacitor C131 (0.01 μF), and the other side of C131 is connected to R78 (1M). At the junction of R78 and C131, connect an extension lead labelled "CW stop." Now solder a 1000-pF disc capacitor across R78. (Incidentally, this is the only component added to the T-4XB chassis).

11. Locate C92 (0.02 μF) and CR12 in the anti-vox circuit. Then connect an extension lead labelled "anti-vox" at the junction between C92 and CR12.

12. C142 (0.02 μF) is connected at the junction of R98 (3.3M) and CR9. Connect a jumper across C142.

13. Connect an extension lead labelled "audio and side-tone" to the center wiper contact of the vos gain pot, R89.

14. A shielded cable is connected to the T-R terminal on the circuit board and routed outside the T-4XB cabinet to the designated terminal of the T-R switch. The cable delivering +5 Vdc power to the circuit board and the cable to the T-R switch may be held together and clamped by a convenient screw already located on the T-4XB chassis.

This completes all changes and extension-lead connections to the T-4XB chassis. None of these terminals lack adequate access and offer no problem in location. At trade-in time, the T-4XB may be quickly restored to its original design with no loss in value.

results

In radiotelephone practice, you'll experience some new and pleasant features when using this system. There is no vox-on presence between words or after a word of transmission. Only words and syllables of words appear on the air. Also there are no vox rf transients. The operator may be broken at any time and "doubling" is eliminated. Extraneous local noise does not appear on the air because the voice frequencies overwhelm the noise. If the transmitting operator wishes to force transmission, he may use PTT in the normal manner.

In CW service, the only locally generated noise is from the handkey, keyer, or power-amplifier blower. Only sidetone is heard in the earphones or speaker. Clicks and TVI are absent.

references

1. H.R. Hildreth, WØIP, "Instantaneous Voice Interruption," *QST*, October, 1968, page 40.
2. H.R. Hildreth, WØIP, "More on Instant Voice Interruption," *QST*, June, 1972, page 19.

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electrical units:

their derivation and history

An interesting sidelight on measurement units and their relationship to the metric system

A lot of people who work in electronics have to deal on a day-to-day basis with quantities such as voltage, resistance, and current and the units used to measure these quantities. Surprisingly few, however, know how the units of measurement were established, and it almost never fails to raise an eyebrow when they find out that all the basic quantities — volts, amperes, etc — are metric units that can be defined in terms of kilograms, seconds, and meters.[1]

It's strange to think that the amount of inductance we call 1 henry can be traced back to the diameter of the earth, or that 1 farad of capacitance owes part of its definition to the weight of water. It's stranger yet to realize that these quantities wouldn't have the values they do if it hadn't been for the downfall of King Louis XVI *in the French revolution of* 1789 — but that's how it is.

basic units

The watt, as a unit of power, and the ampere, as a unit of current, provide the foundation for defining the other units in electronics. Watts and amperes owe their definitions to the field of physics, which we'll look at later. The ohm, as a unit of resistance, arises naturally from the ampere and the watt. It was found that only a specific amount of resistance would produce 1 watt of heat if 1 ampere were flowing through it. This value of resistance is *defined* as 1 ohm, and the potential drop developed across this resistance under these conditions is *defined* as 1 volt. Even though an infinite number of combinations of voltage and current will produce 1 watt of heat in a resistor, there is only one value of emf that will do it with 1 ampere of current, just as there is only one value of resistance that will draw 1 ampere with 1 volt across it. Accordingly, there is no ambiguity in the values: they are clearly defined and have no "double values."

The ampere, as a unit of current, is commonly thought of as a quantity, but actually it is a *rate*, or quantity per unit of time. Its hydraulic equivalent would be gallons per unit of time, or liters per unit of time if you want to stay in the metric system. The electrical equivalent of these gallons or liters is the coulomb, which signifies the actual quantity or number of electrons that will sustain 1 ampere of current for 1 second. In round numbers, 1 coulomb equals 6.25×10^{18} electrons. It's more commonly called the unit of charge, and since it defines a quantity instead of a rate, its symbol is Q. Incidentally, the ampere was originally described as the unit defining the *intensity* of a current, hence the use of the letter I in Ohm's and Kirchoff's laws to symbolize it.

derived units

The units of capacitance and inductance owe their definitions to the volt, ampere, and second. In the case of capacitance, it was discovered that no current would flow into or out of a capacitor unless the voltage across it was *changing.* A further investigation revealed that if the voltage changed at a constant rate, the current would also be constant. If the voltage increased, the current would be a charging current; if it decreased, the capacitor would discharge current back into the circuit. A still closer look revealed that, by doubling or tripling the rate of voltage change, the current would proportionately be doubled or tripled. As a result of these discoveries, 1 farad of capacitance is *defined* as the amount of capacitance that will draw a 1-ampere charging current if the voltage across it increases at the rate of 1 volt per second.

Inductance was found to have similar qualities, but with the roles of voltage and current interchanged. That is, no voltage would exist across a coil unless the current flowing through it was changing. If the current was increased at a constant rate, the voltage developed across the terminals of the coil would also be constant; and doubling or tripling the rate of current change would proportionately double or triple the voltage developed across the coil.

One henry of inductance was therefore *defined* as the amount of inductance that would sustain a 1-volt drop across its terminals if the current increased at a rate of 1 ampere per second.

Thanks to this system of definitions, proportionality constants are conspicuously missing in the equations used in electronics. The factor 2π, which crops up in the equations for reactance and resonant frequency, is the only exception; and even here it is unnecessary if the frequency is expressed in terms of radians per second instead of cycles per second — but that's another story.

power and energy

It was stated earlier that the ampere is not a quantity but is a rate of change per unit of time. It would be just as correct to say that a given current was 50 millicoulombs per second as it would be to call it 50 milliamperes. The watt, as a measure of power, is also a measure

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of rate instead of quantity. Just as 1 coulomb per second is commonly called 1 ampere, 1 watt could also be called 1 joule per second. Here arises the distinction between power and energy: energy is a *quantity*, whereas power expresses the *rate* at which this energy is expended.

An example would be appropriate here. Suppose we have a quantity of water at room temperature and wish to heat it to 10 degrees above room temperature. Let's further specify that the quantity of water is chosen so that it will require 5000 joules of heat energy to do this. We could put a resistor into the water and adjust the power fed to the resistor so that it dissipated 50 watts of heat into the water. At this rate, it would take 100 seconds to heat the water. If the power were increased to 200 watts, it would take only 25 seconds; at 500 watts, 10 seconds; and so on. In each case the same amount of work is accomplished; the only difference is in the rate at which it was done and the amount of time required. The joule is a unit of work or energy and has its definition in the roots of the metric system.

measurement systems

A few words about measurement systems are in order. Basically, there are three major systems used in the world to measure quantities and rates on a scientific basis. These are a) the familiar English system, based on the foot for length, the pound as a unit of *force* (not mass), and the second for time; b) the CGS system; and c) the MKS system.

The CGS system (centimeter-gram-second) is the system that Einstein used in his work in relativity. The MKS system (meter-kilogram-second) is the system on which electronic units are based. These last two systems are metric and are based on the meter, gram, and second, which were established in France as a new system of measurement after the downfall of the ruling nobility in 1789. The meter, as a measure of distance, was defined at the time as representing one ten-millionth of the distance from the North Pole to the equator. The gram was also defined as the amount of mass represented by 1 cubic centimeter of distilled water cooled to 4 degrees Celsius. (This is the temperature at which water is most dense). As a unit of time, the second was retained, having its definition in the motion of the earth. The liter, as a unit of volume, was incidentally defined as representing 1000 cubic centimeters.

force and work

In the MKS system, the first derived unit of measure is the newton, which is a measure of force, and is defined as follows: If a 1-kilogram mass were placed on a frictionless surface, this mass could be accelerated from a dead stop to any speed by exerting force on it. The greater the force, the more quickly the mass would accelerate. One newton of force, if exerted on this 1-kilogram mass for 1 second of time, would bring it from a dead stop to a speed of 1 meter per second.

The unit of work in the MKS system is the joule and derives its meaning from the newton and the meter. It is defined as the amount of work expended by moving an object across a rough surface for a distance of 1 meter, if 1 newton of force is required to move it. If the object requires 2 newtons to move it, and it is pushed 2 meters, the work would be 4 joules. The work expended equals the product of the force and the distance.

Ham Radio is designed to provide something for everyone. This broadbush treatment of electrical units, their derivation, and history is presented to encourage further reading in the world of physics, the basis of our electronic heritage. A more rigorous definition of international units including a table of physical constants with their symbols, values, units, prefixes and least-squares error adjustments appears in reference 1. Also included in reference 1 are ten pages of conversion factors to get you out of the awkward English system and into the scientific metric system of measurement. **Editor**

power

Notice that the amount of time required to perform this work does not change the amount of work that is done. Whether the work is done slowly or quickly, the same amount of energy is expended to move the object. The rate at which this work is done is called the *power,* or energy per unit of time. In the first example above, 1 watt of power would accomplish the job in 1 second. In fact, this is the definition of 1 watt: expending energy at the rate of 1 joule per second. In the second example, 1 watt would accomplish the job in 4 seconds, 2 watts in 2 seconds, or 4 watts in 1 second.

So much for the newton, joule, and watt as mechanical units. Only the ampere now has to be defined in mechanical units to completely relate all the electronic units to the world of physics. The ampere is defined under these conditions: If two very long wires (theoretically, infinitely long) are placed parallel to each other and spaced 1 meter apart, and if current is run through them (same value of current for each wire), then the magnetic fields they set up will tend to force the two wires either together or apart, depending on whether the currents are flowing in opposite or in the same direction, respectively.

If the amount of this force exerted on a 1-meter length of either of the wires has a magnitude of 2×10^{-7} newtons then the current is defined as having an intensity of 1 ampere. I don't know why 1×10^{-7} newtons wasn't chosen to define the force created by one ampere, but perhaps it was thought that each wire actually does exert this force, so that the total force acting on a 1-meter length would be 2×10^{-7} newtons – half of it due to each wire's magnetic field. This is all pure speculation, however, and should not be construed as accurate.

I hope this article has shed some light on our heritage in the field of physics without too much confusion. In a country that must go through the withdrawal pains of the English system of measurement, it's reassuring to know that we work in a field that has always been metric.

reference

1. *The International System of Units,* NASA Publication SP-7012, Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402, price 30 cents.

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More Details? CHECK-OFF Page 126

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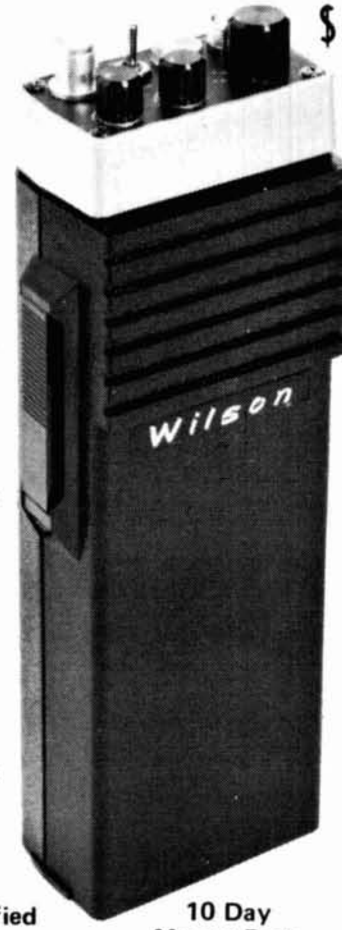

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More Details? CHECK—OFF Page 126

i-f/detector receiver module

Design and construction data for an i-f strip, detector, and audio filter that can be used for a high-performance amateur-band receiver

This article provides design and construction details for a high-performance *i-f/detector receiver module. By itself* the circuit is an 80-meter-band receiver, requiring only a frequency-determining oscillator and an audio stage for on-the-air use on the high-frequency amateur band. Performance data is given in **table 1**.

The sensitivity of this receiver is less than 0.1 microvolt rms for a signal-to-noise ratio of 10 dB. What does this really mean? Looking through the ads, you'll find a number of claims as to the sensitivity of a receiver at some signal-to-noise ratio. What's most important (and always missing in these claims) is a definition of terms. For example, one manufacturer tells us that his receiver has a sensitivity of 1 microvolt for a 10-dB signal-to-noise ratio. The question is, "Is this 1 microvolt rms or 1 microvolt peak-to-peak?" By proper attention to details in the construction of the receiver described here, you should be able to measure its sensitivity at less than 0.1 microvolt *rms* for a 10-dB s/n ratio, not 0.1 microvolt peak. At 1 microvolt rms, you should be able to measure a 10:1 voltage difference (20 dB) between signal and noise. I made my measurements using an HP-606A signal generator and a Tektronix 513 oscilloscope.

design features

The block diagram of the i-f/detector (**fig. 1**) illustrates the overall design. Good image rejection is obtained by initial amplification at the higher frequency followed by a mixer stage with low noise and modest conversion gain. Amplification and passband characteristics are obtained with a multipurpose IC, using either a crystal or ceramic filter, or an LC tuned circuit at the conversion frequency. In any receiver system the detected audio will contain beat oscillator and spurious noise components; these are reduced by passing the signal through an active low-pass filter IC, which is matched to an IC amplifier stage.

The i-f/detector schematic is shown in **figs. 2A** and **2B**. Values for resonant-circuit components are shown in **table 2**. Note that in the schematics, capacitors are shown as a fraction; for example, 0.1/C. This means that the capacitor is a 0.1 μF ceramic. An M below the value means the capacitor is mica.

The 3.5 to 4.0 MHz rf signal is amplified by Q1 and Q2. Tuning of both stages is accomplished using a varactor diode, which eliminates the need for a large variable capacitor. Dual-gate mosfets allow external agc and rf gain control if desired. For applications where an agc and rf gain control isn't required, a single-gate device could be directly substituted, or the control gates, G2, of Q1 and Q2 may be tied to some fixed voltage value. Down-frequency conversion is accomplished by Q3, a standard mosfet mixer. The input tuned circuit of Q3 is fixed and broadly tuned; the output is matched to a commercial i-f transformer. Biasing of gate 2 is accomplished by using a gate-to-source resistance that establishes the operating point for the device.

Overall system voltage gain, detection, and passband characteristics are developed by U1, a multipurpose IC. The schematic illustrates a Collins mechanical filter for

table 1. Performance data for the high-performance *i-f/detector* module.

frequency range	3.1 to 4.5 MHz with suitable hfo
nominal operating frequency range	3.5 to 4.0 MHz
high frequency oscillator range	3.045 to 3.545 MHz ≈ 2V, peak-to-peak
sensitivity	<0.1 μV rms for 10 dB s/n ratio (distinguishable signal of 1 kHz)
noise margin	>20 dB at 1 μV rms
shape factor	dependent upon filter. Normally 60 dB with Collins of Murata ceramic filters
in-band spurious signals	none using chassis and shields as indicated
audio	0 to 3 kHz; 5 kHz notch; 15 to 20 dB minimum attenuation on signals >5 kHz
nominal gain	80 to 100 dB depending on link positioning and *control voltages*

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passband shaping. You may prefer a ceramic or simple transformer-type component instead; the overall operation of U1 will remain the same regardless of the type of filter used. Variations in gain can be expected using crystal and transformer filters, which have operating impedances differing from the ceramic type shown.

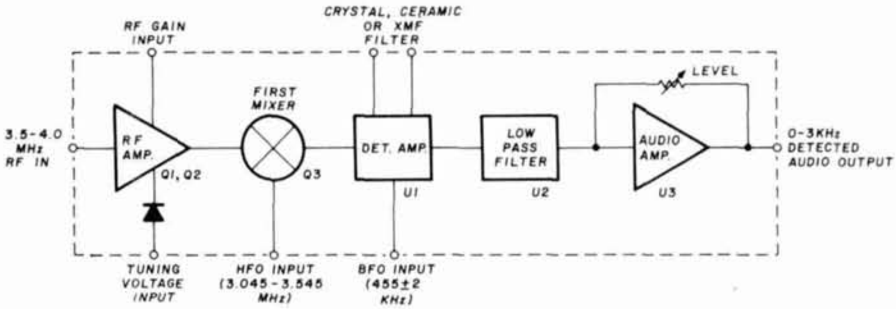
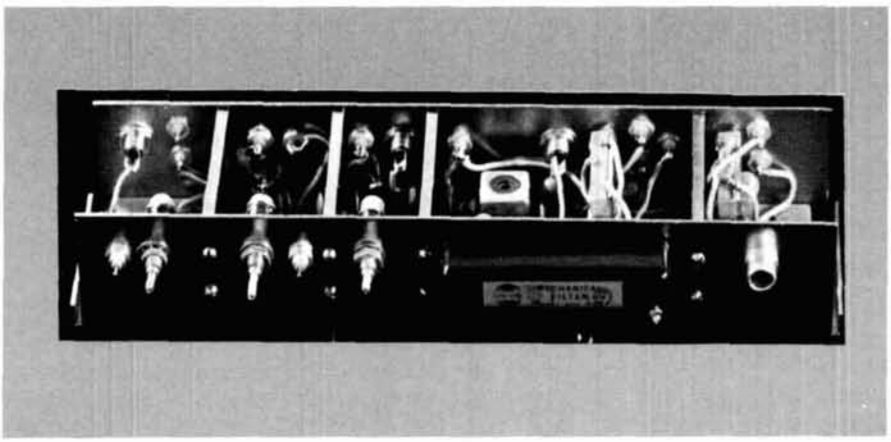


fig. 1. Block diagram, high-performance i-f/detector receiver module.

The detected audio signal on U1, pin 7, will contain many beat-oscillator audio components or hiss-type noises in addition to the desired detected audio. The most common method of eliminating these undesirable noise products is to use a toroidal LC or multiple op-amp active low-pass filter. These methods work well but require large areas on the chassis and can be expensive.

An inexpensive active device is available that operates as an elliptic lowpass filter having a rolloff frequency of 2.5 kHz. The attenuation at 5 kHz is approximately 18 dB with increasing attenuation of all higher-order components. U2's output is matched to the input of U3, which operates as a voltage amplifier to compensate for the loss in gain through U2. U3 is a straightforward operational amplifier with internal frequency compensation.



High-performance i-f/detector receiver module, cover removed. Enclosure is an LMB 850 box. Rf input is at right; audio output at left.

construction

The photographs show the construction approach. A PC board carries all active components with input-output of the stages arranged along the LMB 850 chassis sides through feed through terminals.* This circuit is a classic i-f amplifier approach that will provide excellent interstage shielding and minimum feedback coupling which will eliminate oscillation. Interstage shields are aluminum strips fastened to the PC board and chassis sides. These shields provide all the mechanical rigidity required.

There are two methods of building the i-f/detector module: the right way and the sure-failure way. The *right way is to build the assembly one stage at a time,* testing each stage as you go. The sure-failure way is to install all components, assemble, and apply power expecting it to work!

Construction should start with U3. Install its associated components, temporarily apply power and signal inputs, then verify the correct output signal. The next step is to install U2, its associated components, and temporarily apply power and signal levels to U2 to verify its operation. This procedure should continue with U1, Q3, Q2, and finally Q1. Proceed through each stage to ensure proper operation. Using the step-by-step approach, you'll become familiar with circuit operation and can optimize the stages for your own preference.

Figs. 3A and **3B** illustrate the PC board component locations. The board is designed to accomodate ¼-watt resistors and low-voltage capacitors; larger components may be used by mounting them vertically with respect to the PC board.

*Printed-circuit boards are available from the author for $3.00 plus postage.

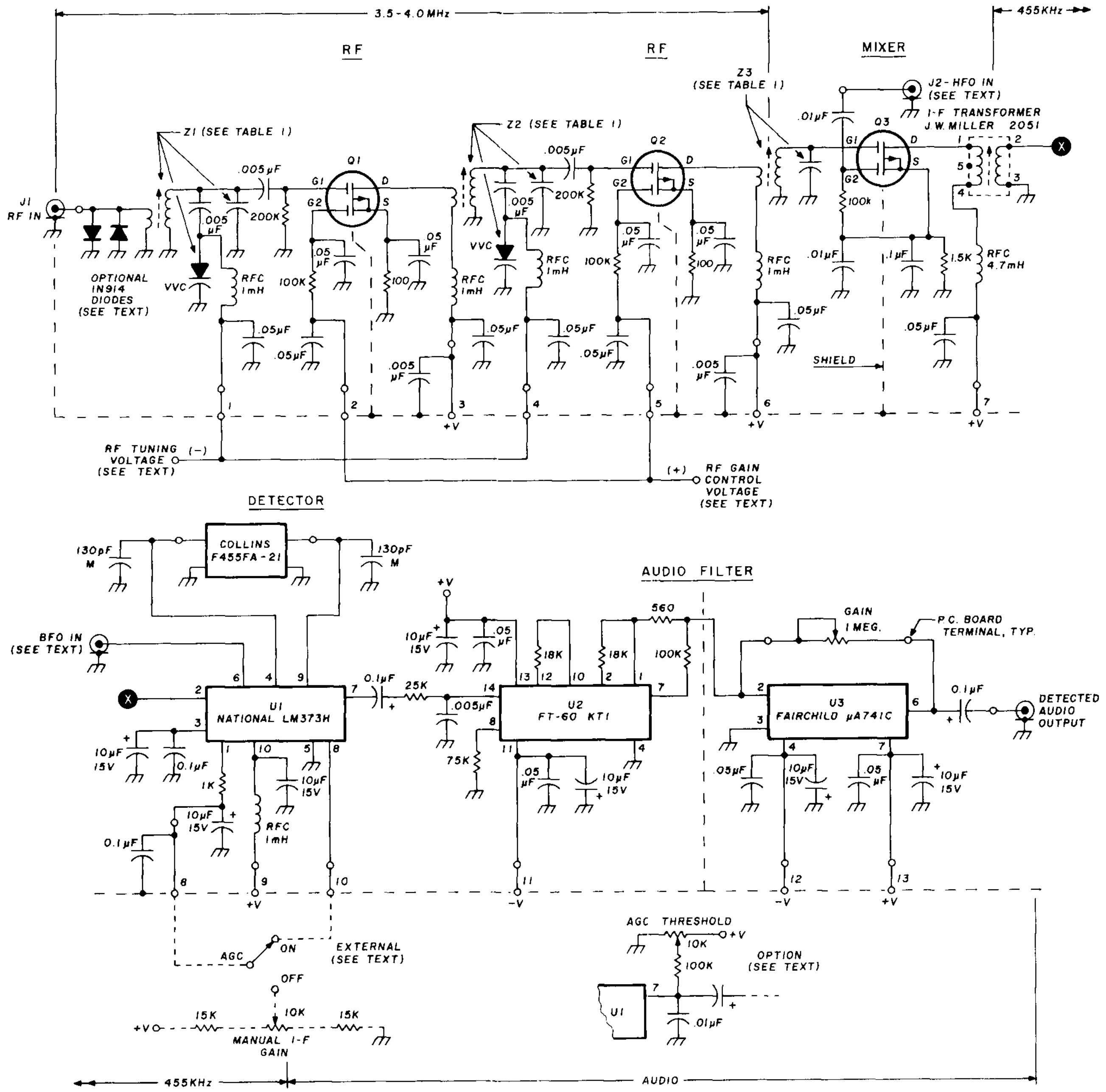
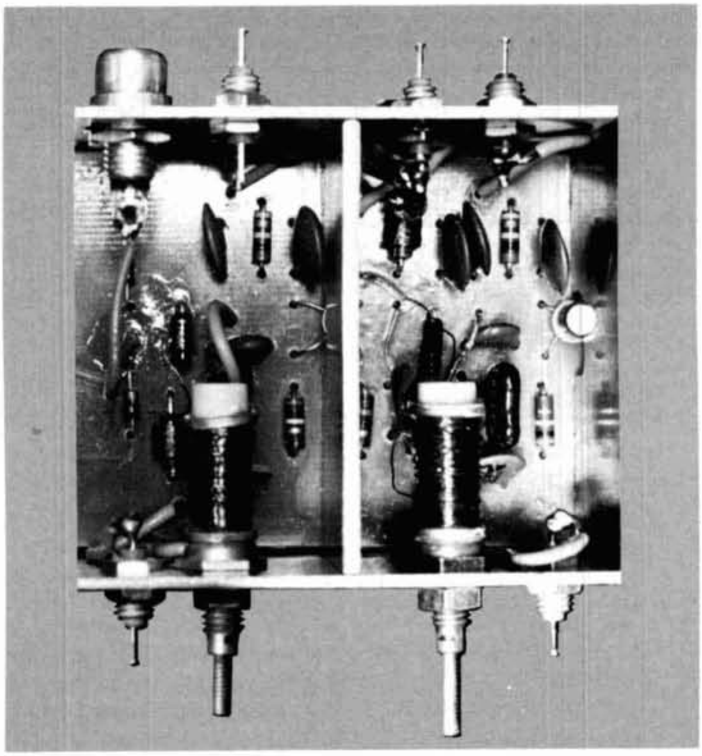


fig. 2. Schematic diagram of the high-performance i-f/detector module. Tuned circuit values are listed in table 2.

step 1. U3 installation. Install U3 and all associated fixed PC-board components. The 1-megohm feedback resistor may be left freely supported over the top of U3 using bus wires from the PC board. Temporarily install one end of a 100-ohm resistor in the +12 and −12V terminals. These resistors limit the current during test in the event of a wiring or circuit error. Apply a 15mV 1-kHz audio signal to pin 2 of U3, and with power applied, adjust the feedback pot so that a 15-volt peak-to-peak signal appears at pin 6; this signal may be taken from the output side of 0.1/C capacitor of pin 6.

step 2. U2 installation. Install U2 and all associated components up to the 0.1/C coupling capacitor from pin 7 of U1. Install the 100-ohm current-limiting resistors for both the +12 and −12V terminals, as you did for U3. With power on both U2 and U3, apply a 100mV 1-kHz audio signal to the U1 pin 7 side of the 0.1/C capacitor. While monitoring the output of U3, adjust the U3 feedback pot for an output voltage of 100 to 150 mV. Increasing the frequency of the audio input signal should show decreasing output values at frequencies above 3 kHz.

U2 can be optimized by adjusting the resistor from pin 8 to ground and the 560-ohm series output resistor from pin 1, so that gain and rolloff slope are sharpened. The CW buff will probably want to increase the 18k



Rf amplifier section. Q1, Q2 gates are in the center. Note shielding and cutouts for devices.

resistors between pins 10 and 12 and pins 1 and 2 to 22k, which will bring the rolloff to approximately 1 kHz.

step 3. U1 installation. Install U1 and all associated components up to the 0.01/C coupling capacitor from the 455-kHz i-f transformer secondary. During initial test a 0.01/C capacitor can be installed between pins 4 and 9 rather than a filter, and a jumper wire can be added between pin 8 and the output side of the 1k agc feedback resistor from pin 1. Use the PC board terminals provided for this installation. Initially apply power to U1 only and verify that the first section of U1 is working correctly by injecting a 1-mV modulated 455-kHz signal into the input of U1 at pin 2 through the 0.01/C coupling capacitor. By monitoring pin 4, a gain of 25 to 30 dB should be apparent.

table 2. Resonant-circuit component data.

Z1 and Z2	Z3
inductor: 45 to 50 turns no. 32 AWG (0.2mm) enamelled wire on Miller 4500-2 or 4500-3 coil form	45 turns-same coil form as Z1 and Z2
VVC: Motorola MV1646	None
capacitor: 15 to 20 pF npo or mica type	None
link: 8 to 10 turns, evenly spaced over winding (see text)	same as Z1 and Z2

Fig. 4 is a schematic for the suggested bfo circuit. Initial bfo injection levels should be set at 50 to 75 mV rms; higher values may cause the second section of U1 to oscillate. With injection of both bfo and carrier signals a detected audio signal of 50 to 75 mV should appear at pin 7 of U1. The bfo voltage injection level should be increased to the point of U1 oscillation then decreased slightly. Remove the coupling capacitor between pins 4 and 9 of U1, and install the bandpass filter, using jumper wires from the PC board. Apply power to U1, U2, and U3 and inject the bfo and 455-kHz carrier as before. With 1 mV applied at the input of U1 a clean 90 to 100 mV audio signal should be apparent at U3 output. To simplify initial alignment and test, the bfo frequency should be approximately ±1 kHz from the center of the filter. For instance, using a Collins ceramic filter, I use a

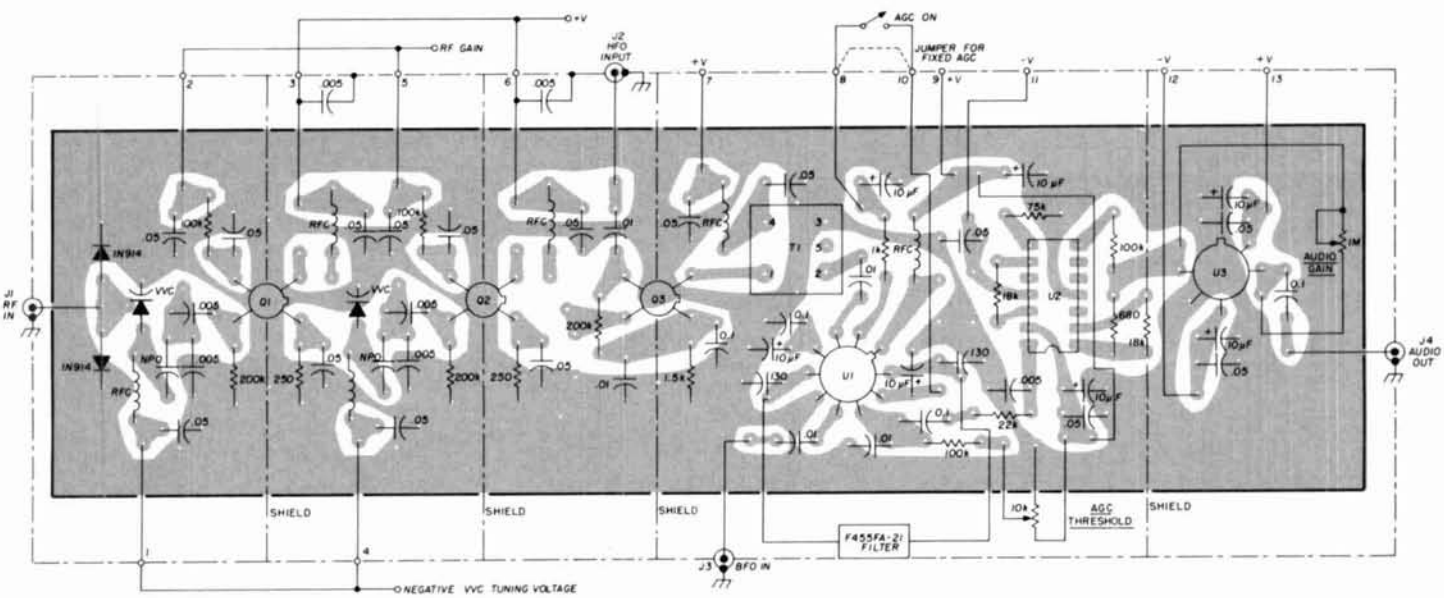


fig. 3. Component layout of rf amplifier, mixer, detector, and audio filter.

455.95-kHz crystal for the bfo frequency, which develops an easily distinguishable 1-kHz audio signal at the output with unmodulated carrier injection.

step 4. Q3 installation. Install all components associated with Q3. Adjust Z3 coil slug for resonance at 3.750 MHz. The input link should be loose along the outside of the coil but terminated to the PC board. Apply power to Q3 and U1 only, and inject a carrier voltage of 3.750 MHz at the link of Z3 and a beat oscillator signal from an external oscillator into G2 of Q3. Reference 1 discusses the design of a suitable high-frequency oscillator that matches this unit. By spreading and sliding the link portion of the Z3 coil input for maximum signal output,

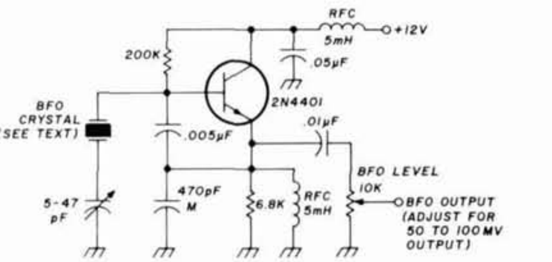
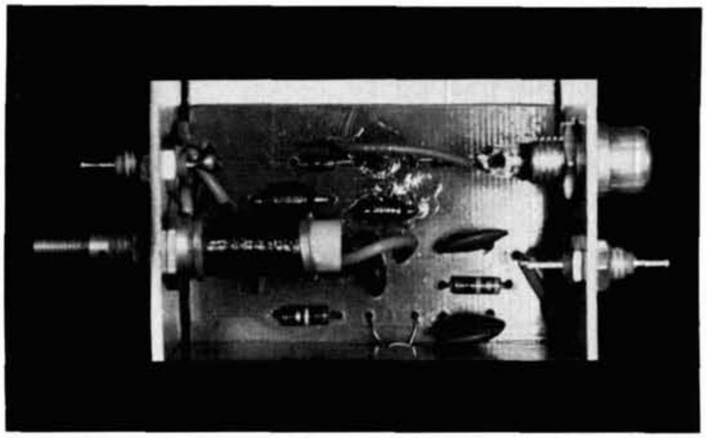


fig. 4. Bfo schematic. Crystal frequency depends on filter, as discussed in text.

a 50-μV signal at Q3 link input should provide a 30-mV detected audio signal at U1, pin 7.

step 5. Q2 installation Install Q2 and all associated components except Z2. Apply power to Q2 and Q3 only, and inject a 3.750-MHz carrier into G1 of Q2 while monitoring the drain of Q3. Temporarily apply +12 volts to the input side of the 100-k resistors of G2. Make the final link adjustment on the output of Q2 for a voltage gain of Q2 equal to 8 or 10. Apply a small amount of



Mixer input section. Q3 gates are on the right; HFO input is through phono jack at upper left.

approximately −10 volts while positioning Z2 slug for resonance. The link may be moved over the surface to the coil to optimize this coupling.

step 6. Q1 installation. Repeat all installation procedures used for Q2 on Q1. Adjust Z1 slug so that the same corresponding VVC tuning voltage is applied to both resonant circuits in the Q1 and Q2 stages. The operating characteristics of Q1 and Q2 can be verified by alternating grounding G2 of Q1 and Q2 for a short time while monitoring the output of Q3 at either the drain or at the i-f transformer secondary. Temporary grounding of G2 for either Q1 or Q2 should result in an immediate decrease in the output signal level. This should demonstrate the AGC operating feature of the dual-gate mosfet. With the ground removed, the output signal should slowly rise to full value. If you're not interested in using an external rf gain control, you could substitute a single-channel fet for Q1 and Q2 and adjust the source

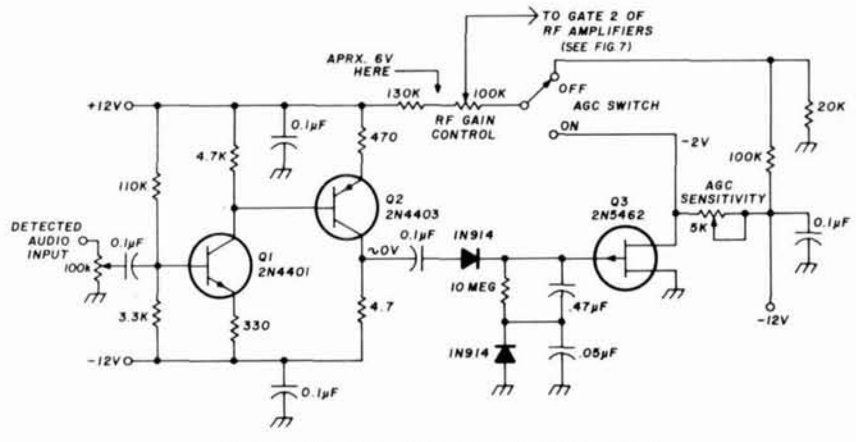
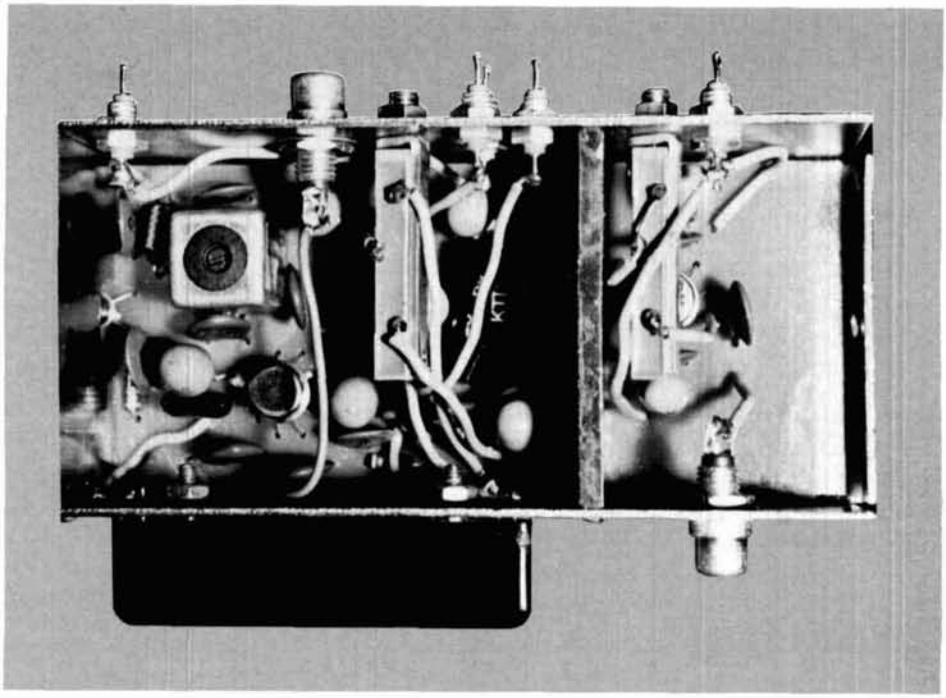


fig. 5. Suggested circuit for agc and rf gain control.

coil dope to the link to secure it in this position. Install Z2 and the associated VVC components. Apply power to Q2 and Q3 while injecting a low-amplitude 3.750-MHz carrier to the link of Z2. Apply −20 Vdc through 100k pot, and adjust the voltage on the VVC to

resistance for adequate feedback. Maximum gain is achieved when the source terminals of Q1 and Q2 are at ground potential; however, optimum noise margins are obtained using a source resistance of approximately 220 ohms.



Mixer output, i-f amplifier, and audio filter. Bfo input is through phono jack at right of the 455-kHz i-f transformer. Audio output is at phono jack, lower right.

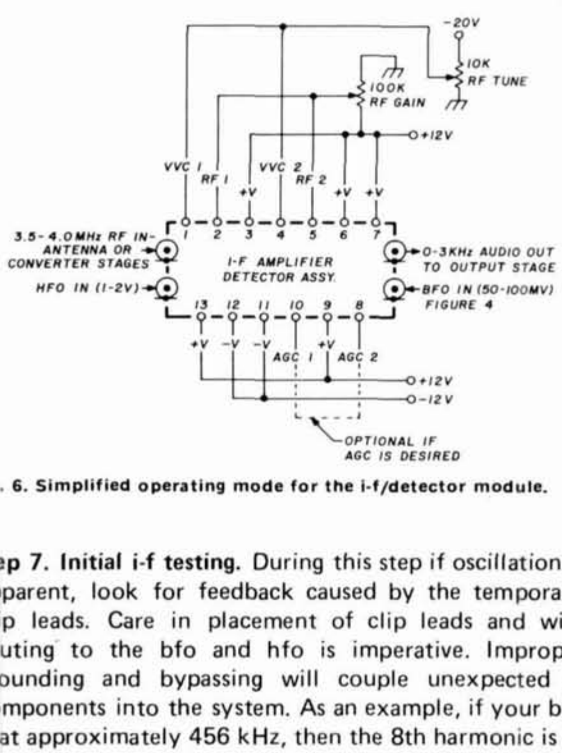


fig. 6. Simplified operating mode for the i-f/detector module.

step 7. Initial i-f testing. During this step if oscillation is apparent, look for feedback caused by the temporary clip leads. Care in placement of clip leads and wire routing to the bfo and hfo is imperative. Improper grounding and bypassing will couple unexpected rf components into the system. As an example, if your bfo is at approximately 456 kHz, then the 8th harmonic is at 3.468 MHz, which can be picked up by the module if you have poor shielding in the high-gain stages.

While keeping the current-limiting resistors in place, carefully apply power to each succeeding stage, starting at U3 and proceeding through Q1. If you haven't made a polarity error, poor positioning of clip leads, or injected too much bfo voltage into U1, the system should be quite stable. Monitoring the output of U3 should display low-level random audio output noise, which should not change even while Q1 input is alternately grounded or left open when the hfo is tuned over the entire input range.

Inject a 5 μV rms signal at 3.750 MHz into Q1 input. While monitoring U3 output with the hfo tuned to the beat frequency, an audio signal of approximately 0.1 V peak-to-peak at 1 kHz should be obtained. Some adjustment in the VVC tuning pot may be necessary to peak the signal.

Strictly speaking, the gain of a system is defined as:

$$dB = 10 \log \frac{|V_2^2|/R_2}{|V_1^2|/R_1}$$

$$\text{If } R_1 = R_2 \text{ then } dB = 20 \log \left|\frac{V_2}{V_1}\right|$$

This relationship is widely misused, and in voltage ratios the gains are expressed in decibels even though $R_1 \neq R_2$.

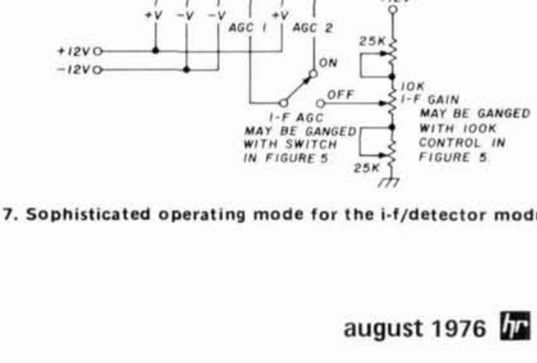


fig. 7. Sophisticated operating mode for the i-f/detector module.

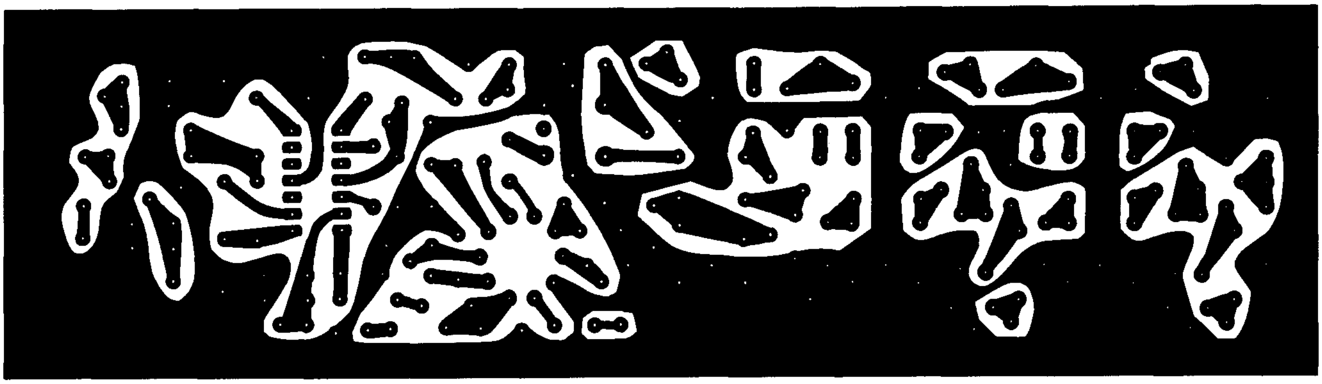


fig. 8. Printed-circuit board layout for the rf amplifier and mixer (left side of board) and the detector and audio filter. Board is 7-7/8" (20cm) long by 2-1/8" (5.4cm) wide.

Furthermore, we had a carrier of 3.750 MHz on the input and are examining an audio signal on the output. If we ignore this small impropriety, we can calculate the i-f/detector gain as:

$$dB = 20 \log \left| \frac{output\ voltage}{input\ voltage} \right|$$

The measured output voltage is 0.1V peak-to-peak, or

$$V2\ rms = \frac{0.1}{2.828} = 0.0353\ rms$$

and the input

$$V_1\ rms = 5(10^{-6})\ V$$

$$dB = 20 \log \frac{0.0353}{5(10^{-6})} = 76.9\ dB$$

final assembly

Remove the temporarily installed current-limiting resistors and install short pieces of bus wire for feedthrough access from the chassis. Install the PC board assembly into the chassis as shown in the photographs, terminating the bus jumpers to feedthrough terminals. Twist the inductors slightly then slide them into mounting holes along the side of the chassis.

For final testing the assembly should be wired as shown in **fig. 6.** Readjust the coils slightly, and if the agc threshold trimmer is used, adjust it for approximately 1 volt. The agc jumper wire between pins 8 and 10 may be removed for initial testing. Apply a 1-μV signal to the input, adjusting the coil slugs and the rf gain and rf tune resistors. A detected 0.1 volt audio signal should appear at the output, which represents a maximum gain of 90.9 dB. The slug settings on Q1, Q2, and Q3 may be adjusted to optimize the tracking characteristics. It won't be necessary to use the full voltage range control from the 10k rf tuning pot since only a few volts are necessary for maximum gain in the rf stages. Peak gain for the mosfet is achieved between 4 to 10 V. Where fixed gain of the i-f/detector is desired, pins 2 and 5 may be tied directly to the +12 volt lines through 100k current-limiting resistors.

By experimenting with the 100k agc-threshold setting, you'll find that a 4 to 5 volt setting on U1 pin 7 will provide a 50% decrease in gain at 300 μV input. You might wish to vary this setting to suit your specific needs. For strictly CW applications this option is not necessary and the circuit shown in **fig. 6** will be easiest to implement. For maximum operation flexibility, the assembly should be operated as shown in **fig. 7.** This arrangement provides agc control at both the rf and i-f stages and, additionally, allows a variable control in the rf and i-f gain to achieve the most ideal signal-to-noise margin in receiving both CW and ssb signals. Referring to **fig. 7,** with the trimpots shown on the i-f gain, a setting range for both the rf and i-f gain control can be achieved that will allow the pots to be ganged, which is also true of the agc switches shown in **figs. 5 and 7.**

With both crystal and ceramic i-f bandpass filters, the ideal ssb bfo frequency is at the plus and minus 20-dB points from center frequency. CW reception is most easily distinguishable at 800 to 1000 Hz. The crystal oscillating frequency then is usually determined after obtaining the filter and measuring the center and 20-dB points. For both upper and lower ssb and CW reception, individual bfo circuits can be used following the design suggested in **fig. 4,** having a common output bus and separate 12-volt control lines so that the desired frequency may be selected.

The selection of Q1, Q2, and Q3 is a matter of preference. The Motorola MFE3006, MFE3008, and RCA 40673 work well in the circuit shown. Recently, RCA announced an economy device, the 40841. Comparison tests between the 40841 and MFE3006 indicate that the RCA device is a comparable unit and could become the receiver builder's most versatile transistor. It may be used as a single-gate device by tying the gates together for most fet oscillator and amplifier applications, or as a general-purpose unit operating to 50 MHz.

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Frequency coverage

1.8 MHz Band	1.8 - 2.0 MHz
3.5 MHz Band	3.5 - 4.0 MHz
7.0 MHz Band	7.0 - 7.5 MHz
14 MHz Band	14.0 - 14.5 MHz
21 MHz Band	21.0 - 21.5 MHz
28 MHz Band A	28.0 - 28.5 MHz
28 MHz Band B	28.5 - 29.0 MHz
28 MHz Band C	29.0 - 29.5 MHz
28 MHz Band D	29.5 - 30.0 MHz
WWV	RX only 10.0 MHz

MIC. input impedance
50KΩ

Frequency
300 - 2700Hz (−6dB)

Sensitivity
SSB less than .25 μv for 10 dB S/N+N ratio
CW less than .15 μV for 10 dB S/N+N ratio

Modes of operation
SSB (LSB or USB)
CW

Input power
200W
1.8-21.5MHz
100W
28-30MHz

ANT. impedance
50Ω - 75Ω Unbalanced

Carrier suppression
More than 50dB

Side band suppression
More than 50dB

Spurious and harmonic suppression
Greater than 40dB

3rd order distortion products suppression
Greater than 30dB

IF Frequencies
1st IF 9MHz
2nd IF 50kHz

Selectivity
SSB 2.4kHz (−6dB)
4.0kHz (−66dB)
CW 400Hz (−6dB)
1.8kHz (−66dB)

Audio output into 8Ω load
2.5W (10% distortion)
3.0W (MAX)

Power source
AC 120V 50/60 Hz (can be re-wired for 240V)

Power drain
400VA TX
78VA RX
48VA RX (Power tube OFF)

Semi-conductors

Transistor	98
IC	43
Diode	120
Tube	3
Digital Ind.	1

Weight
44 lbs. 6 ozs. (23kg)

Dimensions
16-3/4" x 7" x 13-5/8" (420 x 172 x 340mm)

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an introduction to slow-to-fast-scan television converters

Digital scan converters are a new innovation in slow-scan television — here's how they work

One aspect of amateur radio that appears to be leading all others in technical advancement is slow-scan television. Unquestionably, the accelerated advancements in this field are due to the application of digital techniques to circuit design. Most digitized sstv units are also available on PC boards. This means that an enthusiatic amateur can build highly sophisticated sstv circuits without a full knowledge of how they operate. With these thoughts in mind, this article is presented as a guide to understanding one of sstv's latest innovations: the slow-to-fast-scan converter. I'd like to stress that this is an explanatory rather than a construction article. Amateurs can build kits; I'm attempting to describe how their circuits operate – in this case, the slow-to-fast-scan converter.

If you haven't kept abreast of sstv evolution, here's a thought worth considering: begin with a technical starting point (such as this article), expand from there, then follow subsequent advancements. This beats following behind new innovations because you're not familiar with in-between areas.

converter memories

Dynamic mos shift-register memories are the heart of all scan converters. The typical converter memory has about 16,000 digital words by four-bit plane capacity. This means that about 64,000 bits can be stored in such a memory. The four-bit planes usually store from binary 0000 to 1111, or 16 shades of gray. A simplified example of this type memory is shown in **fig. 1**. The top left corner of this hypothetical unit is shown storing a 5. Its binary equivalent is written into the corresponding four-bit planes. Similarly, the binary equivalent of 8 is written into its corresponding planes, and zeros are written into the four right-hand planes. The number of digital words showing on the memory's front area indicate video resolution capability, while the number of bit planes determine gray levels.

If you can visualize 1800 memory cubes as in **fig. 1** wired in series, you'll have a good idea of the memory size used in a slow-to-fast-scan converter. Also, gray coding would be used rather than straight binary-coded decimal coding. This gray code is easier to use because only one bit changes state between successive digital words. Now let's consider the overall concept of scan conversion.

slow-to-fast-scan converter

This converter represents a revolution in sstv technology because it allows you to view slow-scan television pictures on a conventional (fast scan) home television set. Although there are presently only two basic designs of this unit,[1,2] their operational concepts follow a definite pattern. A typical slow-to-fast-scan converter is shown in **fig. 2**. Incoming slow-scan television signals are fed to an sstv demodulator and sync separator. These circuits, shown in the dotted lines of **fig. 2**, represent a typical P7 sstv monitor without a cathode-ray tube and high-voltage supply. In fact, a conventional sstv monitor could be used for this part of the converter if desired.

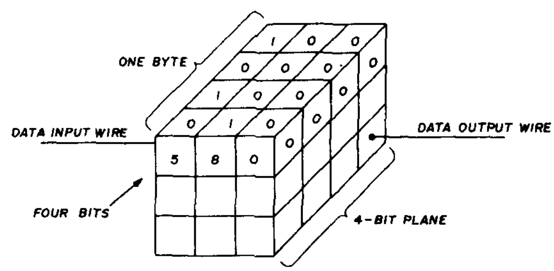

fig. 1. Three-dimensional representation of a shift-register memory used in a slow-to-fast-scan television converter. In this example the total memory capacity is 9 bytes or 36 bits.

Varying video voltages from the demodulator enter the analog-to-digital converter, which changes these voltages to their binary equivalents. This A/D converter employs a voltage divider string connected to voltage comparators to produce digitized equivalents proportional to the varying video levels. This conversion is accomplished periodically by sampling each line of slow-scan video, as illustrated in **fig. 3**. The sampling rate is usually 256 times per line. This sampling rate results in 256 words, each containing four bits of gray-code information. Immediately following each line of sstv video, the sync separator extracts a slow-scan sync pulse, which

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triggers the slow-speed clock (fig. 2). This clock, in turn, opens the proper line buffer input to allow each digitized line to enter the 65,000-bit shift register memory. (Meanwhile, the fast-speed clock directs "read from memory" and "recirculate fast scan information" functions. It does this by sending high-speed shift pulses to the memory and timed, binary-1-level signals to the D/A converter).

this operation for, say, 20 times, you'll see that the red cars will again be on the track. This rather poor analogy is similar to scan-converter memory functions.

concluding remarks

This information has been presented with the hope of enlightening interested amateurs on the basic concepts of digital slow-to-fast-scan conversion. Remember that phenomenal progress is being made in slow-scan television technology. While by no means describing the final word in scan-converter design, I think you'll find this article helpful in following the operation of these circuits.

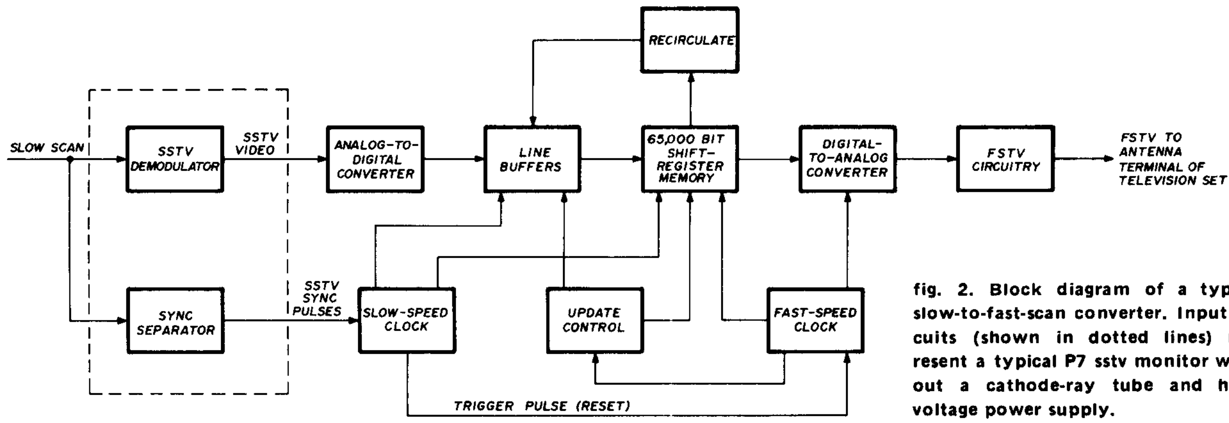

fig. 2. Block diagram of a typical slow-to-fast-scan converter. Input circuits (shown in dotted lines) represent a typical P7 sstv monitor without a cathode-ray tube and high-voltage power supply.

Now, at the precise start of the proper fast-scan line, the fast-speed clock sends a pulse to the update control. This control then directs a binary-1-level signal to the appropriate buffer output, which allows a full line of slow-scan information to enter the fast-speed 65,000-bit memory. This controlled loading operation is performed approximately 128 times during an eight-second period to load a complete sstv picture into memory.

The shift register memory continuously operates at a fast-scan rate, so this information is output (at the proper times) into the D/A converter, where it's converted to fast-scan video. Every 1/15750 second the fast-speed clock sends a sync level pulse to the D/A converter to produce sync-level signals. The D/A converter uses a resistor-transistor network arranged in a *ladder adder* configuration to produce voltage level equivalents to the binary weight of incoming counts.

The final step involves modulating a simple uhf signal generator with the composite fast-scan television signal. This signal is applied to the antenna terminals of a conventional television receiver. Now the resultant slow-scan television picture can be viewed in a well illuminated room.

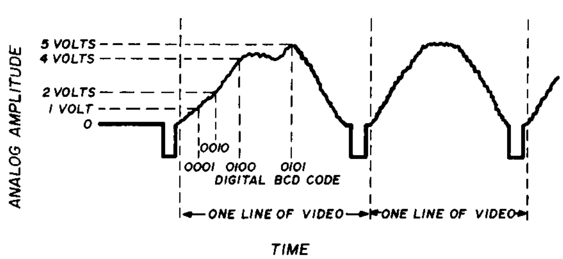

fig. 3. Each line of slow-scan video voltage is sampled at a periodic rate (approximately 256 times per line), which results in 256 digital words per line, each containing four bits of gray-code information.

memory analogy

Memory operation could be illustrated by using an automobile race track to symbolize data flow. First, consider the race track as a circle filled to capacity with very fast blue cars. Fifty green cars are on the entry ramp at one end of this circle, and fifty red cars are on the exit ramp at the opposite end of the circle. A horn sounds and ten red cars quickly leave the exit ramp. At the same time, ten blue cars quickly enter the exit ramp, and ten green cars slowly enter the race track (but quickly gain speed). The red cars then drive around the track and wait in line with the green cars. If we follow

Other sstv innovations are being developed, such as the color system that uses a 500-Hz subcarrier to carry I and Q color information, while black and white information is contained in the usual 1200 to 2300 Hz range. You'll probably be hearing more on slow-scan television developments in the near future.

I'd like to thank Dr. Robert Suding, WØLMD, for his criticism on the final version of this article. If reader interest warrants, future articles on digital electronics or slow-scan television may be presented.

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ham radio

a revealing analysis of the coaxial dipole antenna

This analysis of the coaxial dipole antenna shatters many of the myths — it also explains why its swr bandwidth doesn't measure up to some of the claims

At some time or another most radio amateurs have used a folded-dipole antenna. And, wisely or not, more and more amateurs are now trying the *coaxial* dipole, sometimes called the "double bazooka."[1] Those amateurs who favor the folded dipole generally do so because of its flexible input-impedance characteristics; many amateurs are also acquainted with its somewhat broader bandwidth characteristics as compared with the single-conductor dipole. Many amateurs also try coaxial dipoles in hopes of obtaining increased bandwidth.

Some of the proponents of the coaxial dipole apparently use it because they have been misled into believing it also possesses certain other exotic characteristics, such as supplying its own balun when fed with a coaxial transmission line, as if providing an inherent unbalanced-to-balanced input (not so); exhibiting more than 1.5 dB gain over a simple dipole[2] (not so); and even offering a noise-cancelling capability because the antenna is said to be "shielded" (some amateurs have claimed that "the coaxial dipole has no man-made noise pickup whatever").

One purpose of this article is to show why the coaxial dipole cannot perform according to these utopian specifications and that, except for restricted differences in bandwidth and impedance, its performance is the *same* as that of a simple dipole. Although the coaxial dipole can provide a limited broadening of the impedance bandwidth as compared to a simple dipole *if* engineered correctly, the coaxial dipole configuration in general use by amateurs is not engineered correctly. Consequently, it does not provide the broadband operation erroneously attributed to it.

A second reason for this article is to discuss the *reasons* why the engineering requirements for broadband operation are *not* fulfilled in the amateur configuration, thereby alerting the eager but unsuspecting builder before he wastes valuable time and expensive coax in building a complex dipole that will perform no better than a simple dipole.

impedance bandwidth

The impedance bandwidth of an antenna is derived from the impedance mismatch between the antenna and its feedline as the operating frequency is varied between specified limits. The principal contributor to increased mismatch, as the operating frequency departs from the self-resonant frequency of the antenna, is the reactance which appears in the input impedance of the antenna. The effect of reactance and other parameters is shown in **appendix 1**. This reactance is developed whenever current reflected from the ends of the radiating elements arrives back at the input terminals with other than a 0° or 180° phase relative to the incoming current from the feedline. A zero or 180 degree relation is obtained only when the operating frequency is the same as the self-resonant frequency. This is why the resonant frequency of an antenna is sensitive to the length of the radiating element.[3]

In an *ideal* antenna having infinite bandwidth all the power delivered by the feedline would be radiated by the time the outward-flowing current reaches the end of the radiator, so no reflected current would return to generate a reactance at the input terminals. In other words, the ideal antenna would simply be a broadband transformer, matching the feedline impedance to the 377-ohm intrinsic impedance of free space at *all* frequencies. In our quest for increasing dipole bandwidth we are looking for some scheme which will either cancel or compensate this reactance as it appears (as in folded and

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coaxial dipoles), or reduce the reactance by reducing the amount of out-of-phase current reflected from the ends of the radiator. The out-of-phase currents arriving from the ends of a conically-shaped radiator (or a fan-shaped multi-wire dipole) are smaller than those in a thin dipole so less off-resonance reactance is generated in wider-ended radiators.

The reactance compensation obtained in folded and coaxial dipoles results from applying reactance of the opposite kind, contributed by shorted stub sections, as shown in **fig. 1B** for the folded dipole, and in **figs. 1A, 1C** and **1E** for the coaxial configuration. Each half of the folded dipole and the internal portion of each coaxial section in the coaxial dipole forms a resonant quarter-wavelength stub near the resonant frequency of the dipole, short-circuited at the end opposite the feed point. The two stubs of the coaxial dipole are connected in series through their inner conductors. The series combination is shunted across the dipole input terminals as shown in **fig. 1E**. The shunt-connected stub combination reduces the off-resonance reactance appearing at the dipole input terminals because the stub reactance is inductive below resonance, while the dipole impedance is capacitive, and vice versa. Thus, the off-resonance mismatch to the feedline is reduced.

However, the analysis which follows reveals some facts which will probably come as a distinct surprise to many amateurs, and may cause those who are using the coaxial dipole to contemplate replacing it. The facts are, one, whatever increase in bandwidth is obtained solely by the two coaxial stub sections inside the dipole, the *same* bandwidth can be obtained by using a simple wire dipole of the same outside diameter, but having an *external* shunt stub (equivalent to the internal stubs) connected *directly* across the dipole input terminals as shown in **fig. 1D**. Stuffing the stubs inside the radiator does nothing except to provide a convenient place to hide them. This is because the stub currents flowing on the inner side of the coaxial outer conductor are completely separate from the antenna currents flowing on the outside, and the outside antenna currents are *unaffected* whether the conductor is the outer portion of a coaxial cable, or simply a solid conductor.

Secondly, the amount of bandwidth improvement actually obtainable using the shunt-stub technique depends directly on the relationship between the values of the conductance term of the dipole admittance and the characteristic impedances of both the feedline and the shunt-stub lines. This relationship, which will be explained, involves conversion between equivalent series and parallel circuits. It limits feedlines to those having impedance values, Z_c, within a range whose lower limit is well *above* those commonly used (50 or 75 ohms). The stub lines must have Z_c values in the range from five to ten ohms, which are practically unattainable.

These requirements warrant an explanation which will follow shortly. But first, for the reader who is using a coaxial dipole fed with a fifty-ohm transmission line, how about trying an experiment which will prove that the stubs don't provide the heralded broadbanding? Measure and record the vswr at regular frequency intervals across the entire band. Then open the center conductor of the coaxial sections between the two dipole halves at point A in **fig. 1E**, and replace the antenna to the original height. Now remeasure the swr at the same frequency points and prepare for a shock. I predict that you will find an insignificant difference between the two sets of swr readings.

control of mismatch by R and X

To utilize frequencies in any part of a band, antennas are operated off resonance (except at one frequency).

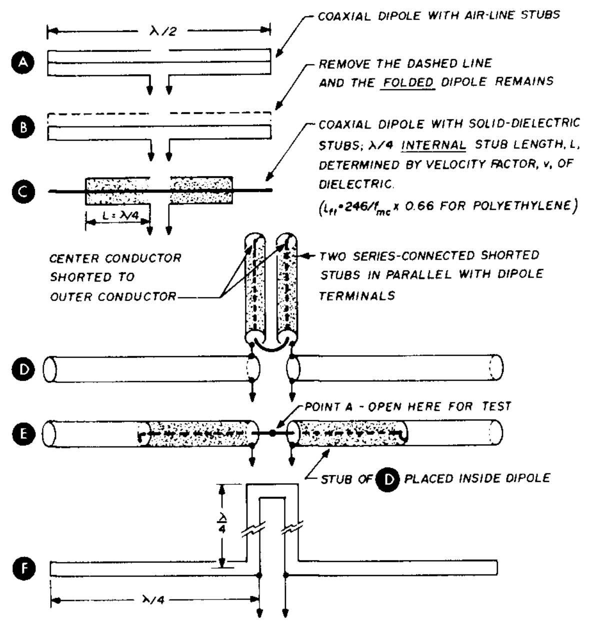


fig. 1. Reactance compensation in coaxial and folded dipoles results from reactance contributed by shorted stub sections. Coaxial dipole with air-line stubs at (A) reduces to a folded dipole (B). Each half of the folded dipole, and the internal portion of each coaxial section in the coaxial dipole (C), form a resonant quarter-wavelength stub. The same bandwidth can be obtained with a simple dipole which has the same outside diameter by using an external shunt stub across the feedpoint (D). Placing the stubs inside the dipole at (E) does not change bandwidth performance. The equivalent configuration is shown at (F).

When operated off resonance we know that the dipole antenna impedance, Z_d, contains both resistance, R, and reactance, X. To obtain zero mismatch at resonance, we know that the line impedance, Z_c, must equal the antenna load impedance, $Z_d = (R + jX)$, which is $R + j0$. To obtain minimum mismatch (which can't be zero) off resonance, the line impedance, Z_c, must equal the absolute value, or magnitude of the load impedance, $|Z_d|$, i.e.

$$Z_c = |Z_d| = \sqrt{R^2 + X^2}$$

A feedline having an impedance which will vary the same as $|Z_d|$ over the desired frequency range does not exist. Therefore a compromise must be found which will now be discussed.

As seen in **table 1** of **appendix 1**, for a value of line impedance which equals load impedance, $|Z_d|$, the mismatch is smaller when X is low and R is high; mismatch is zero for a load impedance of $R + j0$. We know that the magnitude of the dipole impedance, $|Z_d|$ rises *above* its resonant impedance on *both sides* of resonance because of the off-resonant reactance component which appears in the dipole impedance (**fig. 2**). I will show later in the analysis that, when applying the stub-compensation technique, the reactance of the stub shunting the dipole *tends* to cancel the off-resonant dipole reactance.

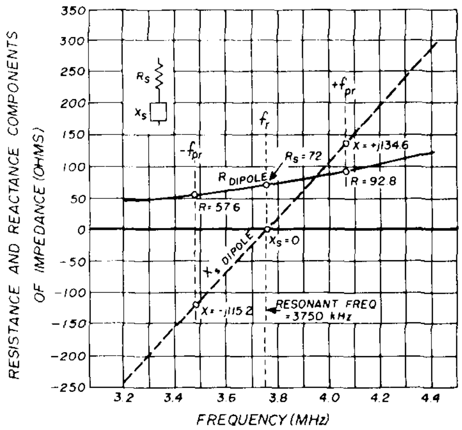


fig. 2. The series resistance, R_s and reactance, X_s components of *uncompensated* dipole impedance versus frequency. Values shown here are for a half-wavelength thin-wire dipole in free space.[11]

In our stubbing operations the following relationship between equivalent series and parallel circuits is of great importance.

When a series circuit containing both resistance and reactance is shunted with a reactance of the opposite kind, the absolute value of the circuit impedance $|Z|$ is increased. However, in applying the technique of shunt-stubbing to thin half-wavelength dipoles, the transformation between series and parallel circuits appears to have been ignored in amateur literature and applications. The impact of ignoring this important relationship is evident when you see that, while the shunt reactance of the stub *is cancelling* the dipole reactance, the stub reactance simultaneously raises the series-circuit resistance in the *resulting parallel circuit.*

Therefore, even though the effective reactance in the load impedance is lowered by the stub reactance, the resulting increase in resistance increases the magnitude, $|Z|$, of the load to values which are even higher than those of the off-resonant impedance of the dipole alone. Thus you have simply exchanged a mismatch caused by reactance for a similar mismatch caused by the increased resistance.* These relationships are illustrated in the following example:

Consider an 80-meter dipole antenna at a height yielding a resonant impedance, Z_d, of $55 + j0$ ohms at 3750 kHz. At 200 kHz below resonance (3550 kHz) the impedance Z_d is $50 - j90$ ohms. Here $|Z_d| = 103$ ohms, and the mismatch is 5.04:1 on a 50-ohm line (see **appendix 2** for calculation method). Since stubbing is a shunting operation, we now examine the equivalent parallel circuit. The component values of the equivalent parallel circuit are $R_p = 212$ and $X_p = -117.8$ ohms. If the capacitive 117.8 ohms were entirely cancelled, the new load impedance would be $212 + j0$ ohms, and the mismatch would be reduced only to 212/50 = 4.24:1.

Note that the poor swr improvement is because the resistance has been raised from 50 to 212 ohms. However, the situation is even worse if the reactance compensation is performed with RG-58/U stubs generally found in amateur coaxial dipoles. Two 50-ohm RG-58/U stubs in series, resonant at 3750 kHz, yield an inductive shunting reactance of 1190.9 ohms at 3550 kHz. If they are shunted across the above parallel circuit of $R_p = 212$ and $X_p = -117.8$ ohms, the resulting values are $R_p = 212$ and $X_p = -130.7$ ohms. The equivalent series impedance is now $58.4 - j94.7$ ohms, $|Z_d| = 111.3$ ohms, and the mismatch is 4.9:1. Obviously neither of these two mismatch reductions is worthwhile.

However, a stub can provide *broadbanding if it is* properly engineered. The approach is clarified if we recall how mismatch is produced by a constant value of $|Z_d|$ when its components X and R vary if line impedance, Z_c, is nearly identical to dipole impedance, $|Z_d|$: low mismatch when X is low and R is high, as shown in **appendix 1, table 1**. However, in the example above the dipole impedance $Z_d = 55 + j0$, and the line impedance, $Z_c = 50$ ohms, are nearly identical *at resonance*. Thus the mismatch is lowest at resonance: $Z_d/Z_c = 55/50 = 1.1:1$. But as we depart from resonance the minimum obtainable mismatch increases for either of two reasons: The uncompensated dipole impedance, $|Z_d|$ becomes increasingly higher than the line impedance, Z_c, or the parallel-circuit resistance, R_p, becomes increasingly higher than the line impedance. These reasons explain why, in the example, so little mismatch reduction is obtained with stubbing, even with the reactance entirely cancelled. In other words, if there is a substantial difference between the line impedance and the absolute magnitude of the load impedance, it makes little practical difference in the amount of the resulting mismatch whether the load is predominantly resistive or reactive.

On the other hand, as will be shown later, you can reduce the mismatch (using stubs) over a limited frequency range by choosing a line impedance, Z_c, intermediate between the extreme values of the compensated dipole impedance encountered over the frequency range of interest. Since the magnitude of the complex

*A typical article contributing to the misunderstanding on this point, by failing to appreciate the fundamental reactions resulting from *parallel-connected* circuit elements, may be found in *73 Magazine*, June, 1973, page 80 (John Schultz, W2EEY, "The Double-Coaxial Antenna"). The author's statement that a 50-ohm feedline must be used to feed coaxial-dipole antennas further illustrates his lack of appreciation for the actual principles involved.

dipole impedance rises off resonance (and is raised still further by the shunt compensation), the line impedance required to reduce off-resonance mismatch must be higher than that which yields the best match at resonance. Thus we must accept a compromise in the match at resonance in exchange for an improvement in match at frequencies off resonance.

Because our control over the conductance values for high-frequency dipoles of any practical length and length-to-diameter ratio is limited by nature, the relationships just described yield the following important facts concerning the feeding of coaxial-dipole antennas: A useful increase in bandwidth over a simple dipole can be obtained with a coaxially-stubbed *monopole* over a ground plane fed with a 50-ohm feedline because the antenna impedance at resonance is *less* than the feedline impedance. A similar bandwidth increase results when a thin, balanced-feed, half-wavelength coaxial dipole is fed with a feedline having a Z_c of 100 ohms or more. However, *no* significant increase in bandwidth can be obtained using the shunt-stub technique when a balanced, thin, half-wavelength coaxial dipole is fed with a 50-ohm transmission line, because the antenna impedance at resonance is usually near to, or greater than, the 50-ohm line impedance.

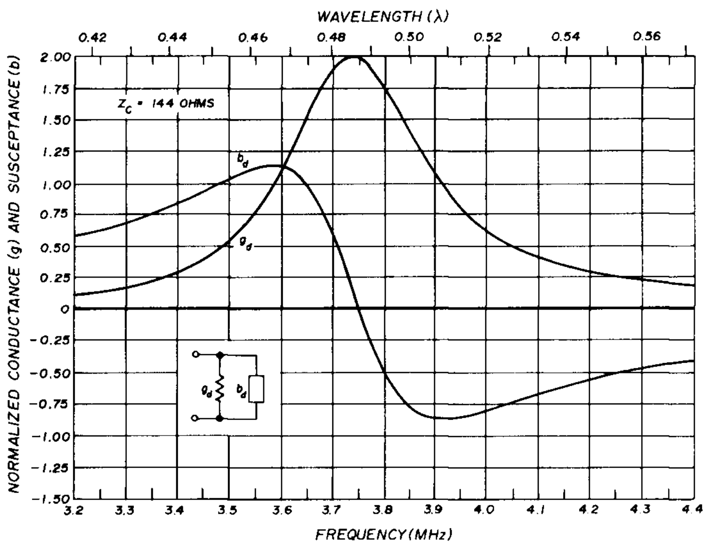


fig. 3. For ease of calculation, the impedance values of fig. 2 have been converted to admittance and plotted in terms of normalized conductance, g, and susceptance, b. Feedline impedance of 144 ohms yields optimum bandwidth, as discussed in the text.

A further requirement for obtaining *significant* bandwidth improvement when using a feedline of suitable impedance is that the shunting-stub reactance drops appropriately with deviation from anti-resonance to compensate the antenna reactance as it increases with deviation from resonance. This requires a stub with a low value of characteristic impedance which is difficult to attain. To satisfy this requirement, the analysis will show that the impedance of the stub-line must be in the range from 10 to 20 ohms if a single external stub is used, or from 5 to 10 ohms each for two stubs *in series*. In typical amateur coaxial-dipole configurations series-connected stubs of RG-8/U or RG-58/U present such high reactance, they are incapable of providing any significant compensation, even if the feedline impedance is of the proper value to obtain *optimum* improvement. Thus two fundamental parameters – low feedline impedance and high stub impedance – cause the typical amateur version of the coaxial dipole to fall short of its goal. The ineffectiveness of these parameters is illustrated in the 80-meter example discussed earlier.

Although this entire situation may seem a bit incredible, the facts are supported explicitly in the published works of Kraus (W8JK),[4] Everitt,[5] Jordan,[6] Coleman,[7] Borton (W9VMQ),[8] and others, in addition to my own experiments and the analysis which follows.* In view of this, how do we account for the apparent success of the coaxial dipole as claimed by so many happy users? There are probably several reasons. First, how many of those happy users performed the test described earlier to obtain a true comparison between the same dipole with and without the coaxial feature? Probably very few, if any. If not, how many amateurs know the *absolute* accuracy of their swr indicators in the *actual* 4:1 or 5:1 swr range? Many indicators give erroneous readings in this range, far *lower* than the true value, resulting in an indication of a wider bandwidth than actually exists.

Secondly, the apparent bandwidth improvement claimed by those using the coaxial dipole with stubs built from 50-ohm cable and fed with a 50-ohm feedline includes the cumulative effects of additional phenomena which are overlooked without realizing that an insignificant amount of improvement is actually contributed by the coaxial configuration itself.

If we assume the swr values presented by Charles Whysall are accurate, they seem to show acceptable results obtained with his Double-Bazooka antenna.[1] However, his design includes *two* broadbanding features which contribute simultaneously – the coax stubs plus capacitive loading of the dipole end sections using two-wire ladder line extending beyond the coaxial portion to obtain a full half-wavelength radiator. This type of multi-wire construction increases dipole capacitance and decreases inductance in the same way as increasing the radiator diameter. It can be shown that this type of loading provides a greater contribution to the increased bandwidth than the coaxial feature, yet Whysall makes no mention of any analysis or experiment performed to determine the *amount of bandwidth* contributed separately by each feature; he simply states that the effective reduction of length-to-diameter ratio provided by the ladderline contributes to a lowered radiator Q.

This is not intended as a criticism of his method of obtaining more bandwidth, or of his article, which con-

*My analysis and experiments were reviewed by *QST's* Associate Technical Editor, Gerald Hall, K1PLP, who performed a verifying experiment at the ARRL Lab. Result: The coaxial-dipole antenna has not been included in the *ARRL Antenna Book.*

tains valuable and pertinent information; it simply isn't correct to attribute the combined effect of *both* features to the coaxial feature alone because it's misleading to the uninitiated.

Since the mechanical construction of the coaxial dipole is not simple, another purpose of this article is to help you to understand what is actually happening so you can reorient your design approach toward a construction method which will yield increased bandwidth performance. Antennas which meet this requirement include the multi-wire, fan-shaped, bow-tie dipole mentioned earlier, invented by Philip S. Carter of RCA in 1937 to obtain the bandwidth necessary for television.[9] Since the beginning of television broadcasting millions of TV receiving antennas have used the Carter bow-tie configuration, and in 1955 it was suggested for 80-meter use by Camillo and Purinton,[10] and again in 1975 by Borton.[8] Additional information on the fan dipole has also been published by Meier.[7]

coaxial dipole principles

In this section we will examine broadbanding principles using stubs under four different conditions. The first two will provide useful bandwidth improvement, while the third and fourth illustrations will provide negligible improvement.

For the analysis to be realistic it will be centered around a half-wavelength dipole which has the physical characteristics of a typical amateur coaxial-dipole antenna, as shown in **figs. 1C** and **1E**. Since the characteristic impedances of both the feedline and the shunt-stub line are critical factors in controlling bandwidth, I will first show how to determine the value of feedline impedance which will yield the maximum possible dipole bandwidth with a given mismatch, or swr, and the impedance required of the shunt-stub line. Next, for comparison, I will determine the bandwidths obtainable (and the stub-line impedances required) when using feedlines of 100 and 50 ohms, respectively. Finally I will show graphically how to assess the bandwidth performance with 50-ohm feedline and 50-ohm stubs.

Since the parallel-circuit relation of the two stubs with respect to the dipole terminals is sometimes difficult to perceive, I will also show why the shorted stub lines are actually in shunt with the input terminals in *both* types of dipoles. Since I will be discussing circuit elements connected in shunt, or parallel, it is convenient to use the less familiar terms of conductance G, and susceptance, B, (components of admittance, Y,) in addition to the somewhat more familiar resistance, R, and reactance, X, components of impedance, Z. Since conductance and susceptance are the reciprocals of *parallel-circuit* resistance and reactance, respectively, they permit the use of direct algebraic addition of the component values associated with each of the parallel-connected elements to determine the total value of the combination. Handling susceptances in this way simplifies the understanding of reactance compensation, especially in the graphical representation in the illustrations.

As an aid to understanding the relationship between equivalent series and parallel circuits, if the series components of impedance are represented by R_s and X_s, and the parallel components by R_p and X_p, then

$$R_p = \frac{R_s^2 + X_s^2}{R_s} \qquad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

while $$G = \frac{R_s}{R_s^2 + X_s^2} \qquad B = \frac{-X_s}{R_s^2 + X_s^2}$$

The series resistance and reactance components of the *free-space*, uncompensated, thin-wire dipole impedance versus frequency are plotted in **fig. 2**. This graph shows an impedance of 72 + j0 ohms[11] at resonance.* However, since the effect of elements added in *parallel* with the antenna is best shown on an admittance diagram, the impedance values have been converted to admittance and replotted in **fig. 3** in terms of *normalized* conductance, g, and susceptance, b. The normalizing technique will be explained presently. The fundamental relationship between dipole conductance, feedline impedance, and bandwidth will now be described for conditions which yield the maximum possible bandwidth.

As a point of departure, we decide on a maximum practical value of mismatch, or swr, which can be tolerated over the band of interest. We will identify this value by either "swr limit," or "mismatch limit." We will define the band of interest, or bandwidth, as the difference between the frequency extremes where the mismatch limit is reached. As is well known, maximum bandwidth is obtained in the conventional, uncompensated, coax-fed dipole when the feedline impedance, Z_c, closely matches the antenna impedance, Z_d, at resonance. However, as I pointed out earlier, to increase bandwidth we must accept a mismatch at resonance in exchange. In fact, I will show that for any swr limit we may select, we obtain maximum bandwidth by deliberately causing the mismatch to attain the selected limit at resonance. We cause the mismatch to reach the limit at resonance by choosing the line impedance, Z_c, higher than the dipole resonant impedance, $Z_d = R_d + j0$, by the ratio equal to the desired mismatch limit. Thus to obtain maximum bandwidth for any swr limit, the line impedance, Z_c, must be $Z_c = R_d \times swr_{limit}$, where R_d is the value of the dipole resistance at resonance. We will use a mismatch limit of 2 for this entire discussion. Thus the proper line impedance, Z_c, for feeding a 72-ohm

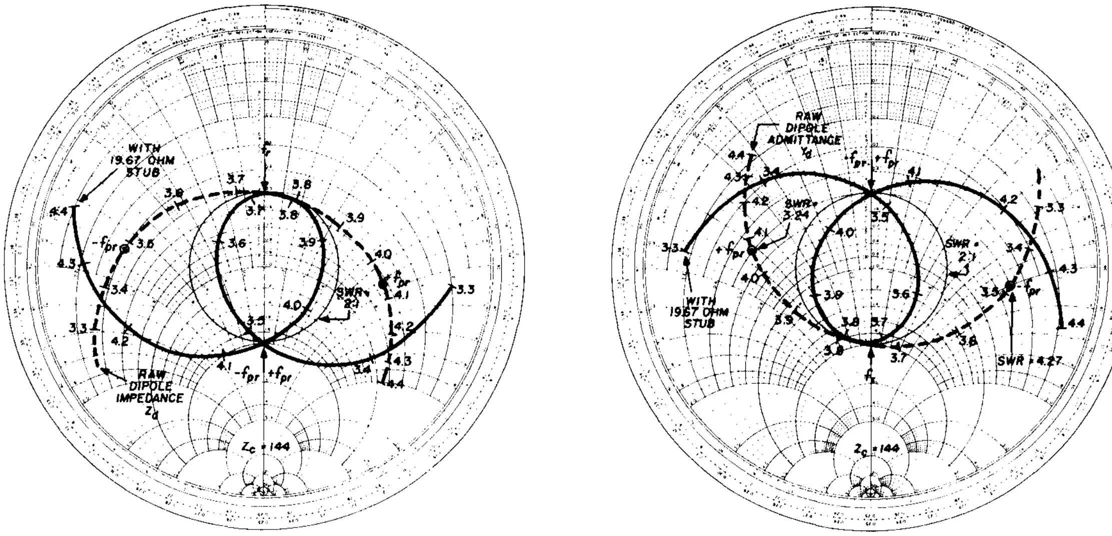
*A free-space dipole was chosen for illustrating the shunt-stub broadbanding principle because its resistance and reactance components of impedance are precisely known. It also avoids complicating the presentation with the effects of mutual coupling with the ground-reflected dipole image at different heights above the ground. Impedances and bandwidths obtained with actual earth-oriented dipoles will therefore differ slightly from the values presented here. The bandwidth of an uncompensated 80-meter dipole at typical heights of 0.25λ, or less, with a 50-ohm feedline will be slightly wider than that of the free-space dipole because the resistance of the earth-oriented dipole at resonance is reduced from 72 ohms to some value closer to 50 ohms, due to the mutual coupling with its image. However, the percentage bandwidth *improvement* obtained when using shunt stubs with *earth-oriented dipoles will not differ significantly from that* presented herein using the free-space dipole data.

stub-compensated dipole to obtain maximum bandwidth over 2:1 swr limits is $Z_c = 72 \times 2 = 144$ *ohms.*

I will now explain the normalizing procedure and conversion between resistance and conductance, which we will be using in the analysis which follows. An impedance, Z, is normalized by dividing it by the line impedance Z_c, to which it is being referenced; it is indicated by the lower case, z. In our example the normalized resonant-dipole impedance $z_d = r_d + j0 = R_d/Z_c = 72/144 = 0.5$. Conductance, G, is the reciprocal

similarly the minimum dipole impedance, z_d, at resonance equals 0.5, as seen in **fig. 4A.**

Since maximum g_d occurs at resonance this establishes the center frequency. The line impedance chosen establishes the band edges of the mismatch limit at the frequencies on either side of resonance where g_d equals the reciprocal of the mismatch limit (where $g_d = 0.5$). Application of this rule in the analysis will be presented shortly. However, it can be seen in **fig. 4** that this relationship between Z_c and g_d is chosen for the com-



A IMPEDANCE

B ADMITTANCE

fig. 4. Smith chart plots of normalized impedance (A) and normalized admittance (B) from 3.3 to 4.4 MHz for both uncompensated and stub-compensated dipoles, based on the use of a 144-ohm feedline. Selection of 19.7 ohm stubline impedance is discussed in text.

of resistance, R, so $G_d = 1/R_d = 1/72 = 0.0139$ *mhos (13.9 millimhos).* Normalized conductance, $g_d = Z_c \times 1/R = Z_c \times G$. Thus the normalized resonant dipole conductance, $g_d = 144 \times 1/72 = 2.0$.

These relations are sometimes best appreciated when displayed on Smith charts, because, in general, plots appearing on a Smith chart represent the normalized values of impedance or admittance. **Figs. 4A** and **4B** show Smith chart plots of the normalized dipole impedance, z_d, and admittance, y_d, respectively; they show conditions both with and without stub compensation. Because of their reciprocal relation these impedance and admittance plots are seen to be mutually inverted. The circle marked "swr = 2:1" encloses all impedances (or admittances) which give rise to mismatch values less than the selected 2:1 limit. These plots will be explained later on.

Since our analysis will be using admittance parameters, the following rule is convenient for determining maximum bandwidth obtained with stub matching in relation to dipole conductance, line impedance, and mismatch limit: The optimum line impedance, Z_c, must be chosen (as explained above) so that the normalized maximum dipole-conductance value, g_d, is equal to the mismatch limit. The maximum conductance, g_d, equals 2 at the dipole-resonant frequency, f_r, as seen in **fig. 4B;**

pensated dipole to yield an identical mismatch value at the center frequency and at the band edges.

Fig. 4 also shows that selecting the line impedance, Z_c, at 144 ohms gives rise to the 2:1 limiting mismatch at resonance, and places the normalized uncompensated dipole-impedance (or admittance) locus on the graph so it intersects the 2:1 swr-limit circle at resonance (f_r). The importance of this line-impedance selection will be more fully appreciated after an explanation of the reactance-cancelling action. However, it can be seen that placing the uncompensated impedance (or admittance) locus on the graph as just described allows the maximum length (frequency range) of the locus to be warped inside the swr-limit circle as the result of stub compensation. Note that although the mismatch reaches the swr limit at both band center and band edges (because the compensated locus passes through the swr-limit circle at these points), the mismatch is less than the limit everywhere between the center and the edges. This is because all impedances represented by the stub-warped locus lying inside the swr-limit circle give rise to mismatches that are less than 2:1.

The importance of selecting proper line impedance is further emphasized in **fig. 10.** Observe the dramatic reduction in frequency range of dipole impedances warped into the swr-limit circle using 100- and 50-ohm lines shown in **fig. 10,** in contrast with the range obtained with the optimum 144-ohm line (**fig. 4**). The impedances plotted in **fig. 10B** illustrate the ineffectiveness of stub compensation when using 50-ohm feedline, with either an optimum stub, or 50-ohm RG-58/U stubs. The frequency range of impedances moved into the swr-limit circle with compensation is nearly the same as with no compensation. Thus with 50-ohm feedline, the only conceivable purpose a stub can serve is to enlarge the antenna profile as an announcement to the neighbors, "here is a ham, a potential source of RFI."

This early comparison of the plots in **figs. 4** and **10** was simply a preliminary view of the results obtained by examining the four broadbanding conditions using stub-reactance compensation. So we will now return to the explanation of how these results were obtained. It is the mismatch value to the 2:1 limit. The normalized compensated-dipole resistance, r_d, at the band edges is 2.0, as previously stated, and shown in **fig. 4A.** Thus the real resistance at the band edges equals $r_d \times Z_c = 2 \times 144 = 288$ *ohms.* This reemphasizes that the resistances 72 ohms (at the center frequency) and 288 ohms (at the band edges) yield the selected 2:1 swr limit when they terminate the optimum 144-ohm feedline. Note that this value of line impedance, which yields the maximum bandwidth with a 2:1 swr, is centered between 72 and 288 ohms; it is thus the geometric mean value between the resonant, band-center dipole resistance and the band-edge circuit resistance. With these values and procedures established, we can delve into the operation of the shunt-stub configuration to see how the reactance compensation is effected.

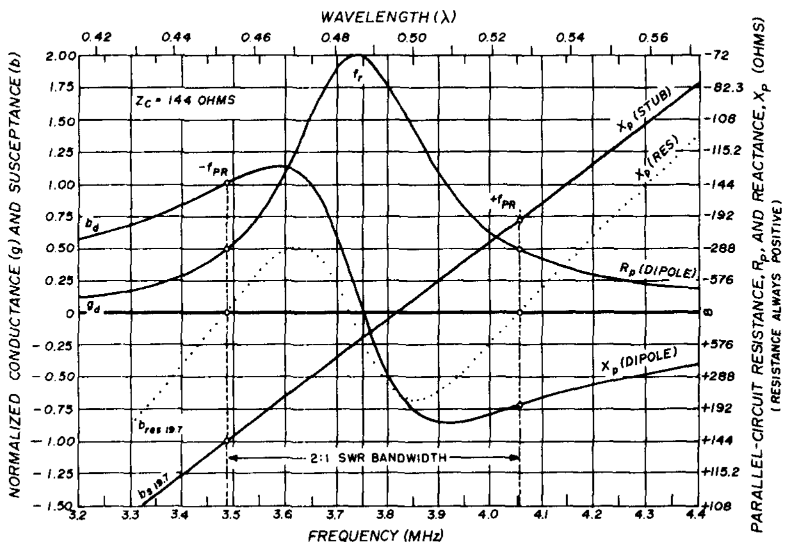


fig. 5. Normalized admittance parameters of a half-wavelength, shunt-compensated dipole. Conductance, g_d and susceptance b_d of the dipole itself are the same as in fig. 3. Since the stub susceptance, b_s, tunes out the dipole susceptance, b_d, at the upper and lower parallel resonant frequencies $-f_{pr}$ and $+f_{pr}$, the resultant susceptance b_{res}, is zero at these frequencies.

convenient to use a graph containing dipole admittance data as a worksheet in determining the stub parameters and the bandwidth obtained through compensation using various feedline and stub combinations. We will use the Smith admittance graph of **fig. 4B** as a worksheet a little later, but we will now use the rectangular-coordinate graph of **fig. 5**, because it affords graphical construction advantages that assist in clarifying the principles of stub matching. **Fig. 5** replots **fig. 3**, the dipole conductance, g_d, and susceptance b_d (normalized to the optimum 144-ohm line), plus other parameters which will be explained as we proceed.

Using **fig. 5**, the band edges of the 2:1 mismatch are *determined from the dipole conductance curve,* g_d, by finding the frequencies on each side of resonance (shown at f_r) where the normalized conductance is 0.5 (the *reciprocal* of the mismatch limit 2.0). We then find a *shunt stub of proper impedance to completely cancel* the dipole susceptance, b_d, simultaneously at both of these frequencies. As seen in both **figs. 4B** and **5**, complete cancellation of dipole susceptance at each band-edge frequency leaves a normalized, pure conductance of 0.5 at both of these frequencies, as required to reduce

shunt stubs

Referring now to **fig. 1F**, you can see the terminals of *the half-wavelength dipole (shortened to resonance) connected* in parallel with a *shorted* quarter-wave stub transmission line. It is well known that the impedance at the input terminals of a shorted quarter-wavelength line of any characteristic impedance, Z_c, is a pure resistance of a very high value. Thus, at the resonant frequency, f_r, the stub connection has a negligible effect on the 72-ohm dipole impedance. However, at frequencies below resonance both the dipole and shorted stub line become *electrically* shorter, and the dipole becomes capacitive, while the stub line becomes inductive. Due to the parallel connection the inductance of the stub line tends to cancel the capacitance of the dipole.

Conversely, at frequencies above resonance the dipole impedance becomes inductive and the stub line becomes capacitive so a similar compensation is again obtained. Unfortunately, the compensation is far from perfect because, although the dipole and stub susceptances are of opposite polarity, they are not equal nor do they change at the same rate. You can see this in **fig. 5** which shows how the dipole susceptance, b_d, varies with frequency in contrast to the stub susceptance shown by the straight line, b_s. The resultant susceptance remaining from the imperfect compensation equals the sum, $b_d + b_s$, shown in the curve b_{res}. This resultant susceptance, b_{res}, combined with the dipole conductance, g_d, will yield the locus of the compensated-dipole admittance in **fig. 4B,** and the corresponding impedance locus in **fig. 4A.**

Still referring to **fig. 5,** if you look far enough above dipole resonance you will find a frequency where the dipole and stub susceptances are equal and opposite, and a perfect susceptance compensation is obtained because this is the higher of the two band-edge frequencies at which the stub line was selected to cancel the dipole susceptance. These equal and opposite susceptance values, $b_d = -0.725$, and $b_s = +0.725$, are seen at points on the ordinate line which intersects the dipole conductance curve at $g = 0.5$, and the frequency scale at $+f_{pr}$ (f_{pr} = *parallel-resonant frequency*). Since the stub susceptance, b_s, tunes out the dipole susceptance, b_d, at this frequency, the resultant susceptance, $b_{res} = 0.0$, and parallel resonance is established. (These values may also be seen in **fig. 4B.**) From network theory we know that when parallel resonance is obtained by cancelling dipole reactance with a shunt-stub reactance of the opposite sign, then the relatively-low value of series dipole resistance is converted to the higher value of its equivalent parallel-circuit resistance component.

Thus the impedance at the antenna terminals is a pure resistance of 288 ohms for an swr of 2.0 as stated previously. **Fig. 2** shows the raw (uncompensated) dipole impedance at this frequency to be $Z_d = 92.8 + j134.6$ *ohms*($|Z_d| = 163.5$ *ohms*). This impedance would yield a 3.24:1 swr on the 144-ohm line in the absence of the stub. However, the *equivalent parallel-circuit* components are $R_p = 288$ and $X_n = +198.6$ *ohms*, but with the susceptance cancelled, leaving the 288 ohms of pure resistance as shown in both **figs. 5** and **6.**

Similarly, a second parallel-resonance frequency, $-f_{pr}$, will be found *below* the dipole resonant frequency where the dipole conductance again equals 0.5. This can

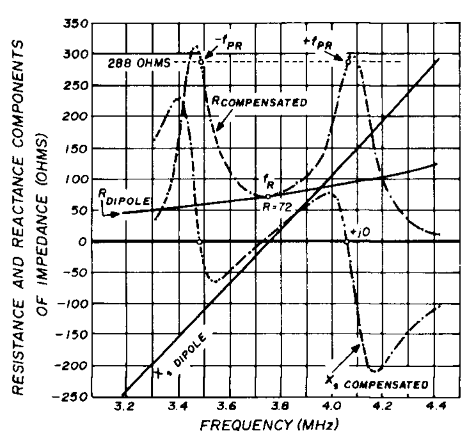


fig. 6. Resistance and reactance components of compensated and uncompensated half-wavelength antennas. Reactance is cancelled at the upper and lower parallel-resonant frequencies $+f_{pr}$ and $-f_{pr}$ leaving a resistive component of 288 ohms.

be seen in **figs. 4B** and **5.** This is the *lower* band-edge frequency at which the stub was selected to cancel the dipole susceptance. The susceptance values at this frequency are $b_s = -1.0$, and $b_d = +1.0$, while the raw dipole impedance shown in **fig. 2** is $Z_d = 57.6 - j115.2$ *ohms*, ($|Z_d| = 128.8$ *ohms*) for an swr of 4.27:1 without the stubs. The equivalent parallel-circuit components are

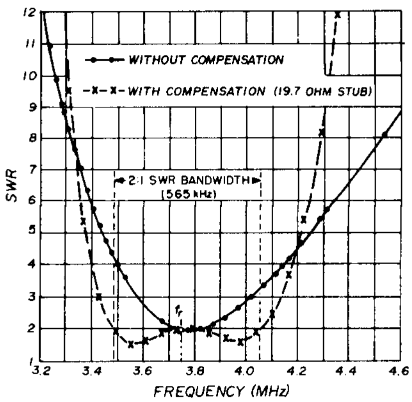


fig. 7. Swr vs frequency for compensated and uncompensated dipoles fed with a 144-ohm feedline. Characteristic impedance of compensating stub is 19.7 ohms.

$R_p = 288$ and $X_p = -144$ ohms, and as at $+f_{pr}$, with susceptances and reactances cancelled, the circuit impedance is again $288 + j0$ ohms for an swr of 2.0:1.

The procedure for calculating the precise value of the shunt-stub line impedance, Z_c, which cancels the dipole susceptance at both band-edge frequencies, from the slope of the stub susceptance line b_s, in **fig. 5,** requires more space than is available here, but sufficient accuracy will be obtained by using the procedure outlined in **appendix 3.** However, the Z_c value for this slope is 19.7 ohms and must be divided by two if two stubs are used in series.

We know the mismatch and impedance values at the band center and at the band edges. So we now want to determine impedance and mismatch in the frequency range *between* the band center and the edges. Susceptance, b_{res}, (and reactance) are present in this frequency range, because the stub provided complete susceptance cancellation only at the band-edge frequencies. Determining impedance and mismatch in this range is simplified by using the Smith admittance chart of **fig. 4B** to calculate graphically the effect of the stub compensation. While we see conductance and susceptance components of admittance plotted separately in **fig. 5,** these components are plotted together in a single continuous locus on the Smith chart. The component values in the locus are identified separately by the appropriate conductance- or susceptance-circle graduations on the chart grid. We perform graphical addition of stub and dipole susceptances to determine the effect of the stub compensation by adding point-by-point values of the stub susceptance, b_s (obtained from **fig. 5**) to the corresponding dipole susceptance values (b_d) appearing on the locus of the uncompensated dipole admittance, y_d, on fig. 4B. Plotting the results of these additions yields the locus of the stub-compensated dipole admittance. Corresponding values of impedances are found by simply inverting the plots on the admittance graph (fig. 4B); we thus have plots of the equivalent impedances, as shown in fig 4A. To show how this works, the normalized, uncompensated dipole admittance is seen to be $0.5 - j0.725$ at the upper band-edge frequency, $+f_{pr}$. From fig 5 the stub susceptance, b_s, is $+j0.725$ at $+f_{pr}$. Returning to fig. 4B, we add these dipole and stub susceptances graphically: Following the arrow line along the $g = 0.5$ *circle*, we move point $+f_{pr}$ clockwise for a distance of 0.725 units of normalized conductance, to the chart center where $b = 0$. This move represents the capability of the stub to "pull" or warp the $0.5 - j0.725$ point on the uncompensated dipole admittance locus to the new, compensated admittance point, $0.5 + j0$. Going to the corresponding point in fig. 4A, the admittance-impedance inversion yields the expected normalized impedance of 2.0. The band-edge point used in this example was chosen for simplicity, but the same procedure may be used to convert any point on the uncompensated dipole admittance locus to determine the corresponding compensated value.

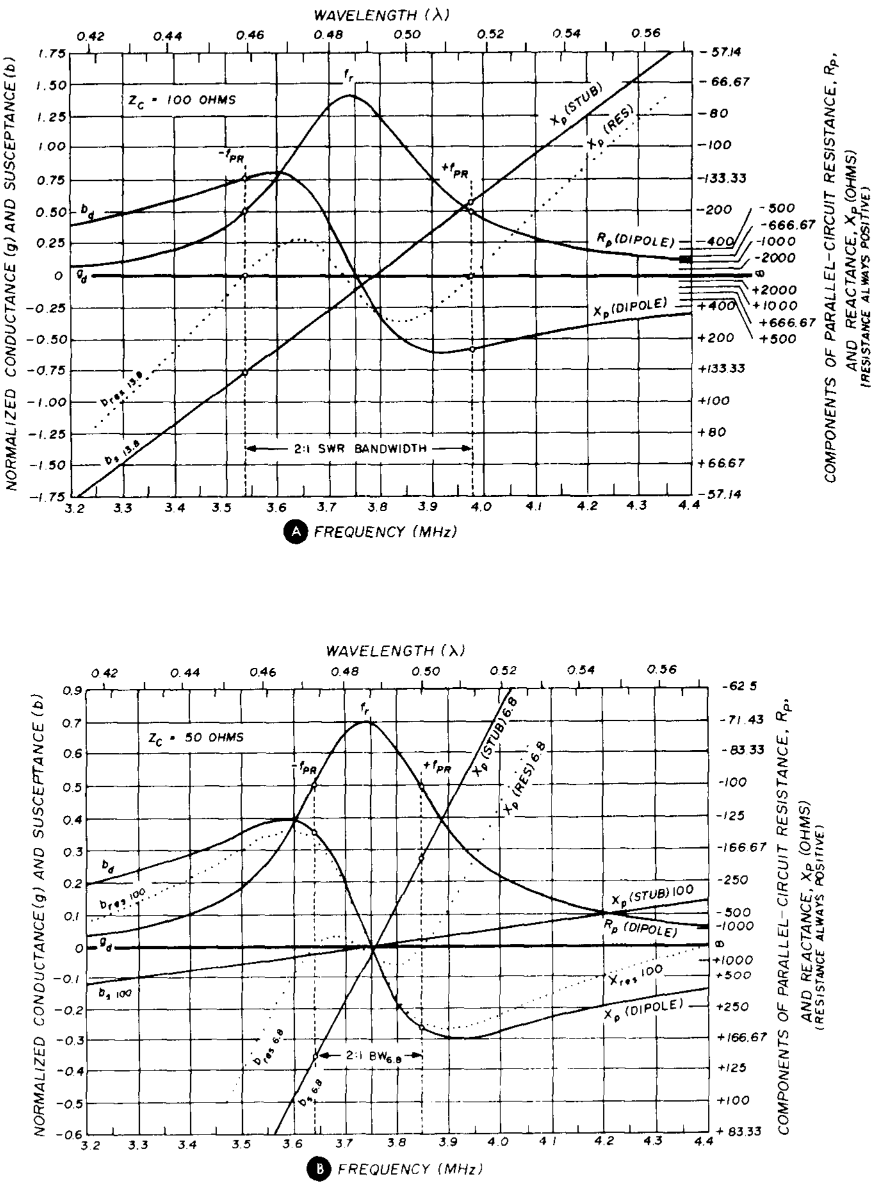


fig. 8. Normalized admittance parameters for uncompensated and compensated half-wavelength dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines. Conductance, g_d, and susceptance, b_d, are for the uncompensated case. Stub susceptance, b_s, and resultant susceptance, b_{res}, are also shown. The optimum stub susceptance for the 50-ohm case (b_s = 6.8 ohms) gives maximum swr bandwidth.

The concept of shunt-stub compensation will now be further clarified by a brief review of some relationships between equivalent series and parallel circuits which form the basis for our operations. These relationships were stated earlier in the mathematical expressions for converting between series and parallel circuits. This review which follows describes those expressions:

1. Provided Q is greater than 1, the component value of a parallel-circuit resistance, R_p, is higher than its corresponding series-circuit value, R_s.

2. When a series circuit containing resistance, R_s, and reactance, X_s, (an antenna) is shunted by a pure reactance of the opposite kind (a stub), the following changes result:

a. The resistance, R_p, and the conductance, G, components of the equivalent parallel circuit remain constant;

b. The reactance component, X_p, of the equivalent parallel circuit is increased, and the susceptance, B, is decreased.

c. The series-circuit resistance, R_s, is increased, and as the value of the shunting reactance (the stub) is changed in the direction toward cancellation of the net reactance, the series-circuit resistance, R_s, continues to rise until the circuit becomes parallel resonant. At this point the series-circuit resistance, R_s, becomes equal to the parallel-circuit resistance, R_p.*

In view of these relationships, it is important to remember that parallel-circuit resonance exists at both band-edge frequencies because the stub and dipole susceptances cancelled each other to zero. The pure resistance of 288 ohms at the dipole terminals at the band-edge frequencies has been raised to this value because it is the *parallel-circuit resistance* value of the uncompensated dipole impedance, which at both of these frequencies is simply the reciprocal of dipole conductance.

Let's now look at the effect of stub compensation on the separate series resistance, R_s, and reactance X_s, components of the dipole impedance over the entire frequency range extending somewhat beyond the band edges. To observe this effect the dipole conductance and *resultant* susceptance components from **fig. 5** are converted into their equivalent series resistance and reactance components of impedance, then plotted in **fig. 6** along with dipole resistance and reactance values plotted from **fig. 2** for direct comparison with the original, uncompensated dipole impedance components. Note the remarkable change in both the resistance and reactance components which resulted from a change in susceptance *only* – the dipole conductance remained *unchanged* by the stub compensation.

In addition to displaying the bandwidth in the Smith charts of **figs. 4A** and **4B**, the bandwidth obtained with shunt-stub compensation of the half-wavelength dipole in combination with the 144-ohm optimum-impedance feedline is also illustrated in **fig. 7** by plotting the feedline mismatch versus frequency. (The mismatch values were computed from the dipole conductance and *resultant* susceptance values in **fig. 5** using a technique applicable to pocket calculators described in **appendix 2**.) For comparison with the bandwidth of the uncompensated dipole, the mismatch values for the uncompensated dipole were computed from the dipole impedance components in **fig 2** and plotted in **fig. 7** along with the compensated values.

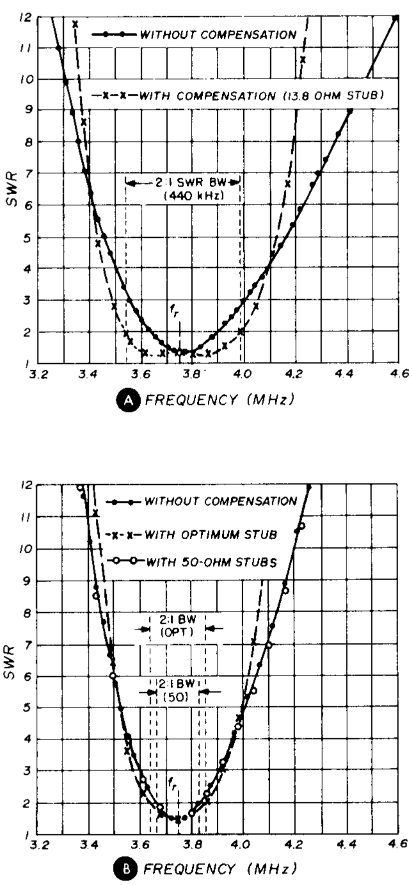


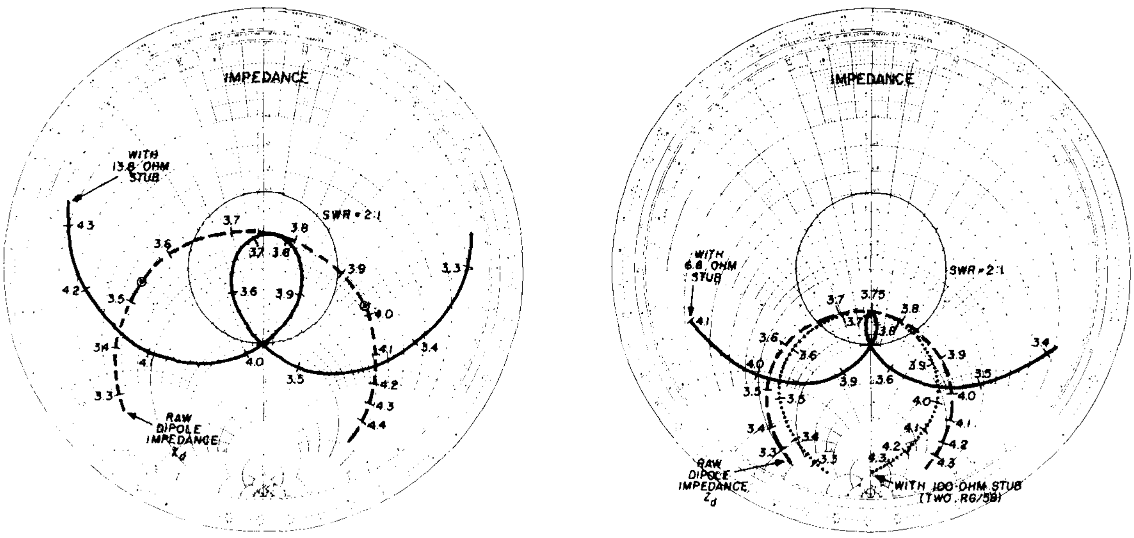
fig. 9. Swr vs frequency for compensated and uncompensated dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines.

In contrast to the data shown in **fig. 5** for a 144-ohm feedline, **figs. 8A** and **8B** show the effects of using feedline impedances of 100 ohms and 50 ohms, respectively. These are less suitably related to the dipole impedance values encountered over the frequency range than those using the optimum-bandwidth 144-ohm feedline impedance. These graphs show the corresponding normalized conductance and susceptance relationships in the same manner as the 144-ohm case shown in **fig. 5**. **Figs. 9A** and **9B** illustrate the bandwidths obtained using 100- and 50-ohm feedlines, both with and without compensation. As in the Smith-chart plots of **figs. 10A** and **10B**, they show clearly how the bandwidth drops as the line impedance is reduced from the 144-ohm optimum to 50 ohms.

*The basis of the hairpin match used in Yagi arrays. The 20-ohm (approximate) driven element is shortened to introduce enough series capacitance to raise the equivalent parallel-circuit resistance to 50 ohms. The resulting series capacitive reactance (–24.5 ohms) is then cancelled by the hairpin shunt-stub inductance, leaving the feedpoint impedance at 50 + j0 ohms.

From this graphical data you can see why the bandwidth decreases as feedline impedance is reduced. We compare resistance data in **figs. 5, 8A** and **8B** at the 2:1 mismatch bandedge frequencies, where the normalized dipole conductance is 0.5. It is seen that the parallel-circuit resistance component of the dipole impedance produces the 2:1 mismatch in each case of the three line impedances; the resistance at the 0.5 conductance point being equal to twice the line impedance. This mismatch-conductance relationship holds for any value of feedline susceptance available with a 100-ohm stub, as obtained with two series-connected 50 ohm stubs made from RG-8/U or RG-58/U coaxial cable.

The low slope of this plot is a clue to its meager compensating capability, vividly emphasized by the almost negligible difference between the raw dipole susceptance, b_d, and its corresponding resultant susceptance b_{res100}. Here the uncompensated 2.33:1 mismatch at the $-f_{pr}$ point is reduced only to 2.28:1 with the 50-ohm stubs. The bandwidth increase over the



A Z_c = 100 OHMS

B Z_c = 50 OHMS

fig. 10. Smith chart plots of normalized impedance for compensated and uncompensated half-wavelength dipoles fed with 100-ohm (A) and 50-ohm (B) feedlines. Optimum stub impedance for 100 ohm case is 13.8 ohms; optimum stub impedance for 50-ohm feedlines is 6.8 ohms.

impedance. Therefore, because the parallel-circuit resistance decreases with the use of lower values of feedline impedance, the bandwidth decreases between the two frequencies where the values of equal resistance appear on either side of resonance. Since these dipole-resistance values determine the frequencies of the 2:1 mismatch points, this demonstrates that the maximum obtainable 2:1 bandwidth decreases as the feedline impedance is reduced from 144 ohms.

Let's return to **fig 8B** for a closer look at the 50-ohm feedline case. This graph contains susceptance plots of two different shunt stubs with 6.8 and 100 ohms impedance respectively, along with their corresponding *resultant* susceptance plots. The straight line labelled $b_{s6.8}$ plots the susceptance of the stub which completely cancels the dipole susceptance at the 2:1 mismatch points, as shown in the corresponding resultant susceptance plot, $b_{res6.8}$. The uncompensated mismatches of 2.33:1 and the 2.34:1 at $-f_{pr}$ and $+f_{pr}$, respectively, are thus reduced to 2:1. (Not a very significant reduction.) While the impedance of this shunting stub is $Z_c = 6.8$ ohms, it requires a characteristic impedance of 3.4 ohms if two stubs are used in series (not a practical value of Z_c). The bandwidth is increased from 165 kHz (no stub) up to 210 kHz. On the other hand, the plot labelled b_{s100} shows the compensating no-stub, 165-kHz width is negligible; it is even too small to show graphically in **fig. 9B**. The Smith chart plot of **fig. 10B** verifies the disappointing performance of the coaxial dipole when it is fed with 50-ohm line – with *either* 6.8- or 100-ohm stubs.

This analysis of the 50-ohm feedline case, in addition to the mismatch graphs (**fig. 9B**) and the Smith chart impedance plots, clearly shows that there is no significant bandwidth improvement when feeding a coaxial dipole with 50-ohm feedline, especially when 50-ohm coax is used for the shunt-stub lines. Unfortunately, even the optimum 144-ohm line presents nearly insurmountable problems for amateur use. It is true that 144-ohm line could be built using two 72-ohm coax lines in a series, balanced relation – with the outer conductors tied together at both ends, and each inner conductor feeding one dipole half. This would be fine, except that at 80 meters the dipole resistance is usually less than 72 ohms. Unless we find a feedline having an impedance equal to *twice* the resonant dipole resistance, $R_d + j0$, the maximum 2:1 swr bandwidth will not be obtained. Another problem is that determining the optimum feedline impedance ($Z_c = 2R_d$) is simply an academic exercise unless its entire 2:1 swr range of impedances is transformed to the corresponding nominal 50-ohm range required by most amateur transmitters. A possible solution to this problem may be a broadband transformer described by Jerry Sevick, W2FMI,[12] but probably not without some loss in bandwidth. However, the most severe problem of all is the required low stub impedance — less than 20 ohms. This would require a balanced configuration of two series connected stubs, each less than 10 ohms. Their construction would involve an unwieldy combination of series-parallel quarter-wavelength sections of 50- and 75-ohm coax. Further, the positioning of such a kluge with respect to both its supporting members and the feedline presents additional complications which require more space to explain than is available here.

Since the mismatch values associated with the 50-ohm feedline case in **fig. 9B** and **10B** are higher than those measured by many coaxial dipole users, it is tempting to assume that this analysis is incorrect. However, the following factors responsible for lower *measured* values should be kept in mind.

1. The mismatch values plotted here are those which appear at the antenna-feedline junction. Mismatch values measured at the input end of the transmission line will be lower than at the antenna because of line attenuation (the greater the attenuation, the lower the input mismatch).*

2. In many cases swr indicators read lower than the true value.

3. The use of any additional broadbanding feature such as multiwire end loading, or larger radiator diameter when using RG-8/U coax as the radiator, reduces the inherent reactance which is developed, thereby lowering the mismatch.

Let's now consider WA9PIV's assertions concerning the gain and self-balun characteristics of the coaxial-dipole antenna. *Antenna gain is obtained by adding the far-field radiations from each, separate element of any array consisting of more than one dipole element.* The coaxial-dipole antenna is not an array, but a single dipole element, and thus has the *same* radiation pattern and the *same* gain as a simple dipole.

Regarding the self-balun characteristics, the bazooka formed by the shorted quarter-wavelength coaxial skirt surrounding a coaxial feedline, as shown in **fig. 11,** does indeed achieve a balanced-to-unbalanced (balun) action, resulting in cancellation of radiation from the feedline; this would otherwise occur as a result of current on the *inside surface of the outer conductor flowing around the* top and down the outside of the outer conductor in an admittance path which is in parallel with that half of the dipole fed by the outer feedline conductor. The term "bazooka" cannot be applied to the coaxial-stub configuration within the dipole. As a result, the balun function of the bazooka has been wrongly and unwittingly attributed to the coaxial feature of the coaxial-dipole antenna. The coaxial-dipole antenna is strictly a balanced-input device, and as stated in the opening paragraph, it is the same as the simple wire dipole, except for its impedance and bandwidth characteristics.

WA9PIV's further assertion that all harmonics are rejected by the coaxial dipole is not true because only the even harmonics are rejected. The reason is that the shorted stubs are multiples of a half-wavelength on even multiples of the fundamental frequency, and odd multiples of a quarter-wavelength on odd multiples of the fundamental. Thus a short circuit is reflected across the

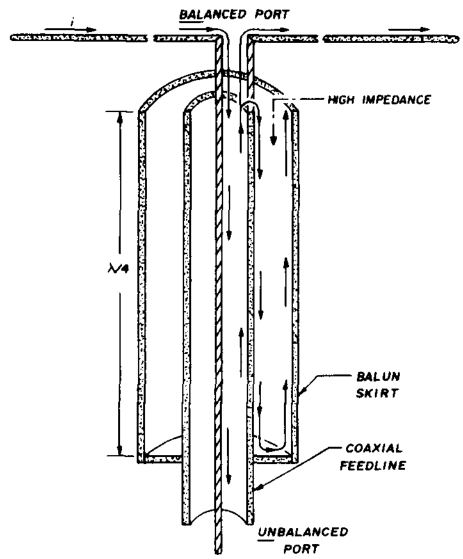


fig. 11. True bazooka formed by shorted quarter-wavelength coaxial skirt surrounding a coaxial feedline provides balanced to unbalanced transformation. This is not the case for the misnamed "bazooka antenna" which is strictly a balanced-input antenna. The high impedance (low admittance) appearing across the open end of the $\lambda/4$ resonant cavity of the bazooka shown here impedes current flow into the cavity, thus admitting practically all of the feedline current into the low impedance (high admittance) of the dipole. Removal of the coaxial skirt increases the admittance of current along the outside of the feedline.

input terminals to energy appearing at frequencies of the even-numbered harmonics, while an open circuit is reflected at odd-harmonic frequencies. As a result, harmonic energy is not suppressed at frequencies which are odd-harmonics of the fundamental frequency. These principles also hold true for the folded dipole.

Incidentally WA9PIV specifies dimensions for his coaxial dipole which make the stubs resonant at a frequency higher than the resonant frequency of the dipole. This simply shifts both band edges to frequencies slightly higher than those occurring when the stubs and dipole both resonate at the same frequency. However, regardless of his statement that mismatch is not greater than 1.5:1 over the entire 3.5 to 4.0 MHz band using his stub dimensions, the overall bandwidth is substantially the same as that shown in **figs. 9B** and **10B.**

It is clear, in listening on the amateur bands, that many amateurs, in their attempts to reduce swr, are afflicted with the coaxial-dipole syndrome. It is also clear that many are still placing unwarranted emphasis

*Instructions for calculating the corresponding mismatches appearing at opposite ends of the transmission line, for any value of line attenuation, are presented in **appendix 4.**

on low swr, usually for the wrong reasons. However, response to my series of articles on this subject[13] indicates that many amateurs are learning that when low-loss feedlines are used (which is generally the case on the high-frequency amateur bands), *a low swr on the feedline does not save any significant amount of power, nor does it noticeably enhance the level of the radiated signal.*

The only significant benefit of a low standing-wave ratio over an entire amateur band is the ease of matching your transmitter output to the *input* to the feedline anywhere in the band. Obtaining an acceptable impedance match is not too difficult on 40 meters, but on 80 meters the line input impedance varies so widely across the band (unless unknown losses reduce the mismatch at the input), it is possible to match the transmitter directly to the line only over a very limited frequency range. Unfortunately, hope for curing this ailment by using the coaxial dipole has been shattered, and anyone dispensing it as a cure should study this analysis.

Following are two suggested prescriptions for an effective-cure of the mismatch problem. As a partial cure try a dose of either W1SX's[10] or W9VMQ's[8] bow-tie dipole configuration mentioned earlier. This antenna provides some improvement over the thin-wire dipole, though not quite enough to permit coupling the average transmitter directly to the feedline across the entire 80-meter band. For a more complete cure, review the *QST* "Reflections" series (especially parts VI and VII),[13] and discover why the mismatch at the antenna-feedline junction that can't be cancelled by the stubs of the coaxial dipole *can be* compensated by conjugate matching at the input terminals of the feedline. Once this is understood, you can live with a three-, four-, or five-to-one swr on the feedline, and still make your transmitter happy by using a tuner to transform whatever impedance appears at the line input to 50 ohms of pure resistance at the tuner input at *any* frequency in the band.

Recommended out-patient treatment for this cure: One hour of operating per day for three or four days using this technique. This treatment will provide sufficient therapy to warrant discharge of the patient, and to guarantee a complete and permanent cure of the coaxial-dipole antenna syndrome.

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appendix 1

The magnitude of an impedance mismatch may be determined as swr from the relationships R/Z_c or Z_c/R *only* when the load is a pure resistance. When the load is a complex impedance, $Z_L = R + jX$, the exact mismatch may be determined in terms of the complex reflection coefficient (**Eq.** 1 of reference 13, part III) from

$$\bar{\rho} = \rho\angle\theta = \frac{Z_L - Z_c}{Z_L + Z_c} \qquad (1)$$

The value of swr may then be obtained from the expression

$$swr = \frac{1 + \rho}{1 - \rho} \qquad (2)$$

Swr, however, cannot be determined from any relationship between the feedline impedance, Z_c, and the simple absolute magnitude, $|Z|$, of the complex load impedance. This is evident from **table 1** where it can be seen that different values of load impedance, Z_L, can have the same magnitude, $|Z| = 50$ *ohms,* yet produce different values of mismatch on a given feedline.

Furthermore, it is improper to specify impedance as a pure number such as 50 ohms (as it is often heard on the amateur bands) unless it is in some way implied to be resistive, as when referring to the characteristic impedance of a low-loss transmission line. The proper way to specify impedance is to either use the complete polar form (magnitude and angle) or the equivalent rectangular form ($R + jX$) as shown in **table 1.**

table 1. Different swr values on a 50-ohm transmission line as a result of terminating the line with various values of impedance, Z, (different R and X) which have the same magnitude, $|Z| = \sqrt{R^2 + X^2} = 50$ ohms.

load impedance, Z			
polar form		**rectangular form**	**swr value**
$\|Z\|$	**angle**	R X	(50 ohm line)
50	$\angle 0°$	50 + j0	1.00
50	$\angle 25.84°$	45 + j21.79	1.59
50	$\angle 36.9°$	40 + j30	2.00
50	$\angle 53.1°$	30 + j40	3.00
50	$\angle 66.42°$	20 + j45.83	4.79
50	$\angle 72.54°$	15 + j47.7	6.51
50	$\angle 78.46°$	10 + j48.99	9.90
50	$\angle 84.26°$	5 + j49.75	19.95
50	$\angle 87.13°$	2.5 + j49.94	39.98
50	$\angle 88.85°$	1 + j49.99	99.99
50	$\angle 90°$	0 + j50	∞

appendix 2

Another method for calculating the exact swr from $R \pm jX$ (or $G \pm jB$), developed by the author after a suggestion by W2KF, and derived from **eq.** 1 in **appendix 1**, is shown in the following steps:

1. Normalize by dividing the complex impedance (or multiplying the complex admittance) by the characteristic impedance of the transmission line, Z_c.

$$\frac{R \pm jX}{Z_c} = r \pm jx \qquad (G \pm jB)Z_c = g \pm jb$$

Note that normalized values are in lower case letters.

2. Find the b term of the quadratic formula

$$\frac{b \pm \sqrt{b^2 - 4ac}}{2a}$$

using values of r and x (or g and b) in the expression

$$b = \left(\frac{x^2 + 1}{r}\right) + r \qquad b = \left(\frac{b^2 + 1}{g}\right) + g$$

3. Calculate the swr from the simplified quadratic formula

$$swr = \frac{b + \sqrt{b^2 - 4}}{2}$$

Note that the a and c terms of the complete quadratic formula in **step 2** reduce to one during the derivation due to the normalizing procedure and can be ignored. The negative root of the discriminant

$$-\sqrt{b^2 - 4ac}$$

is also disregarded.

Example: Determine the swr generated on a 50-ohm transmission line by a load impedance of 40 + j30 ohms.

Normalizing: $r + jx = \frac{40 + j30}{50} = 0.8 + j0.6$

Find the b term: $b = \left(\frac{0.6^2 + 1}{0.8}\right) + 0.8 = 2.50$

Calculate the swr: $swr = \frac{2.50 + \sqrt{2.50^2 - 4}}{2} = 2.000$

This is a very good example because the answer is *exactly* 2.0:1 with no fractional remainder. Other examples that give exact answers (50-ohm lines) are 30 + j40, swr = 3.0:1; and 80 + j90, swr = 4.0:1.

To show that different values of load impedance can yield the same swr, the following complex loads (50-ohm transmission line), will generate an swr of 2.6180:1 — 25 + j25, 50 + j50, 100 + j50, and 130 + j10.

appendix 3

Use the following procedure to determine the approximate value of the characteristic impedance, Z_c, of a shunt-compensating stub line from the slope of the susceptance plot:

1. Let θ equal the length in electrical degrees between $-f_{pr}$ and $+f_{pr}$. Find θ by determining the wavelength difference, $\Delta\lambda$, between $-f_{pr}$ and $+f_{pr}$, then multiply by 360 degrees

$$\theta = (\Delta\lambda)360°$$

2. Let X_{av} equal the average of the parallel-circuit dipole reactances appearing at $-f_{pr}$ and $+f_{pr}$ without regard for the sign.

3. The approximate impedance of the stub line can be found from

$$Z_c \approx \frac{X_{av}}{\tan(90 - \frac{\theta}{4})}$$

Example: Determine the impedance of the stub line from the b_s, susceptance plot in **fig. 5** ($-f_{pr} = 3490$ kHz, $\lambda = 0.453$; $+f_{pr} = 4055$ kHz, $\lambda = 0.527$).

$$\Delta\lambda = 0.527 - 0.453 = 0.074$$

$$\theta = 0.074 \cdot 360° = 26.64 \text{ degrees}$$

parallel-circuit reactances:

$-f_{pr}$	144 ohms
$+f_{pr}$	192 ohms
X_{av}	168 ohms

$$Z_c \approx \frac{168}{\tan(90 - \frac{26.64}{4})} = \frac{168}{\tan 83.34} = \frac{168}{8.56} = 19.62 \text{ ohms}$$

appendix 4

The mismatch SWR_A at the antenna feedline junction is higher than the mismatch SWR_I measured at the input to the transmission line because of line loss. When one mismatch is known use the following procedure to calculate the unknown mismatch at the opposite end of the line. Let

ρ_A = reflection coefficient at antenna (point A)

ρ_I = reflection coefficient at input (point I)

α = line attenuation in dB (multiply dB per foot times length of the line in feet)

r = decimal value of the output/input power-loss ratio of the feedline:

$$r = antilog_{10}\left(\frac{\alpha \text{ in } dB}{10}\right)$$

Example: If the line attenuation is 0.5 dB, what is the output/input power-loss ratio? (0.5 dB is expressed as a negative quantity since it is loss.)

$$r = antilog_{10}(-0.5/10) = 0.891$$

A. Use the following steps to calculate the mismatch at the antenna (SWR_A) from an swr measurement at the input to the transmission line (SWR_I).

1. Calculate ρ_I from SWR_I $\quad \rho_I = \frac{SWR_I - 1}{SWR_I + 1}$
2. Calculate output/input power-loss ratio, r, from line attenuation, α
3. Calculate ρ_A from ρ_I/r (ρ_A is larger than ρ_I)
4. Calculate SWR_A from ρ_A $\quad SWR_A = \frac{1 + \rho_A}{1 - \rho_A}$

Example: The input swr to a 120-foot RG-8/U feedline is 3.5:1 at 4.0 MHz. What is the swr at the antenna? (Attenuation of RG-8/U is 0.32 dB per 100 feet at 4.0 MHz so attenuation of 120 feet is 0.384 dB.)

$$\rho_I = \frac{SWR_I - 1}{SWR_I + 1} = \frac{3.5 - 1}{3.5 + 1} = \frac{2.5}{4.5} = 0.556$$

$$r = antilog_{10}\frac{-0.384}{10} = 0.915$$

$$\rho_A = \frac{\rho_I}{r} = \frac{0.556}{0.915} = 0.607$$

$$SWR_A = \frac{1 + \rho_A}{1 - \rho_A} = \frac{1 + 0.607}{1 - 0.607} = \frac{1.607}{0.393} = 4.093{:}1$$

B. Use the following steps to calculate the swr at the input of the transmission line (SWR_I) from a mismatch measurement at the input to the antenna (SWR_A):

1. Calculate ρ_A from SWR_A $\quad \rho_A = \frac{SWR_A - 1}{SWR_A + 1}$
2. Calculate r from line attenuation, α
3. Calculate ρ_I $\quad \rho_I = \rho_A \times r$ (ρ_I is smaller than ρ_A)
4. Calculate SWR_I from ρ_I $\quad SWR_I = \frac{1 + \rho_I}{1 - \rho_I}$

Example: The swr at the input to an antenna is 5:1 at 4.0 MHz. What is the swr at the input of a 156.25 foot length of RG-8/U transmission line? (Attenuation of RG-8/U is 0.32 dB per 100 feet at 4.0 MHz so attenuation of 156.25 feet is 0.5 dB.)

$$\rho_A = \frac{SWR_A - 1}{SWR_A + 1} = \frac{5.0 - 1}{5.0 + 1} = \frac{4.0}{6.0} = 0.667$$

$$r = antilog_{10}\frac{-0.5}{10} = 0.891$$

$$\rho_I = \rho_A \times r = 0.667 \times 0.891 = 0.594$$

$$SWR_I = \frac{1 + \rho_I}{1 - \rho_I} = \frac{1 + 0.594}{1 - 0.594} = \frac{1.594}{0.406} = 3.926{:}1$$

ham radio

differential keying circuit

Low-cost TTL logic devices are combined in a keyer design with optional weight control

An investigation of differential keying circuits in tube, transistor and relay form revealed that such circuits could be designed using TTL or cmos IC logic. Some IC multivibrators were found in the Texas Instruments *TTL Data Book.*[1] One was the SN 74123, which has two multivibrators in one package, and the other was the SN74121, which has only one. The SN74123 was chosen since this meant only one 16-pin socket, and the second multivibrator could be used for another function such as weight control.

circuit description

Reference 2 indicated that the SN74121, SN74123-series inputs require a clamped threshold near ground or at a slightly negative voltage (about -1.2 volt). I selected a clamp at ground because otherwise another power-supply bias level would be necessary; this clamped threshold at ground was adequate for the keyer application.

Waveforms and a diagram of the circuit are shown in **figs. 1** and **2** respectively. By following these two illustrations, we can run through the operation of how the differential keying is formed. The NOR and NAND logic is necessary to develop the gating needed. An SN7402 was selected as the NOR gate (J or N type – the pin-out is different for other types), and the SN7410 was selected as the NAND element simply because I had some left over from a previous purchase from Poly Paks (one of many sources for these devices). **Fig. 2** shows one of the four dual-input NOR gates used as an inverter and the clamp to ground for the input of the SN74123 (U1A-B). Both inputs are connected through a current-limiting resistor to a high logic level, thus producing a low at the output, which is very close to ground (0.8 volt or so). The input biasing by means of R1 also provides bias to my keyer output transistor switching stage.

When the keyer output goes to ground U3C output goes high and U1A, being connected for a positive trigger input, produces the inhibit pulse, P3, seen in **fig. 1A**. The completion of the keyer pulse, P1, results in a negative trigger, which is received by the second half of the SN74123, (U1B). Thus, U1B develops pulse P5. Pulses P2 and P3 are processed by NAND gate U2. The output conforms to NAND logic in that an output will occur only with like inputs (positive logic is used, so this means both the inputs must be high). Therefore pulse P3, being negative, inhibits the output during its time interval, and output pulse P4 is formed.

Now the gate pulse that will keep the oscillator on must be developed. This gate pulse is formed by adding the output of U3C with that of U1B (positive or Q output in this case). The final gate output becomes that shown by P6. The third NOR gate of the SN7402, U3B, is used for another inverter, so a positive pulse, P7, is generated.

interface circuits

We now come to the all-important function of interfacing the developed logic pulses with the equipment being considered. Transistors Q2 and Q3 perform this interface and are general-purpose, high-voltage npn and pnp transistors needed for the hybrid interface between logic IC levels and the usual high bias levels of vacuum tubes. R7 provides limiting for the NOR load current, and Q2 base current. The positive output pulse, P7, switches Q2 on, which in turn switches Q3 on, and the

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transmitter oscillator plate and/or screen receives its positive bias voltage.

Instead of using complete on-off switching of bias voltage or voltages, a resistor should be added between the positive bias and the oscillator circuit being controlled, so that the bias level is decreased but still allows low-level oscillation. When full positive bias is applied, the output increases to its normal level in a smooth transition from low to high levels rather than starting from zero oscillator output, which could result in transients. R8 limits Q3 base current and provides additional isolation should the transistors fail. The output switching transistor collector and emitter can be connected in various ways to satisfy the interface requirement of whatever circuit switching is needed.

The keying-pulse interface is made fairly easy by merely using the same type pnp transistor used in the gate interface and turning it on when the NAND gate output goes on (pulse-switches from a high positive level to a low-level ground). R6 provides the gate load and transistor base-current limiting, while Q1 switches the grid-block circuit to ground, thus keying the transmitter exciter and starting the pulse shaping process. The keying pulse is symmetrically located within the gate time interval and therefore performs the desired differential keying function. Normally, grid-block keying circuits are of a high-impedance level so that no further protection is needed should the transistor fail. If Q1 failed, the exciter would remain keyed to ground, but no harm would befall integrated circuit U2.

weight control addition

If weight control is not included in the keyer (or if a bug or hand key is used), then another multivibrator, such as the SN74121, must be used with an additional NOR gate. In fact, the NOR gate used for forming the acceptance gate must be a three-input type (the SN7427 is available from most sources). The three-input NOR can also be used for the two-input requirement by merely connecting two of the three inputs together (see **fig. 3**).

The logic of these gate arrangements is most easily understood by describing the different timing waveforms again (see **fig. 1B**). Input pulse P1 is the keyer or hand-key output and turns off U3C when grounded or when the contacts are closed. This action increases positive

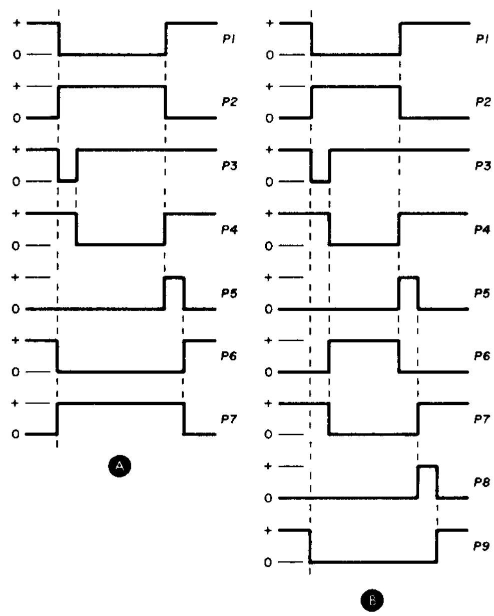


fig. 1. Timing waveforms of differential keyer logic with weight control included in keyer (A) and with weight control added (B).

pulse P2 from ground (about 0.8 volt) to about +3.4 volts. One-half of the SN74123 still forms the negative-going inhibit pulse, which forms the keyer output pulse delay, P3, by preventing any output from NAND gate U3A until all inputs are at a high level. The second half of the dual multivibrator U1B (SN74123) is connected to accept a negative trigger input, thus forming the weight-control pulse, P5.

The trailing edge of P5 triggers acceptance-gate pulse

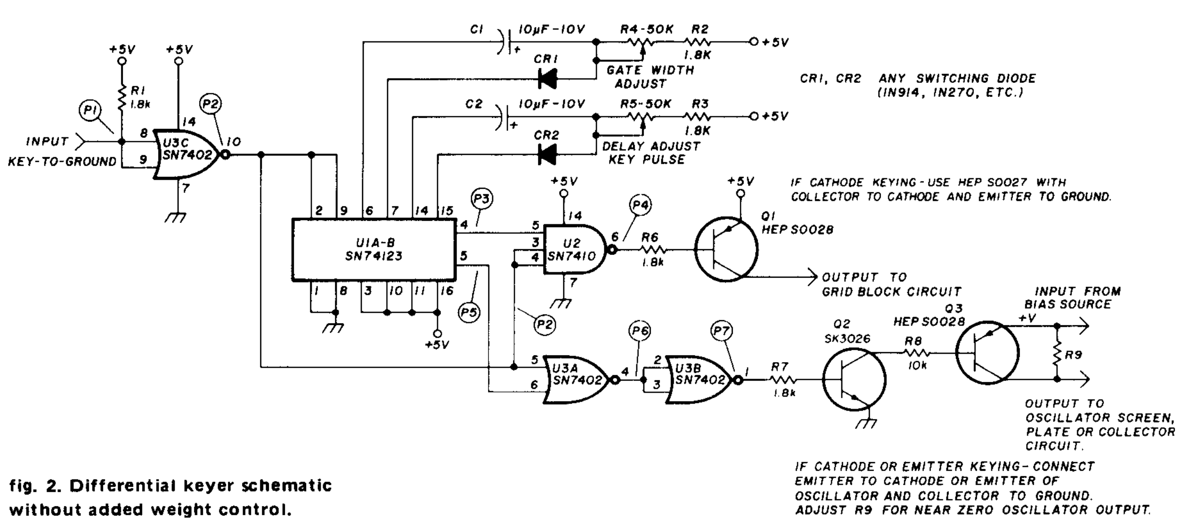
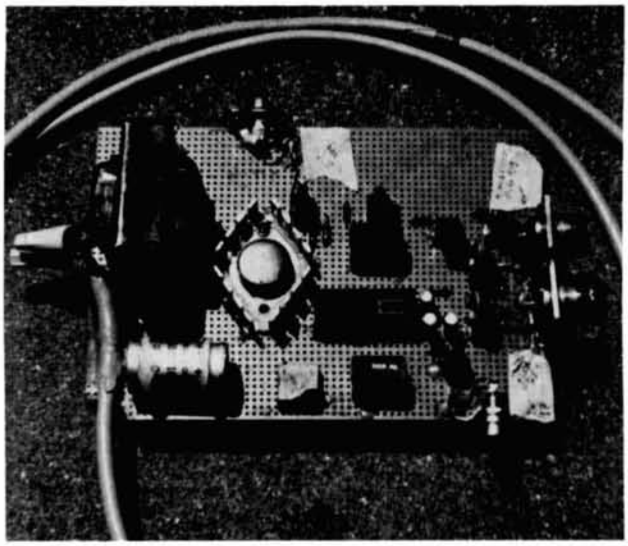


fig. 2. Differential keyer schematic without added weight control.



Component side of perf board showing general layout.

stretcher multivibrator U2 (SN74121) and forms P8. Now, P2, P5, and P8 are added by the three-input NOR gate, U4B, producing the required acceptance-gate pulse, P9, that will turn on the VFO or oscillator before and after the keying pulses have been completed (P5 and P6). Keying pulse P4 is inverted by U3B, forming P6, which is then added to the weight-control, multivibrator (U1B) output P5 by U4A, forming the desired keying-pulse interval, P7.

Note that the acceptance-gate interval, P9, is formed by the keying-pulse width, P2, the weight-control multivibrator output, P5, and the gate-extension pulse width, P8, such that any variation in the weight-control pulse also varies the gate acceptance interval, thus inadvertent extension of the keying pulse interval, P7, beyond the acceptance-gate interval, P9 is prevented. Such extension of P7 would cause spurious radiation in the form of key clicks. Also, if any variation is desired, only the weight-control multivibrator time interval adjustment is necessary.

It may be found that when the speed is varied, the weight control may have to be changed slightly. Since exciter or transmitter keying characteristics are usually fixed and drive levels change with frequency, the weight control may have to be changed to obtain the same desired keying quality or characteristic (with class-C amplifiers).

The same low-to-high level interface transistor switching used in the differential key circuit without weight control is used here; however, U3B was required to allow the proper positive level input for U4A with respect to the keyer pulse, P4. U3B merely serves as an inverter by connecting all inputs together.

construction

Nothing is critical in this circuit. Straightforward point-to-point insulated wiring was used. One rf decoupling capacitor was found necessary at the input of the differential keyer circuit (0.001 μF ceramic). My glue gun came in very handy for strapping down sockets and components before wiring. The glue takes a minimum length of time to dry, but be careful about using the glue on temperature-sensitive components since it's initially very hot.

The power supply consists of a filament transformer rated at one ampere (much more than required), an LM309K regulator, and a bridge rectifier IC. All were mounted on the same board. The filter capacitor is a 1500 μF 10V electrolytic. A zener diode would work as well as the LM309K and requires less space.

alignment

Alignment was easily accomplished with the aid of an oscilloscope such as an EICO model 460. Alignment

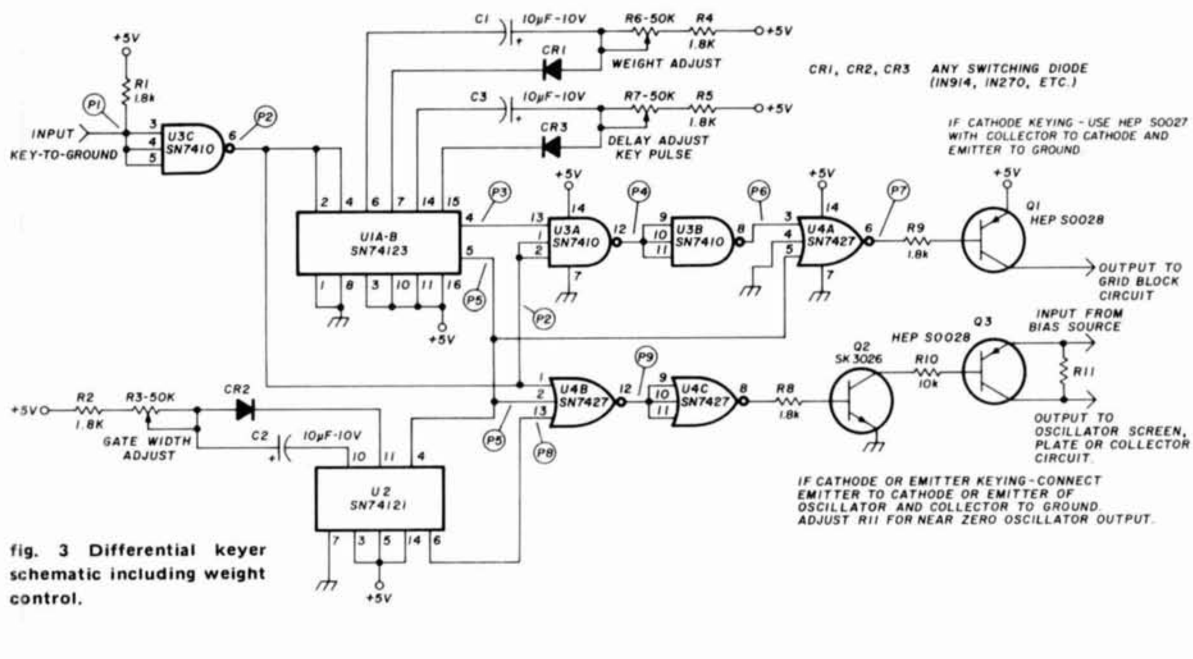
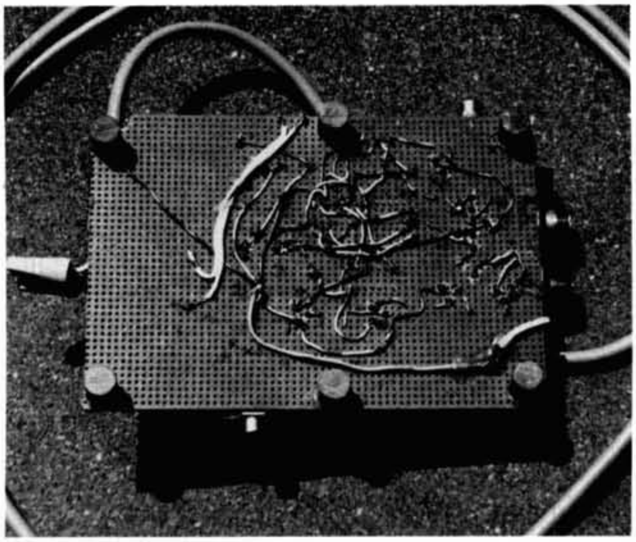


fig. 3 Differential keyer schematic including weight control.

would be almost impossible without an oscilloscope. My Johnson T/R switch has an rf output test point, which allows the transmitted wave shape to be observed when it's within the oscilloscope bandpass, otherwise the scope detector probe can be used and the envelope observed.



Underside of perf board showing point-to-point wiring.

Delay multivibrator U1A pulse width was adjusted by R5, **fig. 2**, until the front of the shaped pulse started to decay with near zero rise time, then R5 was backed off until the shaped pulse reappeared. This procedure was also used by varying R4 while observing the pulse trailing edge with respect to the acceptance-gate pulse-width stretcher multivibrator, U1B. These controls require no further adjustment, and the keyer weight control is used for any differences that may be experienced with varying code speeds.

If the weight-control logic is added the weight control multivibrator would be set at minimum pulse width, while the acceptance-gate multivibrator pulse width would be varied to eliminate any decay of the shaped-pulse trailing edge. The weight control can then be adjusted to produce the desired keying characteristic.

The circuit has worked well without any malfunctions. Incidentally, the TTL logic has bias limits. It will not work properly or will fail completely if the bias becomes greater than 5.5 volts. A bias value between 4.7 and 5 volts with good regulation by means of a zener or IC regulator device is recommended.

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1. *The TTL Data Book for Design Engineers*, First Edition, Texas Instruments, Inc., Dallas, Texas 1973.
2. *Designing with TTL Integrated Circuits*, Texas Instruments, Inc., McGraw-Hill Book Company, New York, 1972.

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TTL IC tester

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I can remember when you could plug a one-dollar vacuum tube into a ten-cent socket. The situation now seems to be reversed: a twenty-cent integrated circuit plugs into a socket costing anywhere from fifty cents to a dollar or more. When I was planning a construction project that would require 63 ICs, sockets were out of the question – just too darned expensive. This simple tester was then constructed so that each IC could be tested before it was soldered into place. The ICs shown were for the project mentioned above. Some could be deleted or other types added according to your favorites or the contents of your junk box.

circuit description

Nine input stimuli are generated by feeding +5 volts through resistors R1 through R9 as shown in **fig. 1**. Each input stimulus is either a 1 (+5V) or a 0 (ground), depending on the position of its corresponding switch S1 through S9. The output states are indicated by lamps DS1 through DS6. If the lamp is ON, a 1 is indicated; OFF indicates a 0 state. Q1 through Q6 can be any npn transistor that will carry the 150-mA lamp current. I used some TO-5 germanium transistors removed from computer PC boards. An alternative output indicator circuit could be an LED with the proper current-limiting resistor in place of the transistor and lamp. The 7413 Schmitt trigger is used as a "de-bouncer," allowing a single pulse to be produced by manually depressing switch S11 (**fig. 2**). This pulse is used in testing JK flip-flops.

construction

The test sockets and lamp drivers are mounted on a piece of perf board approximately 4½ x 2-3/8 inches (114 x 60mm) (**fig. 3**). This board is mounted on 5/16-inch (8mm) spacers above an aluminum chassis, which is 4 x 5 x 1½ inches (101 x 127 x 38mm). All switches and lamps are mounted on the aluminum chassis, and resistors R1 through R9 are mounted underneath. The test sockets are wired in parallel; that is, all pins requiring a no. 1 input are connected together and to S1 and R1; all

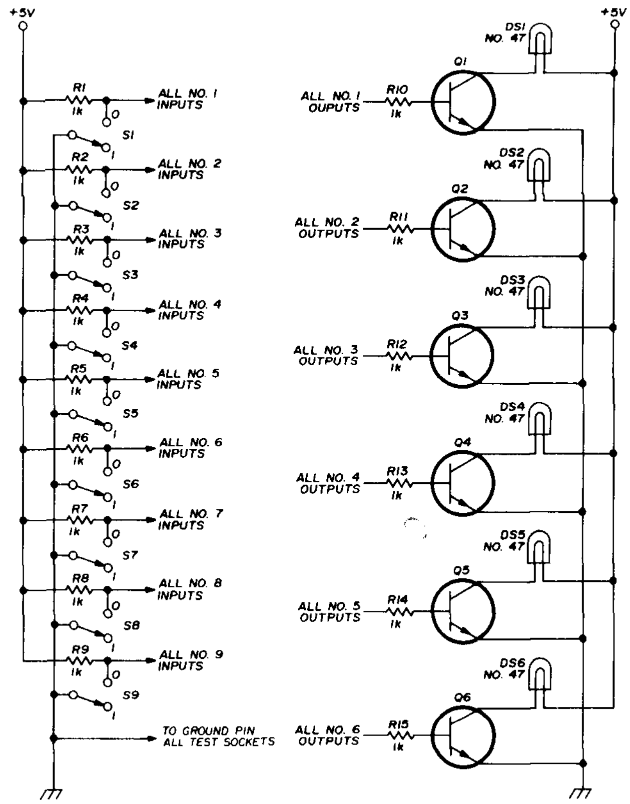


fig. 1. Schematic of the IC tester.

no. 2 inputs pins are connected together and to S2 and R2, etc. A 5-volt power supply at 1 ampere is required to power the IC tester. If you're careful not to light more than two lamps at a time, 0.3 ampere would be sufficient.

operation

The diagrams shown in **fig. 2** may be used as a guide when operating the tester. Perhaps, when testing JK flip-flops, it would also be advantageous to have the specifi-

By **Kenneth H. Leiner, WA4LCO,** 3254 Inverness Court, Orlando, Florida 32806

table 1. Logic matrix for selected ICs.

IC	S1	S2	S3	S4	S5	S6	S7	S8	S9	S10	S11	DS1	DS2	DS3	DS4	DS5	DS6
7400	1	1	1	1	1	1	1	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	1	0	1	0	1	0	1	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
	0	1	0	1	0	1	0	1	NU	NU	NU	ON	ON	ON	ON	NU	NU
	0	0	0	0	0	0	0	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
7402	1	1	1	1	1	1	1	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	1	0	1	0	1	0	1	0	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	0	1	0	1	0	1	0	1	NU	NU	NU	OFF	OFF	OFF	OFF	NU	NU
	0	0	0	0	0	0	0	0	NU	NU	NU	ON	ON	ON	ON	NU	NU
7404	1	1	1	1	1	1	NU	NU	NU	NU	NU	OFF	OFF	OFF	OFF	OFF	OFF
	0	0	0	0	0	0	NU	NU	NU	NU	NU	ON	ON	ON	ON	ON	ON
7410	1	1	1	1	1	1	1	1	1	NU	NU	OFF	OFF	OFF	NU	NU	NU
	1	1	0	1	1	0	1	1	0	NU	NU	ON	ON	ON	NU	NU	NU
	1	0	1	1	0	1	1	0	1	NU	NU	ON	ON	ON	NU	NU	NU
	0	1	1	0	1	1	0	1	1	NU	NU	ON	ON	ON	NU	NU	NU
	0	0	0	0	0	0	0	0	0	NU	NU	ON	ON	ON	NU	NU	NU
7473	0	1	0	1	1	1	NU	NU	NU	X	YES	OFF	ON	OFF	ON	NU	NU
	1	0	1	0	1	0	NU	NU	NU	X	YES	ON	OFF	ON	OFF	NU	NU
	X	X	X	X	0	0	NU	NU	NU	X	NO	OFF	ON	OFF	ON	NU	NU
	1	1	1	1	1	1	NU	NU	NU	X	YES	TOGGLE		TOGGLE		NU	NU
7476	1	0	1	0	1	1	1	1	NU	X	YES	ON	OFF	ON	OFF	NU	NU
	0	1	0	1	1	1	1	1	NU	X	YES	OFF	ON	OFF	ON	NU	NU
	X	X	X	X	1	1	0	0	NU	X	NO	ON	OFF	ON	OFF	NU	NU
	X	X	X	X	0	0	1	1	NU	X	NO	OFF	ON	OFF	ON	NU	NU
	1	1	1	1	1	1	1	1	NU	X	YES	TOGGLE		TOGGLE		NU	NU

X = Don't-care condition (1 or 0). NU = Not used. TOGGLE = DS1 and DS2 (DS3 and DS4) alternate with each S11 pulse.

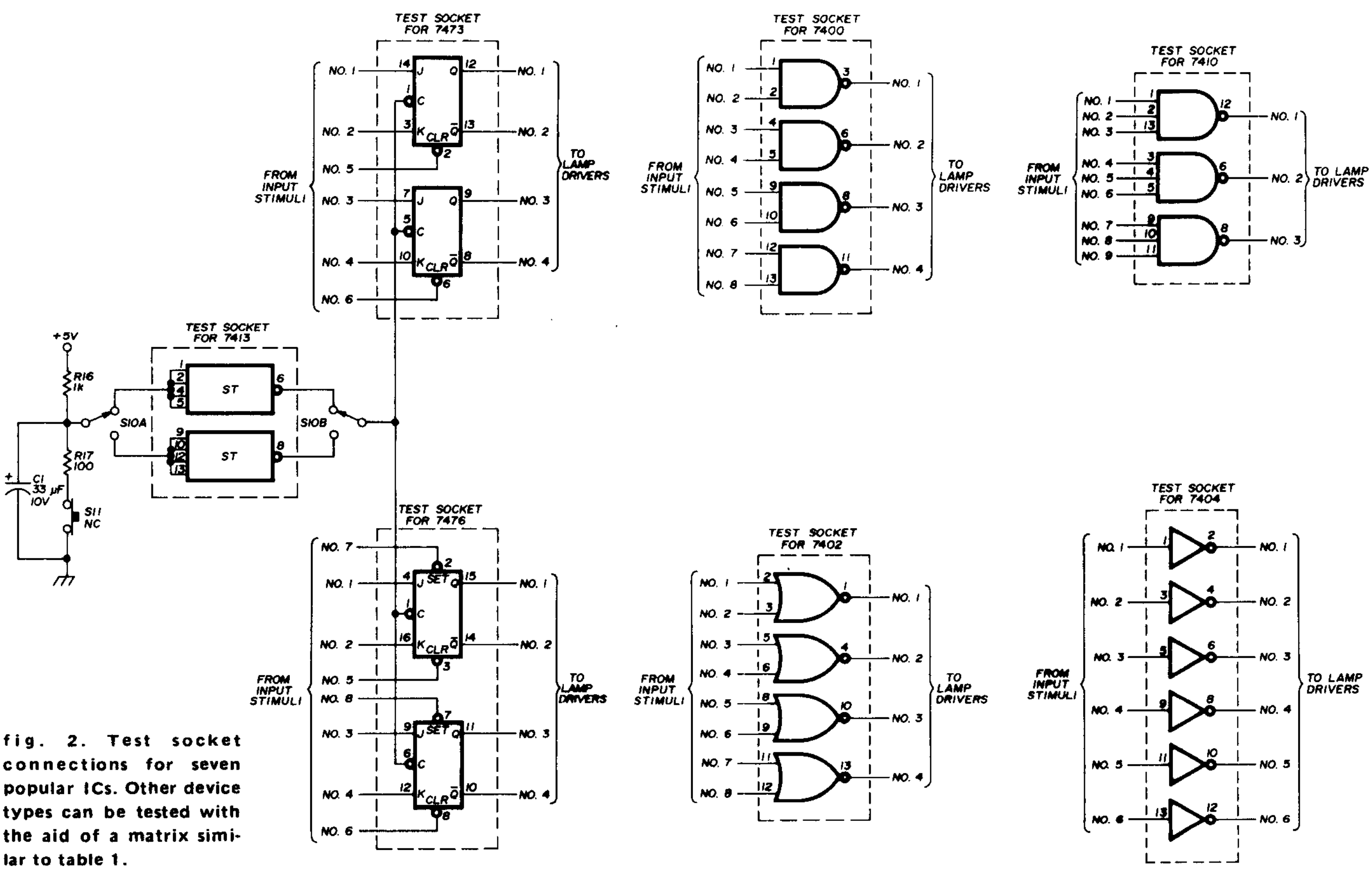


fig. 2. Test socket connections for seven popular ICs. Other device types can be tested with the aid of a matrix similar to table 1.

cation sheet handy. If desired, tables may be constructed similar to **table 1**. Always turn off the 5-volt power when inserting or removing an IC from the socket. In removing ICs from the test sockets, slip a small screwdriver blade under the IC. Then rock the IC gently up and down, raising first one end slightly, then the other end. Repeat until the device is free of the socket. This procedure usually removed the IC without bending any of the metal pins.

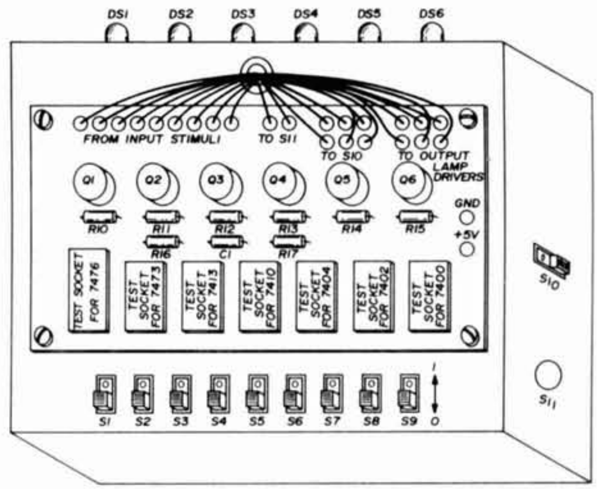


fig. 3. Suggested component layout.

When testing JK flip-flops, a 7413 must also be inserted in its socket. Also, a 7413 can only be tested if a 7473 or a 7476 is in the tester. As wired here, the 7413 has all its input pins connected together, so the test for the 7413 is not complete. The 7413 was really included to produce clock pulses for testing JK flip-flops, so whatever test is available for the 7413 is a bonus. Switch S10 allows both Schmitt triggers of a 7413 to be tested. At all other times, only one IC at a time is inserted into the tester. In testing flip-flops, when power is first applied it is sometimes necessary to depress S11 twice in making the first test. Probably this is because when power is turned on the various internal flip-flops may assume any state, and it takes one pulse to get them into the proper relationship.

conclusion

The tester was used to check about 100 integrated circuits. Three defective units were found. A 7410 had a defective gate; pin 6 remained high at all times. A 7402 showed a dim glow on DS2 and DS4 when they should have been OFF. (This trouble may not have been found if LEDs had been used as output indicators.) Finally, one-half of a dual JK flip-flop remained in the 1 state at all times.

It may be doubtful at times who'll win, but at least when debugging the 63-IC project I had a head start on Murphy and his wretched laws.

ham radio

More Details? CHECK—OFF Page 126

50-MHz bandpass filter

A bandpass filter of unusual design that provides 6% bandwidth at 50.5 MHz with only 4-dB insertion loss

The design of highpass, lowpass, and bandpass filters for use at hf and vhf has been covered in recent amateur literature.[1,2,3] The professional literature has also offered design aids in the form of *slide-rule devices* for use in filter synthesis[4] and in graphs.[5] The article on hf bandpass filters for receivers by W7ZOI[3] is an excellent example of showing what can be done and how simple these filters can be. W7ZOI is to be commended for combining amateur know-how with laboratory equipment to demonstrate the selectivity of his designs. His **fig. 7** (reference 3) at first appeared too complex and at the same time reminded me of a similar filter I had hiding in the garage.

filter characteristics

The garage relic is of unknown origin and as **fig. 1** shows, is rather sophisticated. **Fig. 2** is a plot of this filter's response taken from an x-y recorder (using a hand-tuned signal generator). The insertion loss (4 dB) and a bandwidth of 6 percent at the 3-dB points seem pretty good, considering the 50.5-MHz center frequency and the amount of wire on the coils.

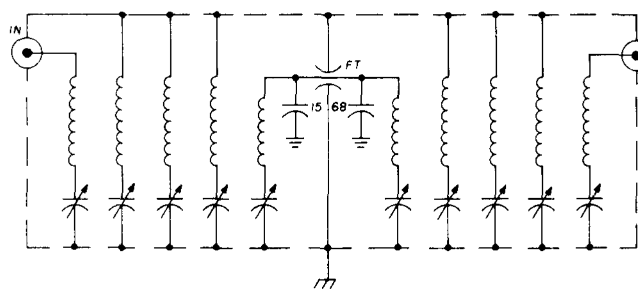


fig. 1. Schematic of the vhf bandpass filter. Center frequency is 50.5 MHz, bandwidth 6 percent, and insertion loss 4 dB. Each of the inductors is about 2.2 μH; variable trimmers are 1.5-7 pF.

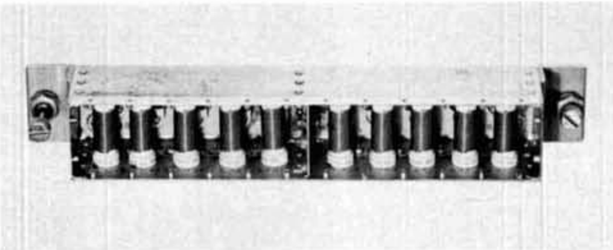
A photo from a Hewlett-Packard spectrum analyzer **(fig. 3)** shows the skirt slope. Vertical divisions are 10 dB and the horizontal scale is 1 MHz/cm.

construction

I don't recommend construction of this filter unless you have a sweep signal generator and a 5-inch (13cm)

By **Paul H. Sellers, W4EKO,** 4002 Columbus Avenue, Norfolk, Virginia 23504

oscilloscope for alignment. Alignment is tedious and quite ticklish. For those brave enough to attempt to duplicate this filter, the photo and **fig. 4** are provided. The coils forms are ribbed Teflon rod, ½ inch (12.5mm) in diameter. Each of the coils are 21 turns no. 20 AWG (0.8mm) wire; winding length is 1-3/64 inches (26.5mm). Coil ends are inserted through holes in each end of the Teflon rod **(fig. 4A)**. Overall coil diameter, including wire and ribs, is 9/16 inch (14.5mm). The coils are spaced as in **fig. 4B,** which is a side view of the filter showing the hold-down screws for the coils. The variable capacitors are about 1.5-7 pF, and the fixed capacitors are Corning type CY10C, 150J and 680J.



Inside the 50-MHz bandpass filter showing coil arrangement and center shield. (Photo courtesy Paul Ireland).

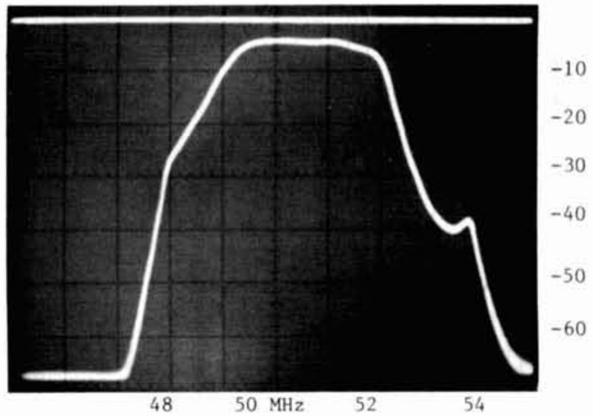


fig. 3. Photo from H-P spectrum analyzer showing filter skirt slope. Vertical scale: 10 dB/division; horizontal scale: 1 MHz/cm.

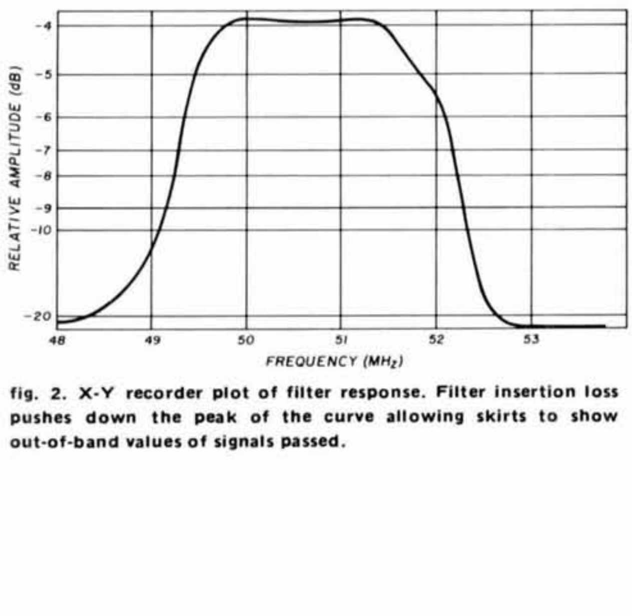


fig. 2. X-Y recorder plot of filter response. Filter insertion loss pushes down the peak of the curve allowing skirts to show out-of-band values of signals passed.

I'm sure you'll appreciate the design of this filter, including the unusual input/output circuits and the purely inductive coupling between stages.

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1. Bob Myers, W1FBY, and Clarke Green, WA1JLD, "Field Day Filter," *QST,* April 1973, page 11.
2. Neil Johnson, W2OLU, "High-Frequency Low-pass Filter," *ham radio*, March, 1975, page 24.
3. Wes Hayward, W7ZOI, "Bandpass Filters for Receiver Preselectors," *ham radio,* February, 1975, page 18.
4. *Genistron Filter Slide Rule,* Genistron Inc., Los Angeles, California, 1965.
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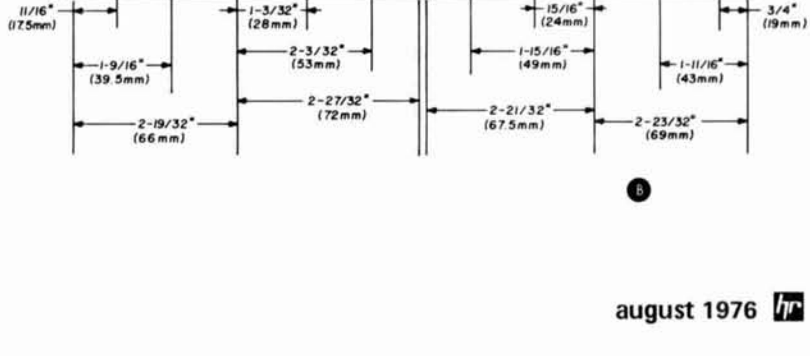

fig. 4. Coil construction details (A) and side view of filter box showing coil spacing (B). Enclosure dimensions are 2-1/8 inches (54mm) deep and 1-7/8 inches (47.6mm) across opening.

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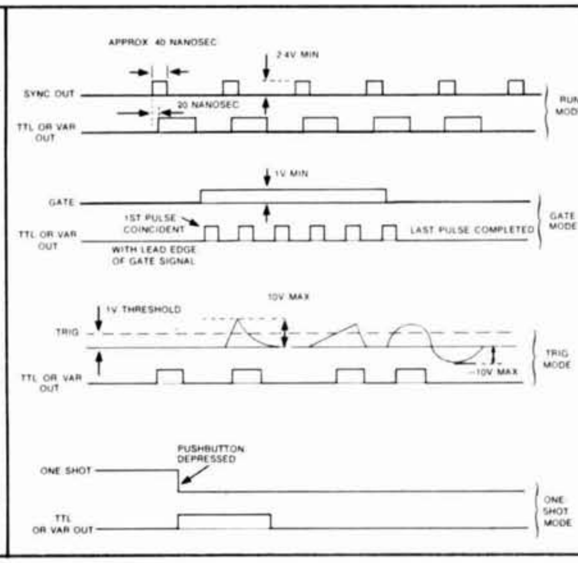
Pulse Width and Spacing Controls:	100 nanosec to 1 sec. in 7 overlapping decade ranges. A single-turn vernier control provides continuous adjustment between ranges.	**OUTPUTS:**	
		VAR OUT	
		Amplitude	0.1-10 V positive
Duty Cycle:	10⁷-to 1 Range: adjustable over entire pulse width/spacing range. 100 nanosec "ON" 1 sec "OFF" to 1 sec "ON" and 100 nanosec "OFF".	Rise/fall time	Less than 30 nanosec
		Impedance	400Ω max
		TTL OUT	
		Fan-out	40 TTL Loads
Operating Modes:		Sink	160 milliamps—0.8 V max
RUN	0.5 Hz to 5 MHz as per width/spacing and amplitude control settings.	Rise/fall time	Less than 20 nanosec
		SYNC OUT	
TRIG	DC to approx. 10 MHz	Pulse width	Approx. 40 nanosec. Other sync pulse spec's same as TTL out.
Input requirements:	Sine waves 2 VP-P, pulses 1 V peak, >40 nanosec wide, maximum input ±10 V (Input Impedance: Approx. 10KΩDC coupled.)	Pulse lead time	Sync pulse leads outputs by approx. 20 nanosec.
GATE	Synchronous gating. Leading edge of gate signal turns generator "ON". Last pulse is completed, even if gate ends during pulse.	**POWER**	117 VAC ± 10%, 50/60 Hz, 5 watts (220 VAC, 50/60 Hz also available, at slightly higher cost.)
		SIZE (WxLxH)	7.5 x 6.5 x 3.25" 191 x 165 x 83 mm
Input requirements:	Same as "TRIG" Mode.	**WEIGHT**	2 lbs. (0.91 Kg)



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MICROPROCESSORS

microcomputer interfacing: how does a microcomputer make a decision?

One of the most important programming characteristics in any digital computer, including a microcomputer, is the ability to make a decision. For a typical microcomputer, we can define a *decision* as the process of determining further action based on the logic state of a *flag.* A *flag* is a single flip-flop that can be either set or cleared in response to operations occurring within the microcomputer system. A change of state of the flag is usually an indication either that a particular operation has been completed, or that a certain condition exists as a result of a microcomputer operation. Flags can be located either internally or externally to the microprocessor chip; those discussed here are the internal flags, which are set or cleared in response to specific types of microprocessor instructions, such as arithmetic and logical instructions.

The flags located within the microprocessor chip are typically associated with the *arithmetic-logic unit (ALU),* a region within the chip where all arithmetic and logical operations are performed. In the 8080 microprocessor chip, for example, five flags indicate the following conditions;

By Jonathan Titus, David G. Larsen, WBHYJ, and Peter R. Rony

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

Zero flag If the result of an arithmetic or logical operation is zero, the zero flag is set to logic 1; if nonzero, the zero flag is reset to logic 0.

Sign flag If the result of an arithmetic or logical operation is negative, the sign flag is set to logic 1; if positive, the sign flag is reset to logic 0.

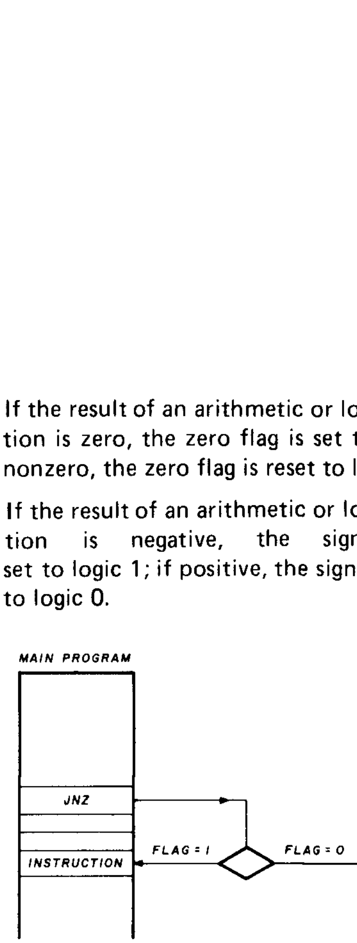


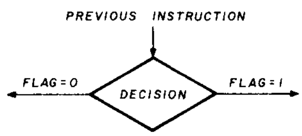
fig. 1. The JNZ instruction. If the zero flag is at logic 1, the instruction is ignored and program control passes to the following instruction.

Parity flag If the result of an arithmetic or logical operation has even parity, the parity flag is set to logic 1; if odd parity the parity flag is reset to logic 0.

Carry flag If the result of an arithmetic or rotate operation has a carry out of the most-significant bit of the 8-bit result, the carry flag is set to logic 1; if not, the carry flag is reset to logic 0. The carry flag is reset to logic 0 after all logical operations.

Auxiliary carry flag If the result of an arithmetic operation has a carry out of bit 3 into bit 4 of the 8-bit result, the auxiliary carry flag is set to logic 1; if not, the auxiliary carry flag is reset to logic 0. The auxiliary carry flag is reset to logic 0 after most logical operations.

Insufficient space is available in this column to discuss all of the above flags, so we shall restrict our attention to the zero flag. Shown below is the traditional flow chart *decision symbol* applied to an 8080 microprocessor decision:



The next instruction executed depends on the logic state of the flag associated with this specific decision. For example, consider the JNZ instruction, where JNZ means "Jump if Not Zero:"

instruction code	mnemonic	description
302	JNZ	If the zero flag is at logic 0, jump to the 16-bit
<B2>		memory address given in bytes <B2> and <B3>
<B3>		of this three-byte instruction; if the zero flag is at logic 1, ignore this instruction and proceed to the following instruction.

The statement "Jump if Not Zero" refers to the 8-bit result of a preceding instruction, not to the logic state of the zero flag. When this result is zero, the zero flag is set at logic 1 and program control passes to the next instruction, as shown in **fig. 1.**

The JNZ instruction is widely used in the creation of programmed *time delay loops,* an example of which is provided in **table 1.** In this program, both the address and instruction bytes are in octal code; it is assumed that the HI memory address byte is 000. The program first moves an 8-bit timing byte into register B; this byte, indicated by an asterisk, has any value between 000 and 377. The value of the byte will determine the duration of the time delay.

At LO memory address 002, a device-select pulse is generated to set the SN7474 flip-flop shown in **fig. 2.** The contents of register B are then decreased by 1.

table 1. Microcomputer program that demonstrates a simple time delay loop based on a decision made on the logic state of the zero flag. This program generates a single output pulse, the duration of which is determined by the timing byte at location 001, at the Q output of the SN7474 flip-flop.

LO memory address	instruction byte	mnemonic	clock cycles	description
000	006	MVI B	7	Move following timing byte into register B
001	*	—	—	Timing byte for register B
002	323	OUT 2	10	Generate device-select pulse that sets the SN-7474 flip-flop
003	002	—	—	Device code for set input to SN7474 flip-flop
004	005	DCR B	5	Decrement contents of register B by 1
005	302	JNZ	10	If zero flag is at logic 0, jump to the memory address given by the following two address bytes; otherwise, ignore this instruction
006	004	—	—	LO memory address byte
007	000	—	—	HI memory address byte
010	323	OUT 3	10	Generate device select pulse that clears the SN7474 flip-flop
011	003	—	—	Device code for clear input to SN-7474 flip-flop
012	166	HLT	7	Halt the microcomputer

*May have any value between 000 and 377. Its value determines time-delay duration.

The JNZ instruction immediately tests the logic state of the zero flag; if the contents of register B are not zero, the flag is at logic 0 and a jump occurs back to LO memory address 004. The DCR B and JNZ instructions are executed repeatedly until the contents of register B become zero, at which time the zero flag becomes logic 1. The JNZ instruction tests the flag for the last time and shifts program control to the OUT 3 instruction at LO memory address 010. This output instruction generates a device-select pulse that clears the SN7474 flip-flop. Once this has been done, the microcomputer comes to a halt.

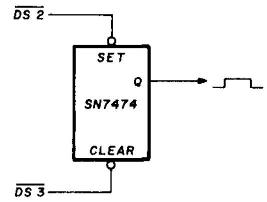


fig. 2. SN7474 flip-flop used as a monostable multivibrator.

The program shown in **table1** generates a single output pulse, the duration of which can take any value between 0.0125 and 1.925 ms in steps of 0.0075 ms. Some typical pulse widths are summarized in **table 2** for an 8080-based microcomputer that operates at a clock rate of 2 MHz. The calculations associated with the conversion of clock cycles to pulse width were discussed in reference 1. The number of clock cycles is a measure of the actual time it takes the microcomputer to execute a single instruction or group of instructions.

table 2. Examples of output pulse widths generated by the program in table 1 with an 8080 microcomputer operating at a clock rate of 2 MHz.

timing byte at LO memory address 001	number of clock cycles	pulse width (ms)
000	3850	1.925
001	25	0.0125
002	40	0.02
003	55	0.0275
004	70	0.035
005	85	0.0425
010	130	0.065
020	250	0.125
050	610	0.305
100	970	0.485
200	1930	0.965
300	2890	1.445
350	3490	1.745
377	3835	1.9175

For a 2-MHz microcomputer, a single clock cycle has a duration of 500 ns. The program in **table 1** and associated SN7474 flip-flop provide an example of what we mean by "the substitution of hardware by software" namely a simple program and a single flip-flop replace a much more complicated hardwired programmable monostable circuit.

reference

1. *Bugbook III. Microcomputer Interfacing Experiments Using the Mark 80 Microcomputer, an 8080 System,* (E&L Instruments, Inc., Derby, Conn., 1975. Available for $14.95 from Ham Radio Books, Greenville, N.H. 03048).

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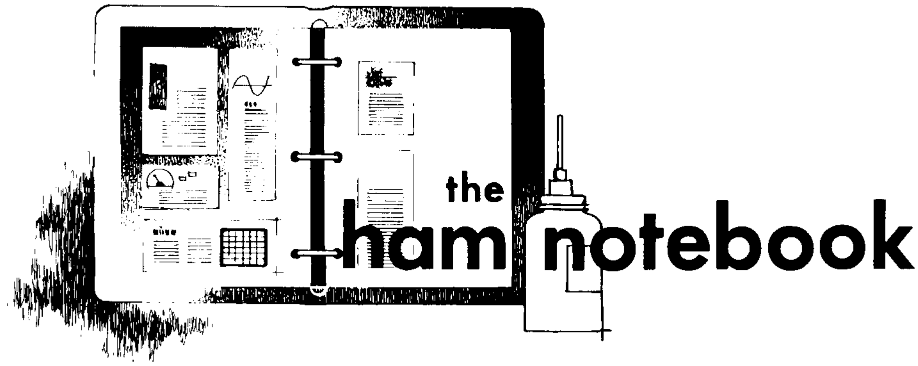

keyer modification

Many CW operators who use electronic keyers prefer a single-shaft paddle of the non-iambic variety, such as the *Vibro Key.* As they develop their speed level and go to close spacing on the paddle, many operators notice the effect of paddle bounce when using keyers with dot memory.

What happens is that after the operator strikes the paddle for a dash, the return motion of the paddle overshoots and causes the dot mechanism to close momentarily, which sets the dot memory. This, in turn, generates an unexpected (and unwanted) dot.

Careful adjustment of the paddle will minimize this effect, but it will still occur whenever the dash side is firmly hit. This is not, perhaps, a problem for the operator with a precise fist, but not all of us meet that description, and the extra bounced dot is very disturbing.

The cure is relatively simple and can be utilized with any TTL keyer. I have installed the circuit in **fig. 1** in both a Data Signal 21B keyer and a keyer built around the new Curtis keyer chip, and

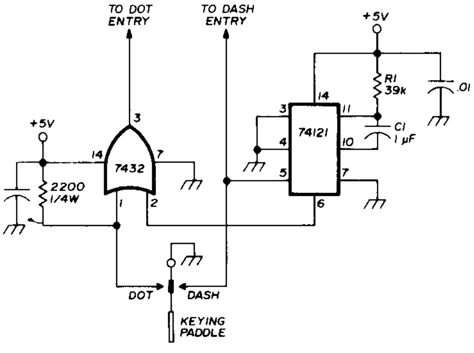

fig. 1. Simple circuit eliminates erroneous dots from being generated by paddle contact bounce (in keyers with dot memory).

in both cases the bounce problem was cured.

The circuit uses a 74121 monostable multivibrator and a 7432 AND gate. The output of the 74121 stays low if the paddle is not in use, if dots are being sent, or if dashes are being sent. However, as soon as the dash paddle is released, the transition from the low to high state causes the 74121 to transmit a high level pulse of short duration to the AND gate. The duration of the pulse is controlled by the values of R1 and C1.

When either of the inputs to the AND gate is in the high state, the output of the gate stays high so the paddle cannot transmit a dot into the dot memory. The duration of the pulse from the 74121 is selected so that it is only long enough to block a dot caused by the dash bounce from being placed in the dot memory. The duration of the pulse is short enough that the operator cannot possibly "reverse fields" with his hand fast enough to lose a dot he intentionally sends.

In fact, with the circuit installed, the only change the operator will notice is that he no longer sends erroneous dots which are caused by the key bounce. Installation in any TTL or CMOS keyer is very simple -- the keyed lines from the paddle are fed through the circuit and connection is made to the +5 volt line. (It should be noted that some CMOS keyers use voltages other than 5 volts, in which case this circuit will not work).

The values of R1 and C1 shown in the circuit were determined experimentally, and should work fine. If you notice any blocking of intentional dots, either R1 or C1 should be reduced in value until the problem disappears.

Bob Locher, W9KNI

Collins KWM-2/KWM-2A modifications

Over the years the Collins KWM-2 and KWM-2A ssb transceiver has undergone a number of modifications, some of which were made during the period the unit was used in military service. Available through MARS libraries, and possibly the Government Printing Office, is an Air Force Technical Manual that lists over 50 modifications to the KWM-2/KWM-2A along with expanded, fold-out diagrams of the circuitry which are a great improvement (over the amateur-style instruction manual) for the bifocal crowd.

Of interest to all KWM-2/KWM-2A owners is a simple modification that consists of adding a 0.01 µF, 400-volt capacitor from the screen (pin 8) of the 6EB8 audio amplifier to ground. This eliminates an ultrasonic oscillation that caused increased noise and audio distortion in some models.

The title of this technical manual is: *TO-31R2-4-183-3.KWM-2A Transceiver.* It also covers changes to the 30L-1 and 30S-1 rf amplifiers. A second technical manual of interest to KWM-2/KWM-2A owners is TO-31R2-4-183-2 entitled, *Technical Manual (Service) KWM-2A Transceiver.* It also covers the previously mentioned amplifiers. This publication provides detailed alignment instructions for the transceiver and linear amplifiers.

William I. Orr, W6SAI

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A convenient method for storing integrated circuits in your parts cabinets is to line the bottom of the drawers with 3/8 inch (1cm) of the polyfoam packing material used to ship electronic equipment. Push the IC leads into the soft material, keeping the ICs in neat rows and all facing the same direction. This way you can see at a glance which circuits you have in stock and keep them damage free.

Gary L. Tater, W3HUC

receiver incremental tuning for the Heath SB-102

A limited amount of receiver incremental tuning (RIT) may be obtained with the SB-102 quite easily. I own a unit with the transistorized linear master oscillator (LMO). At the rear of the LMO is a terminal marked FSK. Unless you are operating RTTY (which does not appear to be recommended in the SB-102 manual) with genuine FSK, this terminal is not used. However, it will provide up to a 1 kHz shift in frequency when directly grounded. By using the circuit shown in **fig. 2**, plus or minus 400-500 Hz shift may be obtained. As shown, the circuit provides a useable amount of RIT which is convenient for netting ssb signals. When used in conjunction with the optional 400 Hz CW filter, it really shines.

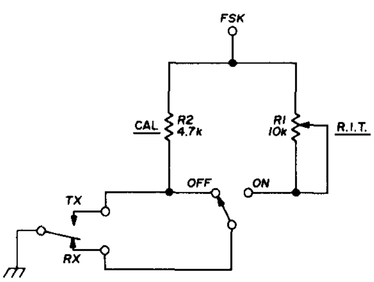

fig. 2. Receiver incremental tuning (RIT) circuit for the Heath SB-102 makes use of the built-in FSK circuit. R1 and R2 are added components.

The calibration resistor, R2, provides an essential mid-point setting and, with R1 centered, a beat note should be of equal pitch. A 10k multiturn pot could be used at R2 if extreme accuracy is desired. The dial calibration will shift approximately 400-500 Hz, but this is easily restored by the zero-set knob. The LMO shaft may be slipped slightly if you're finicky.

Paul K. Pagel, K1KXA

repairing R390 rf transformers

The rf transformers in the R390 and R390A receivers can be used for more than one band. The tuning ranges are: 0.5 to 1, 1 to 2, 2 to 4, 4 to 8, and 8 to 16 MHz. If there is a loss of sensitivity on some bands, or there is difficulty in obtaining proper alignment, here is a trouble to look for. It is more common in the R390 but may also show up in an R390A. In several first rf transformers I have found that the tuning core sticks because of lumps on the inside of the coil form. This is quite easy to check for.*

Look carefully at the rf tuning-slug racks as the mechanism tunes through its complete range on the band in question (or check them all as a precaution). The racks should move up and down smoothly. Check several times from different angles. Also check by pulling them up and down by hand while at the bottom of their range. Look for one end pushing up or causing the rack to deviate from the horizontal.

If you think there is trouble, it is easy to verify. Carefully remove the springs from each end of the slug rack and hang them out of the way under tension (use a bent paper clip). If you let go of the spring it can drop down inside the set and will be difficult to retrieve and reconnect. Lift up on the rack and it will come out quite easily. (When putting the rack back, work slowly, as it is easy to chip the edges of the coil forms when reinserting the tuning cores). When the rack has been removed, shine a light down inside the coil form and look at the side. Lumps show up immediately.

If you have this trouble it is easy to fix, but it must be done with care or the coil form will break. To remove the rf transformer, insert a Phillips screwdriver into the two little holes on the top of the transformer case and loosen the two captive screws. Then wiggle the transformer loose as you pull up. It may help to pry gently with another screwdriver.

*While the coils described are those for the Collins R390 and R390A series of receivers, there are many surplus and commercial receivers which use similar permeability-tuned mechanisms that might be susceptible to the same problem. The repair technique described here could be easily adapted to other, similar tuning mechanisms.

Don't back off on the captive screws any more than you have to. They can go past the point of releasing the transformer, come out of their mounting threads and rattle around loose inside the case. If this happens, take the top off the transformer and use a pair of needle nose pliers to hold the screws in position while you rethread them back.

To repair the coil form you have to remove the lumps from the inside. They appear to be bubbles of varnish or whatever finish was applied to the coil by the manufacturer. The first thing to do is to strengthen the coil form. To do this, spread several layers of *Elmer's Glue-all* on the outside of the form. Be sure each layer has plenty of time to dry; leave it overnight. This will give added strength to the form and coil and help keep either from breaking.

Next, go to work with your box of electric drill bits. Start with 13/64 inch (5mm). Gently insert it into the coil from and twist it by hand to begin removing the crud. When that cuts through, use a 7/32 inch (5.5mm) bit and do the same thing. Finish up with a 1/4 inch (6.5mm) bit. This will take most of it off.

Now make a tube of emery paper (fine sandpaper might work) long enough to reach to the bottom of the coil form and still leave a hand hold. Insert that into the coil form. Take a drill bit thin enough to slip easily inside emery paper but thick enough to give it support. Twist the emery paper around inside the coil form, moving it up and down at the same time. This will smooth off the inside again.

The thing to watch out for here is that you don't chip the top edge of the coil form. If it is chipped or looks about ready to go, it can be strengthened with a thin strip of typewriter paper, spread with *Elmer's Glue-all*, wrapped a few turns around the outside of the coil form top.

Every so often, remove the drill and emery paper and try the tuning core back inside. It should move up and down the entire length of the coil form without binding. When you have completed the operation, clean out the emery and coil form dust by blowing or using a pipe cleaner. Before you put the slug rack back in the set, give the inside of the coil forms and the tuning cores a squirt of silicon spray.

Alexander MacLean, WA2SUT

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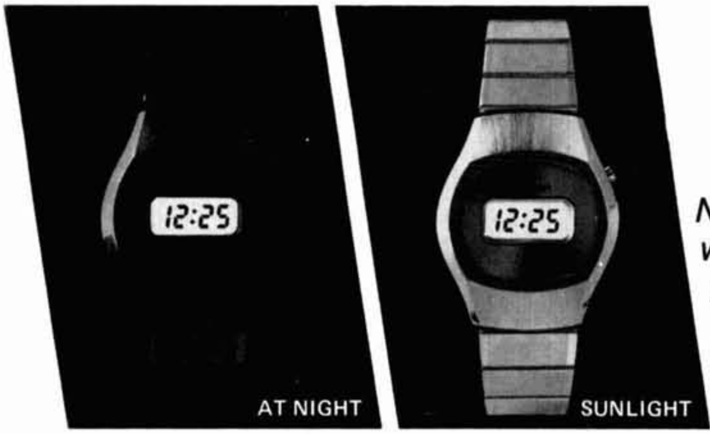

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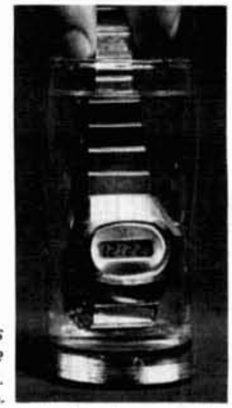

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5. Forget about changing technology The Sensor Laser 220 is so far ahead of every other watch in durability and technology that the watch you buy today, will still be years ahead of all others.

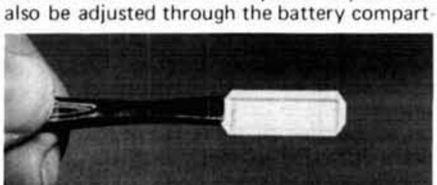
THE ULTIMATE ACHIEVEMENT

Other manufacturers have devised unique ways to produce a watch you can read at a glance. The new $300 LED Pulsar requires a snap of the wrist to turn on the display, but the Pulsar cannot be read in sunlight. The new $400 Longine's Gemini combines both an LED and liquid crystal display. (Press a button at night for the LED display, and view it easily in sunlight with the liquid crystal display.) But you must still press a button to read the time. All these applications of existing technology still fail to produce the ultimate digital watch: one you can read under all light conditions without using two hands. Until the introduction of the Sensor.

PLENTY OF ADVANCED FUNCTIONS

Sensor's five time functions give you everything you really need in a solid-state watch. Your watch displays the hours and minutes constantly, with no button to press. But depress the function button and the month and the date appear. Depress the button again and the seconds appear. To quickly set the time, insert a ball-point pen into the recessed time-control switch on the side. It's just that easy.

Sensor's accuracy is unparalleled. All solid-state digitals use a quartz crystal. So does the Sensor. But crystals change frequency from aging and shock. And to reset them, the watch case must be opened and an airtight seal broken which may affect the performance. In the Sensor, the crystal is first aged before it is installed, and secondly, it is actually cushioned in the case to absorb tremendous shock. The quartz crystal can also be adjusted through the battery compartment without opening the case. In short, your watch should be accurate to within 5 seconds per month and maintain that accuracy for years without adjustment and without ever opening the watch case.



The new exclusive laser-sealed tritium and phosphor light source is a thin solid-state tube that automatically illuminates the display when the lights dim.

STANDING BEHIND A PRODUCT

JS&A is America's largest single source of digital watches and other space-age products. We have selected the Sensor Laser 220 as the most advanced American-made, solid-state timepiece ever produced. And we put our company and its full resources behind that selection. JS&A will warranty the Sensor (even the batteries) for five full years. We'll even send you a loaner watch to use while your watch is being repaired should it ever require repair. And Sensor's advanced technology guarantees that your digital watch will be years ahead of any other watch at any price.

Wear the Laser 220 for one full month. If you are not convinced that it is the most rugged, precise, dependable and the finest quality solid-state digital watch **in the world**, return it for a prompt and courteous refund. We're just that proud of it.

To order your Sensor, credit card buyers may simply call our toll-free number below or mail us a check in the amount indicated below plus $2.50 for postage, insurance and handling. (Illinois residents add 5% sales tax.) We urge you, however, to act promptly and reserve your Laser 220 today.

Stainless steel w/leather strap $129.95
(Add $10 for matching metal band)

Gold plated w/leather strap. $149.95
(Add $10 for matching metal band)

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Your Next Purchase.

ELECTRONIC KEYER
MODEL 10B

Reed relay output (1 amp, 250V, 20VA). 10-30 WPM @ 6V-DC supply, 12 MA drain. 15-45 WPM @ 9V-DC supply, 15 MA drain. 3 MA idle current drain. Fixed spacing. Dots 1:1, Dash 1:3. Self-completing Dot/Dash. Manual dash in tune position. (Batteries not included.) Use the Model 10B Keyer with your paddle or our Model 11B matching paddle.

MODEL 10BWA KEYER with Sidetone assembled	$39.95
MODEL 10BW assembled	$29.95
MODEL 10BK (Kit)	$23.95
200-2K PC BOARD KIT	$14.95
200-3K SIDETONE KIT	$ 5.95
Ship. Wt. 1 Lb., add	$ 1.00

(PA RES. ADD 6% SALES TAX)

PADDLE
MODEL 11B

Dit/Dah travel adjustment. No mechanical switches. No bearings to fail. Paddle assembly weight is 1.5 pounds. Reversible Dit and Dah connections. Rubber feet. Damping on paddle operator lever. Feather glide paddle movement.

MODEL 11BW assembled	$11.95
MODEL 11BK (Kit)	$ 8.95
Ship. Wt. 2 Lb., add	$ 1.35

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CW TRANSMITTER
MODEL 50

15 watts input. Full break-in keying. All solid state. Crystal control. 160, 80 or 40M plug-in coil. Zener regulated chirpless keying. Has built-in 120 Vac power supply. OPTIONS: Built-in keyer and/or sidetone. Paddle Model 11B is compatible with built-in keyer option.

MODEL 50K (Kit)		$49.95
MODEL 50W (Wired)		$69.95
Add-on options:		
SIDETONE	200-21 Kit	$ 5.95
	200-21 Wired	$ 8.95
KEYER	200-22 Kit	$13.95
	200-22 Wired	$18.95
Ship. Wt. 4 Lb., add		$ 2.10

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(All units come with 40M plug-in coil unless otherwise specified. (Additional coil kits $3.95 each postpaid.)

ELECTRONIC KEYER WITH PADDLE

MORSE-1835
MODEL 12

C-MOS circuitry. Solid state output switch. (250V, 1 AMP MAX.) 8-45 WPM. Fixed spacing. Dot 1:1, Dash 1:3. Self-completing Dot/Dash. No on/off switch required. Sidetone has 2-inch speaker. Paddle travel adjustment. Rubber feet. 4 penlight batteries (not included).

MODEL 12 assembled	$49.95
Ship. Wt. 2 Lb., add	$ 1.35

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SPEECH PROCESSOR
MODEL 60A

200K/500 OHM inputs. PTT on connector. Instantaneous attack and release. 2, 9V-DC batteries (not included). 1.5 MA drain. Frequency is ±-1/2 db., 300-3000 Hz. Process gain control has an in/out switch. The process threshold is: 1.5 MV-RMS (HI-Z). 400 micro V-RMS (LO-Z). Output voltage 100 MV-RMS nom.

MODEL 60AW assembled	$29.95
MODEL 60AK (Kit)	$23.95
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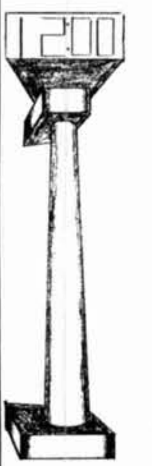
CLOCKS



6 digit AUTOMOTIVE CLOCK KIT complete with a CRYSTAL TIMEBASE accurate to .01 percent. 12 volts d.c. operation — built in noise suppression and voltage spike protection. Readouts blank when ignition is off — draws 25 mA in standby mode. Has .3 in. readouts. Use it in your car or for all applications where a battery-operated clock is needed. Approximate size 3" x 3.5" x 1.75"

WITH BLACK PLASTIC CASE	$34.95 ppd.
WITHOUT CASE	$29.95 ppd.
ASSEMBLED AND TESTED	$45.95 ppd.

CMOS CRYSTAL TIMEBASE KITS with .01 percent accuracy. 5-15 v.d.c. operation. Draws only 3 mA at 12 volts. Single I.C. — very small size — the P.C. board is 7/8" x 1-5/8". Choose a main output of 50 or 100 Hz., 60 Hz., 500 or 1000 Hz., or 1 Hz. Several related frequencies are also available on each board, in addition to the main ones listed above. Be sure to specify the Frequency you want. All kits are $10.95 ppd.

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ME-3 microminiature tone encoder

Compatible with all sub-audible tone systems such as: Private Line, Channel Guard, Quiet Channel, etc.

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- Excellent frequency accuracy and temperature stability
- Output level adjustment potentiometer
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Wired and tested, complete with K-1 element

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K-1 FIELD REPLACEABLE, PLUG-IN, FREQUENCY DETERMINING ELEMENTS
$3.00 each

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HOW'S YOUR BIRD??

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MODEL 43

$110.00

ELEMENT TABLE 1 SUFFIX H .... $40.00 ea.
SUFFIXES A, B, C, D, E .... $35.00 ea.
ALSO AVAILABLE . . . MODEL CC1, CARRY CASE FOR MODEL 43 .... $22.00
EC-1, CARRY CASE FOR XTRA ELEMENTS $14.00
ABOVE PRICES DO NOT INCLUDE SHIPPING.

PLUG-IN ELEMENTS for use with Model 43 THRULINE Wattmeter. Select one or more elements to suit your frequency and power ranges. When ordering, specify catalog number and THRULINE model number.

Table I

STANDARD ELEMENTS (CATALOG NUMBERS)

	Frequency Bands (MHz)					
Power Range	2-30	25-60	50-125	100-250	200-500	400-1000
5 watts	—	5A	5B	5C	5D	5E
10 watts	—	10A	10B	10C	10D	10E
25 watts	—	25A	25B	25C	25D	25E
50 watts	50H	50A	50B	50C	50D	50E
100 watts	100H	100A	100B	100C	100D	100E
250 watts	250H	250A	250B	250C	250D	250E
500 watts	500H	500A	500B	500C	500D	500E
1000 watts	1000H	1000A	1000B	1000C	1000D	1000E
2500 watts	2500H					
5000 watts	5000H					

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This very popular item at the 1976 Dayton Hamvention is **NOW OFFERED TO YOU** at these low prices. Incorporates the ideal "tactile feel" leaving no doubt that contact has been made. These **NEW** keyboards, manufactured by **THE DIGITRAN COMPANY,** are furnished with instructions for combining with a **MOSTEK** or **MOTOROLA** chip and a crystal (plus several small components) to become a Tone Encoder.

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- **Dual tones mixed internally no external mixing circuitry required.**
- **Constructed from CMOS for RF immunity**
- **Transmitter switching transistor on chip**
- **Completely compatible with Digitran keyboard**
- **Typ. 5 external components req'd (incl. crystal)**

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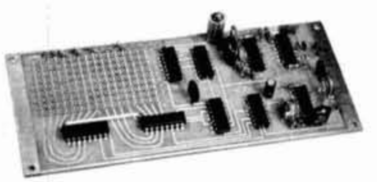
NEW products

programmable cw identification kit



With the new CW ID kit offered by VHF Engineering, you can build a complete identifier for commercial or amateur repeaters in about one evening. The CW ID kit uses high-grade components and comes complete with a drilled epoxy-glass circuit board and programming diodes.

Sufficient diodes are included to allow you to program virtually all repeater calls. To program, all you do is solder the appropriate diodes directly to a matrix on the PC board furnished. The diodes are mounted on the board in a straight-line fashion: three diodes for a dash, one for a dot, and none for a space. Programmed calls can be changed easily by rearranging the diodes. You can program the board for either CW or RTTY, which means added flexibility.

The CW ID kit is available for $39.95 plus postage; wired and tested it's $49.95 plus postage from the manufacturer. Drop a note to VHF Engineering, 320 Water Street, P.O. Box 1921, Binghamton, New York 13902 for more information, or use *check-off* on page 126.

vhf wideband preamplifier

Twin-output preamplifiers are something new offered by Spectrum International for the vhf buff. Designated MMa50, MMa144, and MMa220 (for the 50-, 144- and 220-MHz bands), these preamps feature two untuned stages with twin outputs for feeding two independent receivers. The preamps are built on a glass-epoxy G10 PC board, which is mounted in a standard die-cast aluminum box. Gain and noise figure are quite respectable as shown in the following table:

	MMa50	MMa144	MMa220
freq range, MHz	50-54	144-148	220-225
nominal gain, dB	20	16	15
noise figure, dB	2.5	2.8	3.4

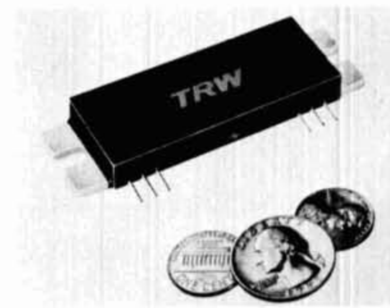
Power requirements are 12 volts dc at 20 milliamperes; size is 1¼ x 2½ x 4½ inches (32x64x114mm). The specifications apply to a 50-ohm input-output system. The MMa50 and MMa144 sell for $29.95 each; the MMa220 for $34.95. Add $1.00 shipping charge for each unit. Write Spectrum International, P.O Box 1084, Concord, Massachusetts 01742 for more information, or use *check-off* on page 126.

rf wattmeter

The trend today in radio transmitters is to have a wattmeter in the transmission line feeding the antenna. The model LPM-880, a new product by Leader Instruments Corporation, is a direct-reading wattmeter that measures radio-frequency power in the 0.5-120 watt range. Power range is selectable by front-panel pushbuttons. Also included is a dummy load for off-the-air power measurements. You can use the LPM-800 for measuring power loss in low-pass filters and coax cables as well as transmitter output. It's supplied with a sturdy tilt stand for easy reading. Input impedance is 50 ohms.

The LPM-880 is priced at $149.95. If you'd like more information contact Pat Redko, Leader Instruments Corporation, 151 Dupont Street, Plainview, Long Island, New York 11903 or use *check-off* on page 126.

power modules for mobile transmitters



Two new vhf rf power modules designed for mobile or marine transmitter applications are now available from TRW Semiconductors. The modules, designated MV20 and MV30, provide in excess of 20 watts and 30 watts output power respectively across the 140-175 MHz band. The modules operate from standard 12-volt automotive supplies and withstand infinite vswr at any angle, with 2 dB overdrive and 16 volts dc applied. The modules also feature 50-ohm input and output impedances, more than 20 dB gain, and are stable when operating into load vswrs as high as 5:1.

When compared to discrete component designs, these modules offer significant savings in size as well as cost of design, production, and repair. Small quantity pricing is $39.50 for the MV20 and $41.50 for the MV30. For further information, contact Sales Manager, Mobile Products, TRW RF Semiconductors, 14520 Aviation Blvd., Lawndale, California 90260 or use *check-off* on page 126.

loudspeaker for voice communications

The Kriket[R] model KC-55 speaker, new from Acoustic Fiber Sound Systems Incorporated, is designed to reproduce the human voice with maximum intelligibility. It has a frequency response between 80-10,000 Hz, essential for voice communications. The secret is in the Kriket[R] 5-inch-diameter (127mm) permanent-magnet speaker and an exclusive AFS Working Wall[R] enclosure, which controls sound by eliminating distortion. A snap-lock mounting bracket permits adjusting the speaker in any desired direction. The KC-55 speaker is packaged for base-station use, but you can adapt it for mobile use simply by removing the base for easy mounting inside a vehicle.

The KC-55 handles 7 watts rms of audio, has an input impedance of 8 ohms, and is furnished with a 6-foot (1.8m) connector cord and a standard miniplug. More information is available from acoustic Fiber Sound Systems, Incorporated, 2831 North Webster Avenue, Indianapolis, Indiana 46219, or use *check-off* on page 126.

digital multimeter

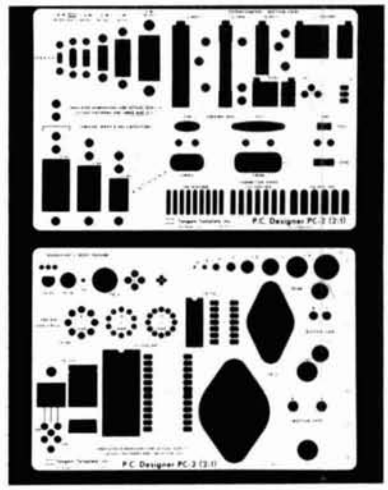
Non Linear Systems of Del Mar, California, announces a new addition to their Volksmeter family. It's the LM-3.5 Volksmeter Plus, a 3½-digit multimeter that fits into the palm of your hand. The 3½-digit feature means that the instrument reads out three digits plus 100 percent overrange. It's a true multifunction, multirange meter, rugged enough for field use yet useful for production or hobby work. Rechargeable nicad batteries and a 115-volt charger are standard equipment.

The LM-3.5 has four ranges for dc and ac volts, to 1000 volts dc or 1000 volts peak ac, with 1-millivolt resolution on the 2-volt scale. The resistance scale has one-ohm resolution and five ranges, from 2000 ohms to 20 megohms full scale. Ac and dc current can be measured in three ranges using shunts furnished. Automatic polarity is featured. Input impedance is 10 megohms on all voltage ranges. A large light-emitting diode display (0.3 inch or

7.6mm) and small package size of 1.9 x 2.7 x 4 inches (48x69x102mm) is a good example of what can be done with large-scale integration technology.

The LM-3.5 Volksmeter Plus retails for $147.00, including input leads, rechargeable nicads, battery charger and current shunts. Optional accessories, such as carrying case, high-voltage probe, desk stand, panel-mount flange, and universal test-lead set are also available. For information on these accessories as well as other data, write Non Linear Systems, Incorporated, P.O. Box N, Del Mar, California 92014, or use *check-off* on page 126.

pc design template



The new *PC Designer* template set has been updated to include additional, frequently used component packages and mounting patterns for printed-circuit layouts and assembly drawings. All patterns conform to guidelines established by Mil-Std 275C and The Institute of Printed Circuits bulletin CM-770. Component mounting patterns are on grid centers which enable the designer to design for automatic insertion assembly equipment.

Fixed and variable resistors, axial and radial lead capacitors, and several semiconductor packages are included. Template use reduces circuit-board design time by eliminating constant referral to manuals and data sheets for package dimensions. The template sets are available from stock in actual, twice, and four times size layout ratios and are priced from $12.00 to $20.00 per set. Write Tangent Template, Inc., Post Office Box 20704, San Diego, California or use *check-off* on page 126.

More Details? CHECK–OFF Page 126

multipurpose cw-operating aid

From SEA International comes the RST 599, an instrument that combines four features in one to enhance CW operation. Packaged in an attractive cabinet, the RST 599 includes a universal keying monitor, code-practice oscillator, CW filter, and a feature called the 599 function.

The RST 599 can be used with any key, keyer, transmitter or transceiver. The CW filter has a nominal center frequency of 850 Hz, with a nominal 3-dB bandwidth. The 599 function extracts and reconstructs the selected signal and, according to the manufacturer, "makes it noise free and RST 599 every time." Inputs are *audio in,* for connection to any receiver output line, and *key in,* for connection to any key or keyer. Outputs are *speaker out,* which drives any speaker of 4 ohms or more; *headphones,* for phones of 4 ohms or more and *transmitter,* which connects your key to any transmitter or receiver. When the 599 function is switched off, receiver output is connected directly to your speaker.

The RST 599 is priced in the $90.00 range and is fully warranteed for one year against material or manufacturing defects. If you'd like more information, write SEA International, P.O. Box 32, Milpitas, California 95035, or use *check-off* on page 126.

programmable scanning receiver

The Opti/Scan, a 10-channel scanning monitor receiver by SBE (Linear Systems, Incorporated), offers some unusual features that will appeal to the vhf enthusiast. Frequencies are digitally synthesized, which means you can forget about buying crystals to obtain desired coverage. You program the receiver yourself to scan channels of interest. Programming is easy. You simply refer to a code list supplied for desired scanning frequencies, program a plastic card according to instructions, insert the card into a slot on the receiver front panel, and internal circuits do the rest. No need to worry about programming errors – you can see what you've programmed and can check for accuracy. As many cards as desired may be programmed – up to 16,000 frequencies may be selected. Program cards are only $2.25 each.

Any ten channels are available between 30-50, 150-170, 450-470, and 490-510 MHz. Two-meter-band coverage (140-160 MHz) is available on special order. The Opti/Scan receiver is furnished with antennas that cover the vhf-uhf utility bands mentioned above. Receiver sensitivity is 0.5 microvolt for 12 dB SINAI. Dimensions are 2½ inches high, 7-3/4 inches wide, and 10 inches deep (6.4x19.7x25.4cm). Operating voltages are 13.8 Vdc or 115 Vac 60 Hz. The list price is $369.95. For further information contact Linear Systems, Incorporated, 220 Airport Boulevard, Watsonville, California 95075, or use *check-off* on page 126.

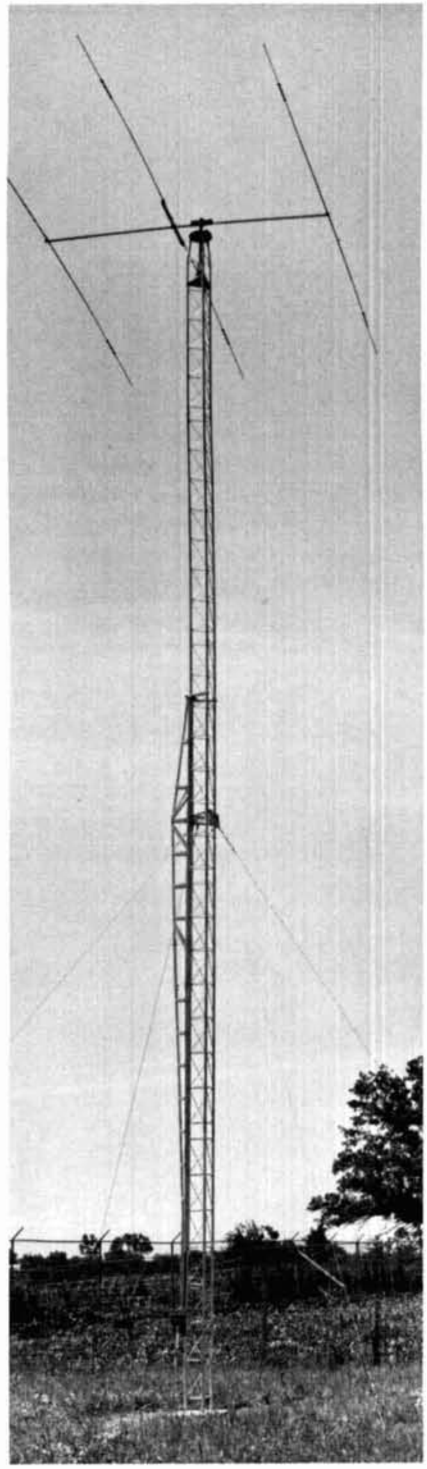
1976 Allied catalog

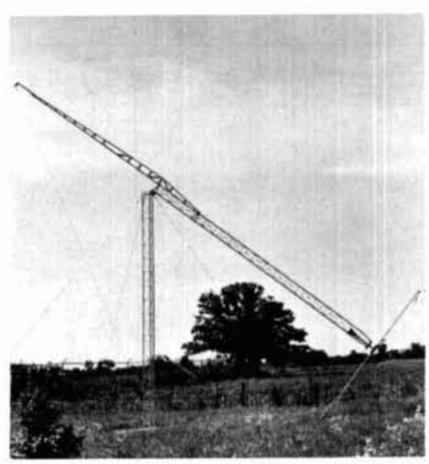
Now available from Allied Electronics is the 1976 *Engineering Manual and Purchasing Guide.* This up-to-date manual is a must for the service bench or engineering library. Engineers, radio amateurs, technicians, and hobbyists will appreciate the complete specifications, illustrations and information which describe each product.

This newest edition of the well-known Allied Guide offers 228 pages of high-quality electronic parts and equipment from Allied and other leading manufacturers. You can choose from a wide variety of new products, in addition to traditional items which have set in the industry's standards. The guide contains cable, solid-state devices, test equipment, connectors, relays, tools, capacitors, and countless other electronic parts for virtually any application. Allied offers bulk pricing for quantity buyers, and nationwide warehouses assure prompt delivery of these often hard-to-get items.

A $5 value for **only $1** to help cover postage and handling. Contact Allied Electronics, Dept. 76, 401 East 8th Street, Fort Worth, Texas 76102 or use *check-off* on page 126.

More Details? CHECK–OFF Page 126





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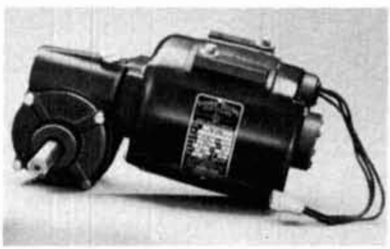

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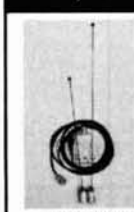

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More Details? CHECK—OFF Page 126

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Small: only 8″x2.6″x1.28″ . . . **Appearance:** slim silhouette all black metal . . . **Serviceable:** easy access to separate receive and transmit circuit boards . . . **PLUS:** 6 pole xtal filter for superlative receiver operation . . . **and:** trimmers on receive and transmit xtals: standard 10.7 MHz 1st IF.

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BARREL KIT #26 PLASTIC TRANSISTORS 100 for $1.98 Untested. Type TO-92 (TO-18), all manufacturers, variety of 2N #'s. Cat.No. 8H 2604

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BARREL KIT #32 TRANSISTORS WITH A HOLE IN IT 50 for $1.98 Cat.No. 8H 2610 Untested. Can't name factory but we bought barrels of 25 watters with mtg. hole in middle. PNP'S and NPN'S.

BARREL KIT #35 NEON LAMPS 40 for $1.98 100% good. Famous NE-2's. All prime, but factory made millions and barrel'ed 'em. Your advantage. Cat.No. 8H 2613

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BARREL KIT #37 1 AMP "BULLETT" RECTIFIERS 100 for $1.98 Untested. Famous style, asstd. voltages, silicon, axial includes all types of voltages to 1KV. Cat.No. 8H 2615

BARREL KIT #39 2N3055 HOBBY TRANSISTORS 15 for $1.98 100% good. From factory to you, these fallouts of the famous 2N3055. We have 10 barrels. Cat.No. 8H2617

BARREL KIT #40 PNP HIGH-POWER TRANSISTORS 20 for $1.98 Popular germanium TO-3 case units, now available at "good ole barrel" prices. Cat.No. 8H 2618 100% good.

BARREL KIT #46 G.E. 3.5 WATT AMPLIFIERS 25 for $1.98 Untested. Hobby type, factory fallouts, we purchased them in barrels. These are unknowns. Cat.No. 8H 2624

BARREL KIT #50 SIGNAL SILICON DIODES 200 for $1.98 Includes many, many types of switching, signal silicon types, all axial leads. Some may be zeners. Cat.No. 8H 2628 Untested.

BARREL KIT #53 JUMBO RESISTOR PAK 100-pc. $1.98 Cat.No. 8H 2721 Assortm metal films, precisions, carbons, metal oxide powers, from 1/8 watt to 7 watts. Color coded & 100% good. Worth $10.

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BARREL KIT #59 POWER TRANSISTORS 40 for $1.98 15 watt Bendix B-5000 pellet transistors, npn, all good, purchased from a pretester, have millions of 'em. 100% good Cat.No. 8H 2727

BARREL KIT #60 DTL'S IC'S 75 for $1.98 Untested. This is prime barrel material. Who wants DTL's? 930, 936, 946's. Your gain is our loss. They're marked too. Cat.No. 8H 2728

BARREL KIT #61 POLYSTYRENE CAPS 100 for $1.98 Finest caps made. As a gamble we bought 10 barrels from factory, mixed values; all good. Cat.No. 8H 2729

BARREL KIT #65 MIXED READOUTS 15 for $1.98 Factory returns — such numbers as MAN-4's, MAN-7's, MAN-3's, 11 barrels & no time to separate. Cat.No. 8H 2733 Untested.

BARREL KIT #68 2 WATTERS 100 for $1.98 100% good. Nobody seems to want 'em! So many suppliers don't count, but throw 'em in the barrel. It's a li'l gold mine. All marked. Cat.No. 8H2735

BARREL KIT #71 CAPACITOR SPECIAL 100 pcs. $1.98 Emptied stockrooms into barrels of mylars, poly's, micas, moldeds, plastics, ceramics, discs, etc. Nifty 100% good. Cat.No. 8H2738

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BARREL KIT #75 400MW ZENERS 150 for $1.98 Factory out of biz! Amazing offer: 6, 8, 10, 12 to 15V. You test. Hermetically sealed glass pak. Double plug. Cat.No. 8H 2740

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BARREL KIT #77 "BROWN" BODY TRANSISTORS 40 for $1.98 G-E D-40 series: has hi-voltage, Darlingtons, hi-current, npn's. Factory line discontinued. Power tabs. Cat.No. 8H 2742 Untested.

BARREL KIT #78 "RED" BODY TRANSISTORS 40 for $1.98 D-42 series. You test—go into your own biz! High current, hi-V. NPN Cat.No. 8H 2743 Untested

BARREL KIT #81 SUBMINI RESISTORS 200 for $1.98 100% good. PC, upright type, color coded, 1/8 watt. Asst values. Came to us in a barrel. Cat.No. 8H 2746

BARREL KIT #83 15 for $1.98 LM-340T VOLTAGE REGULATORS Untested. Factory rejected them for length of leads. May include 5, 6, 8, 12, 15, 18, 24 volts. Power tab Cat.No. 8H 2635

BARREL KIT #86 HOBBY LEDS 40 for $1.98 Untested. Wow! A Litronics dump of all kinds of mixed discrete LEDS, shapes, colors, good, poor, etc. 8H 2859

BARREL KIT #87 NATIONAL IC BONANZA 100 for $1.98 Factory dumps into barrels. Types 8000, 7400 series, DTLs, ROMs, registers, clock & calc. chips, linears, etc. Cat.No. 8H 2860 Untested.

BARREL KIT #88 LITRONICS LED READOUTS 10 for $1.98 747's, 727's, singles, triples, etc. .33 to 0.6. Bot from factory, all mixed; have fun! No. 8H 2861

BARREL #91 SILVER MICAS 100 for $1.98 Cat.No. 8H 3018 For the first time silver micas so low in price! Axial, red case, variety of physical sizes & values. Big savings from distributor prices. Wt. 1 lb.

BARREL KIT #93 HALF WATTERS 200 for $1.98 Untested. Resistor factory tried to fool us by mixing 100% color-coded resistors in barrel. But value is there. 4 oz. Cat. No. 8H3046

BARREL KIT #99 PHOTO ELECTRIC CELLS 10 for $1.98 Asst. GE types, CDS types. Mixed by factory. Big job for us to separate. 100% good. Cat.No. 8H 3052

BARREL KIT #101 RESISTOR SPECIAL 200 for $1.98 Includes: 1/8, 1/4, 1/2, 1, 2-watters, carbon. 8 oz. 100% good. 8H3054

BARREL KIT #104 SLIDE VOLUME CONTROLS 10 for $1.98 Cat.No. 8H3057 Used in hi-fi, volume control maker unloads. Asst. values, what a buy. Worth $1 ea. We've got barrels of 100% material.

BARREL KIT #107 SQUARE OHMS 60 for $1.98 Cat.No. 8H 3096 Factory people are sometimes "squares" when they topple prime square ohms, mix 'em up in barrels. Asst. values watts. Wt. 1 lb.

BARREL KIT #108 TO-5 PLASTIC TRANSISTORS Cat.No. 8H 3101 **40 for $1.98** Includes PNP, NPN, 2N-3638, 2N3641, 2N5000 series, etc. Untested, but guaranteed to a 60% yield.

BARREL KIT #109 TERMINAL STRIPS 150 for $1.98 Wide asst. of terminal strip connectors, from 1 contact up. Strip manufacturers barrel dump is your gain. Wt. 1 lb. Cat.No 8H3136

BARREL KIT #110 SUPPRESSOR DIODES 50 for $1.98 Cat.No. 8H3137 Keeps ignition noises out axial. Untested, but the of your eqpt., car, industrial, etc. Double plug.

BARREL KIT #112 MICRO MINI LEDS 40 for $1.98 All the tiny leds, axial, upright of Monsanto, Litronix, variety of colors. Yield 60% or better. Cat.No. 8H 3139

BARREL KIT #134 CALCULATOR CHIPS 15 for $1.98 National type. Can be MM5736, 38, etc. They have discontinued these chips. FIND A USE. Untested. Cat. No. 8H 3258

BARREL KIT #115 MOLEX SOCKETS 200 for $1.98 100% good. Calculator maker dump! We got a zillion of 'em. Used for IC sockets, etc. Cat.No. 8H 3144

BARREL KIT #116 BUTTONS 'N FEEDTHRU'S 100 for $1.98 100% good. Truthfully worth a small fortune. Wide asst. of button-feedthru caps! HAMS TAKE NOTE! RF, UHF, etc. Wt. 1 lb. Cat.No. 8H3141

BARREL KIT #118 MINI SCRS 50 for $1.98 UNBELIEVABLE! TO-92 plastic SCRS in barrels . . . rite from factory. Includes all voltages up thru 200 prv. 8H3135

BARREL KIT #123 CD-4002 C-MOS IC 15 for $1.98 mostly good. But we have 250,000. Can never sell 'em out. YOUR GAIN! Their number is CD-5602. Cat. No. 8H3217

BARREL KIT #126 UPRIGHT ELECTROS 40 for $1.98 Wide asst. of values from 1mf to 300mf in mixture of voltages. 100% marked 'n good. Why barreled? U-got-a-buy! Cat. No. 8H3226

BARREL KIT #127 AXIAL ELECTROS 40 for $1.98 Truthfully the factories (by mixing 'em in barrels) do all of us a favor. WUT A BUY! Asst. capacities and voltages. Cat. No. 8H3227

BARREL KIT #128 MINI DIP ICS 75 for $1.98 Large mfgr dumped 100's of lbs into barrels. Includes 741s, LM-380-8, 703, 567, 555, 558—but who knows? Factory to you. All mixed, you test. Wt. 1 lb. Cat. No. 8H3245

BARREL KIT #131 TANTALUM ELECTROS 30 for $1.98 Mixed, marked prime, top grade asst. values, voltages. GE, Centralab, etc. Cat. No. 8H3255

BARREL KIT #133 C-MOS IC'S 60 for $1.98 Deliberately thrown in barrels, so we can't test 'em! The famous CD4000series. How good? Who knows?, Who cares? It's only 3c ea. Cat. No. 8H3257

BARREL KIT #138 PANEL SWITCHES 30 for $1.98 Did you hear of OAK? Another eqpt maker barrelled all types of rotaries, electric, slides, etc. 8H 3268

BARREL KIT #140 LAMPS 20 for $1.98 Precision, mini lamp all-metal, panel, with mtg hardware. Lamp maker's overstock. 8H3297

BARREL KIT #141 10 WATT ZENERS 15 for $1.98 Mfr dumps to prepare for new styling, voltages all over the place. Good yield. Cat. No. 8H3298

BARREL KIT #142 DARLINGTON TRANSISTORS 40 for $1.98 TO-92, a Motorola dump, unknown numbers, but high yield to good darlingtons. Retesters didn't get 'em! You will. 8H3285

BARREL KIT #144 RCA PHONO PLUGS 40 for $1.98 1,000,000 RCA phono plugs for this one. You hi-fi-ers know wut they are . . . 100% material. Look at the price. 8H3293

Terms: Add postage Rated: net 30
Phone: Wakefield, Mass. (617) 245-3829
Retail: 16-18 Del Carmine St., Wakefield,
MINIMUM ORDER — $6.00

C.O.D.'s MAY BE PHONED

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P.O. BOX 942H LYNNFIELD, MASS. 01940

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More Details? CHECK—OFF Page 126

DRAKE EXPANDS SERVICE FACILITIES

Effective August 1, 1976 warranty and non-warranty customer service as well as parts for R. L. Drake Co. Amateur Radio Products will be available from the R. L. Drake Co. Western Region Sales & Service Center, 2020 Western St., Las Vegas, Nevada 89106.

William Gaible, WA8ADI, the Manager of this new Drake Company Service Center can be contacted by calling 702-382-9470.

This Western Service Center was established in order to better serve the needs of R. L. Drake Co. customers in the states of:

- Hawaii
- Alaska
- Washington
- Oregon
- California
- Montana
- Idaho
- Wyoming
- Nevada
- Utah
- Colorado
- Arizona
- New Mexico

as well as:

- Western Canada
- Mexico

The office will be staffed with experienced R. L. Drake factory Customer Service employees.

flea market

RATES Non-commercial ads 10¢ per word; commercial ads 40¢ per word both payable in advance. No cash discounts or agency commissions allowed.

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

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

DEADLINE 15th of second preceding month.

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

BUY — SELL — TRADE. Write for free mailer. Give name, address, call letters. Complete stock of major brands. New and reconditioned equipment. Call us for best deals. We buy Collins, Drake, Swan, etc., SSB & FM. Associated Radio, 8012 Conser, Overland Park, Ks. 66204. 913-381-5901.

STOP don't junk that television set. ASE manufactures the world's most complete line of television picture tubes. Over 1700 types. Most types immediate delivery. Tubes for Old or New TV's, black & white and color. 2 year factory warranty. Lowest prices anywhere. Allied Sales & Engineering, Inc., Dept. 22, Pimento, IN 47866. Telephone 812-495-6555.

COMPUTER POWER SUPPLY, +24V/27V/3A +12V/350ma, +5V/5A, —12V/500ma. Compact, extraordinary regulated and filtered. Pulled from new computer. With schematic. $25.00. Many Cartrivision VTR parts. SASE. W6DOM, 6017 Majorca Ct., San Jose, Calif. 95120.

WYOMING RANCH LAND. Wild horses, antelope, deer, elk. 10 acres $30 down, $30 month. Owner - K6ICS, Mike Gauthier, 9550K Gallatin, Downey, CA. 90240.

WANTED. Only 1 copy Ham Radio, May, 1973. Tom Ierlan, WA2SPM, 725 Proctor Ave., Ogdensburg, N. Y. 13669.

VERY in-ter-est-ing! Next 4 big issues $1. "The Ham Trader," Sycamore, IL 60178.

CUSTOM EMBROIDERED EMBLEMS, your design, low minimum. Emblems, Dept. 709, Littleton, New Hampshire 03561.

RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, Ca. 94806.

WANTED: Tektronix types P80 probe, 81 adapter, and 82 plug-in. State price. Mark Hudson, 1218 E. 36th St., Odessa, Tx. 79762.

MOBILE IGNITION SHIELDING provides more range with no noise. Available most engines. Many other suppression accessories. Literature, Estes Engineering, 930 Marine Dr., Port Angeles, WA 93862.

NEW POWERFUL THREE OUTPUT REGULATED POWER SUPPLY, plus 900 parts worth $400.00 list. Solid state TV recorder electronic unit. Schematics, parts cross reference. Free brochure. $17.95 plus $3.50 S&H, USA. Master Charge, BankAmericard. Satisfaction guaranteed. Madison Electronics Company, Incorporated, Box 369, D77, Madison, Alabama 35758.

AMATEUR CALL SIGN PINS $1.00. 1″ x 3″. 1 or 2 lines engraved. Blue, black, red, green. Guaranteed. Holly, Box 3926-S, Hollywood, Florida 33023.

FREQUENCY COUNTER BOARDS, Jan. 76 HR, double sided glass epoxy includes 500 MHz prescaler circuitry and LED board, instructions and parts source listing $15.00 complete. Popular electronic keyer board $2.00. CSJ Electronics, 5201 Cameron Court, Lincoln, NE 68512.

GOVERNMENT SURPLUS Communication Equipment, FREE catalog. Colonel Russell, 9410 Walhampton, Louisville, KY 40222.

WANTED — Housekeeper YL, (1KW Rig) Suburban-W2ISL, A. Porterfield P.E., 41 Winnebago Rd., Yonkers, N. Y. 10710.

KLM PRODUCTS, Larsen ants., Icom, police and fire scanners. Send for prices. Not given over phone. Narwid Electronics, 61 Bellot Road, Ringwood, N. J. 07456.

HOMEBREWERS: Stamp brings list of high quality components. CPO Surplus, Box 189, Braintree, Mass. 02184.

SAVE! Bomar FM, xtals $4.00 ppd. Dentron, Hustler, CushCraft, W. M. Nye, Ameco. Used gear. Complete catalog - write Ferris Radio, 308 E. Harry, Hazel Park, Mich. 48030.

FERRITE BEADS: Ferroxcube beads w/specification and application sheet — 10 @ $1.00 postpaid. Includes latest catalog. CPO Surplus, Box 189, Braintree, Mass. 02184.

COMPUTER HOBBYISTS! Bargain hunt and sell via ON_LINE. 18 issues/year - $3.75. Free sample issue from: ON_LINE, 24695 Santa Cruz Hwy., Los Gatos, CA 95030.

COLLINS 75S1, 32S1, 516F-2, $765. General Radio Type 916-A radio frequency bridge, $245. Heath SB-200, $215. Rohn No. 25, new-never used, complete - 90 ft., my cost $635 - now $450. Copper jacket, low loss 50 ohm coax 280 feet $95. Ham-M rotor complete $100. 4 el. 10 meter beam HB (20′ boom) $40. TA-33Jr. tri-band beam $70. Small prop-pitch rotor - items for complete system $175. "Large" prop-pitch rotor $75. QST, 29 years (328 issues) $90. CQ, 28 years (305 issues) $65. 73 Magazine, 14 yrs. (167) issues) $30. U ship. Moving to W6. M. M. Kovar, W2ZN, 3 Puddingstone Ct., Morristown, N. J. 07960. (201) 386-2512 day - (201) 267-0657 eve.

MANUFACTURERS, DISTRIBUTORS, DEALERS! The Memphis Hamfest had 3,500 registrations last year — even more expected this year! Saturday and Sunday, October 2 & 3, at State Technical Institute, Interstate 40 at Macon Road. Security, motels, restaurants — a great location for a great event! Contact Harry Simpson W4SCF, Box 27015, Memphis, TN 38127. Telephone 901 358-5707.

TECH MANUALS for Govt. surplus gear — $6.50 each: SP-600JX, URM-25D, OS-8A/U, PRC-8, 9, 10, Thousands more available. Send 50¢ (coin) for 22-page list. W3IHD, 7218 Roanne Drive, Washington, D. C. 20021.

RUBBER STAMP, name/call/QTH $2.50 ppd. (CA residents add tax). LWM Press, Box 22161, San Diego, CA 92122.

CANADIAN JUMBO SURPLUS and Parts Catalogs. Bargains Galore. Send $1. ETCO-HR, Box 741, Montreal "A" H3c 2V2.

WANT Wireless book; daughter, Dr. Mahlon Loomis, 1920's. W9LL.

PORTA-PAK the accessory that makes your mobile really portable. $59.95 and $39.95. Dealer inquiries invited. P. O. Box 67, Somers, Wisc. 53171.

DESKFAX PAPER $10 per 1000 sheets, Tom White. 3325 W. Altadena, Phoenix, AZ 85029.

DIRECT CONVERSION RECEIVER KITS for AM or CW. Write WB9MBH, 3132 North Lowell Avenue, Chicago, Illinois 60641.

WANTED: HRO-50 in repairable condition. WA5INP, Art Smith, 5446 Grafton St., Jackson, Mississippi 39206.

FREE Electronics Surplus Catalog. Electronic Specialties, 1659 Wetmore, Tucson, AZ 85705.

H-P 400D $45.00, H-P 650A $75.00, good cond., manuals. FOB Ray Harland, 2602 Mary Ln., Escondido, Ca. 92025.

MOTOROLA CONVERTA-COM CONSOLES. One each, VHF & UHF. Complete with speakers and cables. Make offer. Mike, WA2ZOW, 65 Richard St., Clark, N. J. 07066. (201) 382-0879.

SIDESWIPER only $13. Airmailed USA. Kungsimport, Box 257, Kungsbacka, Sweden.

CIRCUIT BOARDS. Artwork, negatives, etching. SASE for details. Karl Raup, WB4OXG, 630 Albertson Place, Orlando, Fla. 32806.

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American made, 115V Primarys:
6.3 V, 1 Amp, Shielded $1.75 Ea. ppd.
12V — 250 mils — for P.C. Board. Size: 1½" x 1⅛" x 1⅛", 3.5 OZ. $1.40 ea. ppd.

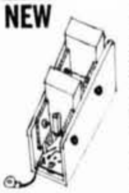
12 volt ½ amp. $1.68 ppd.
12V 1.2 Amp $2.57 ppd.
12V, 3 Amp $4.15 ppd.

NEW NEW NEW

Dual Secondarys; 18-0-18V @ 1.2 Amp and 7-0-7V @ 400 MA, 1 lb. 5 oz., Shielded. 2⅝" x 2" x 2¼" (3¼" MTG) — $3.85 ppd.
22-0-22V — 1 Amp, 6.3V. Tap ¼ Amp — 2" x 2¼" x 2⅜" (MTG 2⅜") Leads from top of shield, 1 lb. 4 oz. $3.20 ppd.
24-0-24 @ 1 amp with tap at 6.3 volt for pilot light. $3.15 ea. ppd.

TRANSISTOR OUTPUT TRANSFORMER — One watt rating 200 ohm Centertapped primary, 8 ohm secondary 85¢ Each ppd.

NEW



Transistor and Relay Assembly — consists of (2) MJE3055 and (2) MJE2955 transistors (10 amp, 90w, 60v complementary pairs) mounted in "U" channel heat sink 2½" x 1⅛" x 1½". (2) XTAL CAN RELAYS, DPDT, 28v, 8000Ω, 5.8 ma DC, 1 amp contacts mounted on PC board with resistors. $2.85 ea. ppd.

NEW SIZES — VERTICAL MOUNT PC BOARD POTENTIOMETERS

American made (CRL) Cermet sizes: 25K, 100K ohms. 5/$1.30 ppd.
CTS Blue wheel. Values: 750, 1K, 1.5K, 50K, 300K ohms. 5/$1.20 ppd.

NEW MINI Imported vertical pots (½H, ⅜W, 1/16D) Values: 1000, 2.5K, 5K, 50K ohms. Slot adj. Price: 5 for $1.20 ppd.

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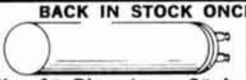
BACK IN STOCK ONCE AGAIN



3000 MFD @ 30 Volt Capacitors.
Size 1" Diameter x 3" Long. 90¢ Each or 3 For $2.25 ppd.

3000 MFD @ 20V Capacitors. Same size as above. 80¢ ea. or 3 for $2.00 ppd.

ALSO 3000 MFD @ 50V 95¢ ea. or 3/$2.65 ppd

ELECTROLYTIC CAPACITOR — PHILCO

Quad section 100x150 MFD @ 400V and at 350V D.C. and 20x50 MFD @ 250V — TWISTAB MTG. A very nice unit for Transceiver Power Supplies etc.
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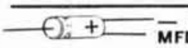
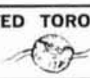
DUAL Electrolytic 1000 & 500 MFD. 15V, long leads. ¾" dia. x 2¼" long. 55¢ each 3/$1.50 ppd.



— Dual Electrolytic 1000 MFD @ 16V, 500 MFD @ 12V. ¾" dia. x 2¼" long. 60¢ ea. or 3/$1.55 ppd.

1¼ inch miniature alligator clips. Bright vinyl red or black. Nickle plated 9 for $1.00

UNPOTTED TOROIDS — center tapped 88MHY 5/$2.95 Limited Qtys 44 MHY 5/$3.95 ppd.



CRL DISC CAPACITORS
.1 MFD, 10V ⅜" dia., long leads. 10 for $1.00 ppd.

Red 6 foot #18-2 line cord with flanged safety plug — 55¢ ea. or 3/$1.45 ppd. Gray 6 foot, 7A, plugs into auto liter socket — 95¢ ea. or 3/$2.55

General Purpose Germanium Diodes. Similar to 1N34a etc. All Cathode banded Full leads. 16 for $1.00 ppd.

Small, glass, silicon diodes. Fairchild FD-type. Full leads and guaranteed. Price: 16 for $1.00 ppd.

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NEW CANADIAN MAGAZINE. "Electronics Work Shop". $5.00 yearly, sample $1.00. ETCOB, Box 741, Montreal, H3C 2V2.

FREE CATALOG. LEDS, strobe lights, UARTS, memories, RF transistors, microphones, IC's, relays, ultrasonic devices, precision trimmer capacitors, digital thermometers, unique components. Chaney's, Box 15431, Lakewood, Colo. 80215.

FOR SALE: Genave GTX-200T, factory tone pad, 4 channels, mint condx. $240. Heathkit SB-102 A-1 shape, make offer. WAØNZQ, Curtis R. Olson, P. O. Box 215, Regent, N. Dak. 58650.

WANTED: Boehme, Creed, McElroy, G.N.T., Wheatsone equipment. ATKO Minikeyer, ATKO, ECCO, Navy 8M3 training set. Manuals booklets. Fisher, 235 Adams Street, Brooklyn, New York 11201.

MODERN 60 MIN. CODE CASSETTES. Novice 0-5 wpm, Progressive 5-13 wpm, General 13-15 wpm, Extra 20-22 wpm. $3 each, 4/$10. Royal, Box 2174, Sandusky, Ohio 44870.

QRP TRANSMATCH for HW7, Ten-Tec, and others. Send stamp for details to Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

COMPLETE LINE KLM, CushCraft, Covercraft dust covers, SCS amplifiers, Regency, Triex Towers. Call or write Radios Unlimited, 86 Balch Ave., Piscataway, N. J. 08854. 201-762-4307.

NAMEBADGES $1.25, name and call sign $1.75. Engraved plastic with pin or clutches. Black, white, red, blue, green, woodgrain. Include payment with order. Club emblem and hamfest badges. SASE for catalog. Donan's Engraving, P. O. Box 07155, Lakewood, Ohio 44107.

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MANUALS for most ham gear, 1939/70. List $1.00. Send SASE (or 25¢) for one specific model quote. Hobby Industry, WØJJK, Box H864, Council Bluffs, Iowa 51501.

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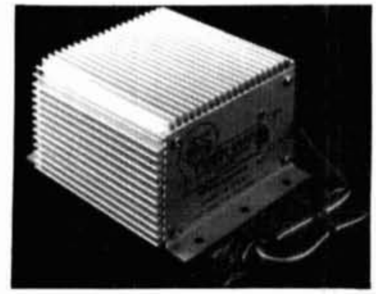
TELL YOUR FRIENDS about the BIG NEW Ham Radio Magazine!



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No rewiring necessary. Engine cannot be damaged by improper installation. Either of three models fits any vehicle or stationary engine with 12 volt negative ground, alternator or generator system. Uses standard coil & distributor now on your engine. Dual switch permits motor work or tune-up with any standard test equipment.

Write for free booklet that not only is the BEST description of CDIs, but also explains the need for such a system. Current prices assured til July 1, '76.

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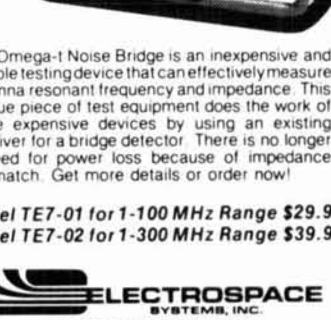
test for resonant resistance with an omega-t antenna noise bridge



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Model TE7-01 for 1-100 MHz Range $29.95
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320 TERRACE VILLAGE
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TELEPHONE (214) 231-9303

Sold at Amateur Radio Dealers or Direct from Electrospace Systems, Inc.

More Details? CHECK—OFF Page 126

Get more operating pleasure with your Atlas Transceiver!

Introducing the new

ATLAS Auxiliary VFO

The Model 206 provides:

- Digital Dial L.E.D. Readout
- Separate control of transmit and receive frequency
- Expanded frequency coverage
- Frequency Counter Capability
- Digital Hold

The Model 206 Auxiliary VFO is designed for simple plug-in installation with all Atlas Transceivers providing increased versatility, greater frequency coverage including full coverage of the 10 meter band, and more flexible operating capability.

DIGITAL DIAL READOUT

All L.E.D. Matrix 6 digit display reads within 100Hz (just 1/10 kHz) of your actual operating frequency. Reads on both Receive and Transmit.

A DIGITAL HOLD feature permits retention of a frequency reading while tuning to other frequencies.

SEPARATE TRANSMIT-RECEIVE FREQUENCY CONTROL

The 206 provides for transmitting on one frequency while receiving on another frequency. Transceive operation on the same frequency may be with either the 206 or 210x/215x.

EXPANDED FREQUENCY COVERAGE

Complete coverage from 1.8 to 3 MHz (with Model 215x only), also 3 to 5, 6 to 8, 8 to 10, 14 to 16, 20 to 22, and with the 210x from 28 to 30 MHz. This permits MARS operation as well as reception over large segments of the HF spectrum.

FREQUENCY COUNTER CAPABILITY

In addition to being a VFO Digital Dial, the 206 will also function as a sensitive frequency counter from 100 Hz to 40 MHz for general use around the ham shack or lab.

The Model 206 is housed in an attractive cabinet matching the Atlas AC console.

With the addition of the 206 to your Atlas station, you have the ultimate in a high performance, maximum versatility, compact, amateur radio station. $299.

For complete details see your Atlas dealer, or drop us a card and we'll mail you a brochure with dealer list.

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More Details? CHECK—OFF Page 126

flea market

KEYER KITS, automatic with weight control. Use ordinary paddle or homebrew "touch" paddle. Draws less than 2 ma at 9 volts. Ideal for portable operation. Includes all P.C.B. components and full instructions. Send S.A.S.E. for details or send $17 to NS. Circuits, Jacob Sollenberger, Route 7, Chambersburg, Pa. 17201.

PC's, Send large S.A.S.E. for list. Semtronics, Rt. #3, Box 1, Bellaire, Ohio 43906.

MAG TAPE, 2,400′ of ½ inch tape for UTRS $7.00/reel. Interested in video, ATV or SSTV? Get on mailing list for free copy of new magazine. W6LVY, Dennis Trimble, 5835 Herma St., San Jose, Ca. 95123.

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada. MIS 3B4.

FREE flyer of unusual surplus equipment and components: military, RF, semiconductors, coils, capacitors, etc. Free gift with first order. Gold Electronics Company, Dept. H, Box 1814, Rochester, NY 14603.

ENGRAVED RADIO LICENSE. Exact reproduction in solid brass. Permanent identification. Send good Xerox copy, with $5.00, to Metal Art Graphics, 1136 Potomac Ave., Hagerstown, Md. 21740.

TELETYPE EQUIPMENT FOR SALE for beginners and experienced operators. RTTY machines, parts, supplies. Special beginners package consists of Model 15 page printer and TH5-TG demodulator, $125.00. Atlantic Surplus Sales, 3730 Nautilus Ave., Brooklyn, N. Y. 11224. Tel: (212) 372-0349.

TELETYPEWRITER PARTS, gears, manuals, supplies, tape, toroids. SASE list. Typetronics, Box 8873, Ft. Lauderdale, Fl. 33310. Buy parts, late machines.

QSL'S — BROWNIE W3CJI — 3035B Lehigh, Allentown, Pa. 18103. Samples with cut catalog 50¢.

NEED HELP holding your PC board while you solder the components? My fixture will sit on your kitchen table and adjusts to fit PC boards up to 6¾″ wide (length unlimited). Order now. Only $7.95 ppd in USA (Mo. residents add 25¢). W. N. Wellman Co., Box 722A, 451 Saline Rd., Fenton, Mo. 63026.

EXCLUSIVELY HAM TELETYPE 21st year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 35¢. $3.50 per year. Box 837, Royal Oak, Michigan 48068.

OSCAR 7, SSB-CW TRANSMIT CONVERTERS. For 28 or 50 MHz input at 20 mw. 432 MHz output at 1 watt. Solid state, for 12 volt supply. 35 watt solid state amplifier available for this converter. Units designed and built by WØENC. Write for information. UHF-VHF Communications, 53 St. Andrew, Rapid City, S. D. 57701.

YOUR AD belongs here too. Why not send it in today.

Coming Events

WALTHAM, MASS. 1st Annual WR1ABV Repeater Flea Market sponsored by advisory committee. August 15, 1976 10 AM to 6 PM. Municipal parking lot, corner of Moody & Pine Streets, Waltham. Refreshments available. Talk-in on 146/64 and 449.05. Sellers $4, Buyers $1, to benefit the WR1ABV Repeater. Contact W1KSZ for further information.

MEMPHIS IS BEAUTIFUL IN OCTOBER! The Memphis Hamfest, bigger and better than the 3,500 who attended last year, will be held at State Technical Institute, Interstate 40 at Macon Road, on Saturday and Sunday, October 2 and 3. Demonstrations, displays, MARS meetings, flea market, ladies' flea market, too! Hospitality room, informal dinners, XYL entertainment, many outstanding prizes. Dealers and distributors welcome, too! Contact Harry Simpson, W4SCF, Box 27015, Memphis, TN 38127, phone 901 358-5707.

HAMBURG INTERNATIONAL HAMFEST, Sept. 18, 1976 at the Erie County Fairgrounds, Hamburg, N. Y., near Buffalo and Niagara Falls. Registration $3/$2.50 (advance). Flee market space $1.00/$5.00. Fri. evening hospitality get-together with pre-registration drawing of a valuable prize, technical and organizational meetings, equipment displays, women's program, R/V and picnic facilities, fun, prizes. Talk in freqs: 146.52, 7.255 (ECARS), and 3.925 MHz. Info: Bert Jones, W2CUU, 143 Orchard Drive, Kenmore, N. Y. 14223, Phone: (716) 873-3984.

More Details? CHECK—OFF Page 126

7400N TTL

Part	Price	Part	Price	Part	Price
SN7400N*	16	SN7453N	27	SN74150N	1.00
SN7401N	16	SN7454N	20	SN74151N	79
SN7402N	21	SN7459A	25	SN74153N	89
SN7403N	16	SN7460N	22	SN74154N*	1.25
SN7404N*	18	SN7470N	45	SN74155N	99
SN7405N	24	SN7471N*	69	SN74156N	99
SN7406N	20	SN7472N	39	SN74157N	99
SN7407N	29	SN7473N*	37	SN74160N	1.25
SN7408N	25	SN7474N*	32	SN74161N	99
SN7409N	25	SN7475N*	50	SN74163N*	99
SN7410N*	18	SN7476N*	32	SN74164N	1.10
SN7411N	30	SN7479N*	5.00	SN74165N	1.10
SN7412N	33	SN7480N	50	SN74166N	1.25
SN7413N	45	SN7482N	98	SN74167N	5.50
SN7414N	70	SN7483N	70	SN74170N	2.10
SN7416N	35	SN7485N	89	SN74172N	18.00
SN7417N	15	SN7486N	39	SN74173N	1.50
SN7420N	21	SN7488N	3.50	SN74174N	1.25
SN7421N	33	SN7489N	2.25	SN74175N	99
SN7422N*	49	SN7490N*	45	SN74176N	90
SN7423N	37	SN7491N	75	SN74177N	90
SN7425N	29	SN7492N	49	SN74180N	99
SN7426N	29	SN7493N*	49	SN74181N	2.49
SN7427N	37	SN7494N	79	SN74182N	95
SN7429N	42	SN7495N	79	SN74184N	1.95
SN7430N	26	SN7496N	89	SN74185N	2.20
SN7432N	31	SN7497N*	4.00	SN74186N*	5.00
SN7437N	27	SN74100N*	1.00	SN74187N	6.00
SN7438N	27	SN74107N	39	SN74190N	1.19
SN7439N	25	SN74121N*	39	SN74191N	1.25
SN7440N	15	SN74122N	39	SN74192N*	89
SN7441N	89	SN74123N*	70	SN74193N*	89
SN7442N	59	SN74125N	60	SN74194N	1.25
SN7443N	75	SN74126N	60	SN74195N	75
SN7444N	75	SN74132N	1.09	SN74196N	1.25
SN7445N	75	SN74136N	95	SN74197N	75
SN7446N	81	SN74141N	1.15	SN74198N	1.75
SN7447N*	69	SN74142N*	4.00	SN74199N	1.75
SN7448N	79	SN74143N*	4.50	SN74200N	5.59
SN7450N	26	SN74144N*	4.50	SN74279N*	90
SN74C. N	27	SN74145N	1.15	SN74251N	1.79
		SN74147N	2.35	SN74284N	6.00
		SN74148N	2.00	SN74285N	6.00

MANY OTHERS AVAILABLE ON REQUEST
20% Discount for 100 Combined 7400's

CMOS

Part	Price	Part	Price	Part	Price
CD4000	25	CD4035	1.85	74C04N	75
CD4001	25	CD4040	2.45	74C10N	65
CD4002	25	CD4042	1.90	74C20N	65
CD4006	2.50	CD4044	1.50	74C30N	65
CD4007	25	CD4046	2.51	74C42N	2.15
CD4009	59	CD4047	2.75	74C73N	1.50
CD4010	59	CD4049	79	74C74	1.15
CD4011	25	CD4050	79	74C90N	3.00
CD4012	25	CD4051	2.95	74C95N	2.00
CD4013	47	CD4053	2.95	74C107N	1.25
CD4016	56	CD4060	3.25	74C151	2.90
CD4017	1.35	CD4066	1.75	74C154	3.00
CD4019	55	CD4069	45	74C157	2.15
CD4020	1.49	CD4071	45	74C160	3.25
CD4022	1.25	CD4072*	45	74C161	3.25
CD4023	25	CD4081	45	74C163	3.00
CD4024	1.50	CD4511	2	74C164	3.25
CD4025	25	CD4518	2	74C173	2.60
CD4027	69	74C00N	39	74C193	2.75
CD4028	1.65	74C02N	55	74C195	2.75
CD4029	2.90			MC4044*	4.50
CD4030	65			MC14016*	56

LINEAR

Part	Price	Part	Price	Part	Price
LM300H	80	LM377N	4.00	LM1351N	1.65
LM301H	35	LM380N	1.39	LM1414N	1.75
LM301CN	35	LM380CN	1.05	LM1458C	65
LM302H	75	LM381N	1.79	LM1496N	95
LM304H	1.00	LM382N	1.79	LM1556V	1.85
LM305H	95	NE501K	8.00	LM2111N	1.95
LM307CN	35	NE510A	6.00	LM2901N	2.95
LM308H	1.00	NE531H	3.00	LM3065N	69
LM308CN	1.00	NE536T	6.00	LM3900N	55
LM309H	1.10	NE540L	6.00	LM3905N	60
LM309K	99	NE550N	79	LM3909*	1.25
LM310CN	1.15	NE553	2.50	LM5556N	1.85
LM311H	90	NE555V*	45	MC5558V	1.00
LM311N	90	NE560B*	5.00	LM7525N	90
LM318CN	1.50	NE561B*	5.00	LM7535N	1.25
LM319N	1.30	NE562B*	5.00	8038B*	4.95
LM319D	9.00	NE565H*	99	LM75450	49
LM320K-5	1.35	NE565N*	1.25	75451CN	39
LM320K-5.2	1.35	NE566CN*	1.95	75452CN	39
LM320K-12	1.35	NE567H*	1.25	75453CN	39
LM320K-15	1.35	NE567V*	1.50	75454CN	39
LM320T-5*	1.75	LM703CN	45	75491CN	79
LM320T-8*	1.75	LM709H	29	75492CN	89
LM320T-12*	1.7	LM709N	29	75494CN	89
LM320T-18*	1.75	LM710N	79	**RCA LINEAR**	
LM323K-5*	9.95	LM711N	39	CA3013	1.70
LM324N	1.80	LM723N	55	CA3023	2.15
LM339N	1.70	LM723H	55	CA3035	2.25
LM340K-5	1.95	LM733N	1.00	CA3039	1.35
LM340K-12	1.95	LM739N	1.29	CA3046	1.15
LM340K-15	1.95	LM741CH	35	CA3059	2.46
LM340K-24	1.95	LM741CN	35	CA3060	2.80
LM340T-5	1.75	LM741 14N	39	CA3080	85
LM340T-6	1.75	LM747H	79	CA308*	1.75
LM340T-8*	1.75	LM747N	79	CA3082	2.00
LM340T-18*	1.75	LM748H	39	CA3083	1.60
LM340T-12	1.75	LM748N	39	CA3086	59
LM340T-15	1.75	LM1303N	90	CA3089	3.25
LM340T-24	1.75	LM1304N	1.19	CA3091	8.25
LM350N	1.00	LM1305N	1.40	CA3123	1.85
LM351CN	65	LM1307N	85	CA3130*	1.49
LM370N	1.15	LM1310N	2.95	CA3600	1.75
LM370H	1.15			RC4194	5.95
LM373N	3.25			RC4195	3.25

KITS EXAR ICS

XR-2206KA **SPECIAL $17.95**
Includes monolithic function generator IC, PC board, and assembly instruction manual

XR-2206KB **SPECIAL $27.95**
Same as XR 2206KA above and includes external components for PC board

TIMERS		**STEREO DECODERS**	
XR-555CP	$.69	XR-1310P	$3.20
XR-320P	1.55	XR-1310EP	3.20
XR-556CP	1.85	XR1800P	3.20
XR-2556CP	3.20		
XR-2240CP	3.25	**WAVEFORM GENERATORS**	
		XR-205	8.40
PHASE LOCKED LOOPS		XR-2206CP	4.49
XR-210	5.20	XR-2207CP	3.85
XR-215	6.60	**MISCELLANEOUS**	
XR-567CP	1.95	XR-2211CP	6.70
XR-2567CP	2.99	XR-2261	3.79
		XR4136	2.00

DATA HANDBOOKS

7400	Pin-out & Description of 5400/7400 ICS	**$2.95**
CMOS	Pin-out & Description of 4000 Series ICS	**$2.95**
LINEAR	Pin-out & Functional Description	**$2.95**

ALL THREE HANDBOOKS $6.95

CONSUMER ELECTRONICS

DIGITAL WATCH
This watch is manufactured by National Semiconductor. It provides 5 functions; hours, minutes, seconds, date, A.M. indicator dot. Accuracy is assured to 5 seconds per month by precision quartz crystal. If something should go wrong with the watch, repair is assured within 48 hours after it is received. Complete with steel black leather band.

ES4-YS
3 MICRON GOLD PLATE BEZEL
$29.95
NOT A KIT

Novus

DIGITAL ALARM CLOCK
This 4 digit Novus Alarm Clock is a very reliable and smartly styled unit. It provides such features as an alarm settable to any minute of the day, a 7 minutes snooze alarm, a power failure indicator, and even an A.M., P.M. indicator.

NOT A KIT **$19.95**

OPTO ELECTRONICS

XCITON
LITRONIX
MONSANTO

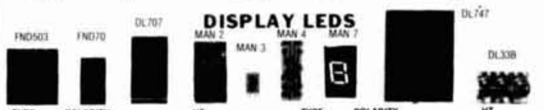
R - RED
G - GREEN
Y - YELLOW
O - ORANGE

DISCRETE LEDS

125" dia.		.185" dia.		.190" dia.	
XC209R	5/$1	XC526R	5/$1	XC111R	5/$1
XC209G	4/$1	XC526G	4/$1	XC111G	4/$1
XC209Y	4/$1	XC526Y	4/$1	XC111Y	4/$1
XC209O	4/$1	XC526O	4/$1	XC111O	4/$1

.200" dia.		.200" dia.		.085" dia.	
XC22R	5/$1	XC556R	5/$1	MV50	
XC22G	4/$1	XC556G	4/$1	.085" dia. Micro red led	
XC22Y	4/$1	XC556Y	4/$1		6/$1
XC22O	4/$1	XC556O	4/$1		

DISPLAY LEDS



TYPE	POLARITY	HT		TYPE	POLARITY	HT	
MAN 1	COMMON ANODE	270	$3.25	MAN 74	COMMON CATHODE	300	$1.50
MAN 2	5 x 7 DOT MATRIX	300	4.95	DL707	COMMON ANODE	300	$1.50
MAN 3	COMMON CATHODE	125	39	DL747	COMMON ANODE	600	2.25
MAN 4	COMMON CATHODE	187	1.95	DL750	COMMON CATHODE	600	2.49
MAN 7	COMMON ANODE	300	1.50	DL338	COMMON CATHODE	110	1.25
MAN 7G	COMMON ANODE-GREEN	300	1.95	FND70	COMMON CATHODE	250	50
MAN 7Y	COMMON ANODE-YELLOW	300	1.95	FND503	COMMON CATHODE	500	1.75
MAN 72	COMMON ANODE	300	1.50	FND507	COMMON ANODE	500	1.75

IC SOLDERTAIL — LOW PROFILE (TIN) SOCKETS

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

SOLDERTAIL STANDARD (TIN)

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

SOLDERTAIL STANDARD (GOLD)

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

WIRE WRAP SOCKETS (GOLD) LEVEL #3

10 pin	$.45	41	37	24 pin	$1.05	95	85
14 pin	39	38	37	28 pin	1.40	1.25	1.10
16 pin	43	42	41	36 pin	1.59	1.45	1.30
18 pin	75	68	62	40 pin	1.75	1.55	1.40

50 PCS. RESISTOR ASSORTMENTS $1.25 PER ASST.

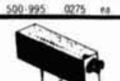
		10 OHM	12 OHM	15 OHM	18 OHM	22 OHM	
ASST. 1	5 ea.	27 OHM	33 OHM	39 OHM	47 OHM	56 OHM	1/4 WATT 5% — 50 PCS.
		68 OHM	82 OHM	100 OHM	120 OHM	150 OHM	
ASST. 2	5 ea.	180 OHM	220 OHM	270 OHM	330 OHM	390 OHM	1/4 WATT 5% — 50 PCS.
		470 OHM	560 OHM	680 OHM	820 OHM	1K	
ASST. 3	5 ea.	1.2K	1.5K	1.8K	2.2K	2.7K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

ALL OTHER RESISTORS FROM 2.2 OHMS - 5.6M AVAILABLE IN MULTIPLES OF 5 ea.
5-25 PCS. .05 ea. 30-95 PCS. .04 ea. 100-495 PCS. .03 ea. 500-995 .0275 ea.

14 PCS. POTENTIOMETER ASSORTMENTS

ASST. A 2 ea. 10 OHM-20 OHM-50 OHM-100 OHM-200 OHM-250 OHM-500 OHM
ASST. B 2 ea. 1K, 2K, 5K, 10K, 20K, 70K, 50K
ASST. C 2 ea. 50K, 100K, 200K, 250K, 500K, 1M, 2M

$9.95 Per Asst.



Each assortment contains 14 pcs of 10 turn pots. All pots are available in single unit quantities $.99 ea.

Astrisk Denotes Items On Special For This Month
Satisfaction Guaranteed. $5.00 Min. Order. U.S. Funds.
California Residents — Add 6% Sales Tax — Data Sheets 25c each
Send a 13c Stamp (postage) for a FREE 1976 Catalog

JAMES

1021 HOWARD AVE., SAN CARLOS, CA. 94070
PHONE ORDERS — (415) 592-8097

74LS00 TTL

Part	Price	Part	Price	Part	Price
74LS00	.39	74LS55	.39	74LS151	1.55
74LS02	.39	74LS73	.65	74LS153	1.89
74LS03	.39	74LS74	.65	74LS157	1.55
74LS04	.45	74LS75	.79	74LS162	2.25
74LS05	.45	74LS76	.65	74LS163	2.25
74LS08	.39	74LS83	2.19	74LS164	2.25
74LS10	.39	74LS86	.65	74LS181	3.69
74LS13	.79	74LS90	1.25	74LS190	2.85
74LS14	2.19	74LS92	1.25	74LS191	2.85
74LS20	.39	74LS93	1.25	74LS192	2.85
74LS26	.49	74LS95	2.19	74LS193	2.85
74LS27	.45	74LS96	1.89	74LS194	2.25
74LS28	.49	74LS107	.65	74LS195	2.25
74LS30	.39	74LS112	.65	74LS257	1.89
74LS32	.45	74LS132	1.55	74LS260	.55
74LS40	.49	74LS136	.65	74LS279	.79
74LS51	.39	74LS138	1.89	74LS670	5.95

THIS MONTH ONLY!!! DL728
The DL728 is a dual 0.5" common cathode red display. It is ideal for use with clock chips, as segments are already multiplexed. **$1.95**

MINATURE TOGGLE

DPDT			PART	1	2-9	10-29	30-99
ON	OFF	ON	221	3.05	2.65	1.97	1.76
ON	NONE	ON	223	2.85	2.25	1.68	1.51
ON	OFF	ON	121	2.35	1.95	1.43	1.30
ON	NONE	ON	123	2.05	1.65	1.21	1.10

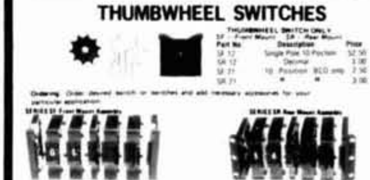
PUSH BUTTON

SPDT - SERIES PB, PB1

	Model Number	1 Each	2-9	10-29	30-99
Maintained Action Switch: Push To Make Contact And Push Again To Release Action	PB-123	$2.35	$1.95	$1.47	$1.30
Momentary Action Switch: Normally Open or Normally Closed. 1 for N.C. circuit use terminals 2 & 1, for N.O. 2 & 3, for N.C and N.O. 1, 2 & 3	PB-126	$2.35	$1.95	$1.47	$1.30

5 AMP RATINGS

THUMBWHEEL SWITCHES



THUMBWHEEL SWITCH ONLY
SF - Front Mount SR - Rear Mount

Part No	Description	Price
SF 12	Single Pole 10 Position	$2.50
SR 12	Decimal	3.00
SF 21	10 Position BCD only	2.50
SR 21	" "	3.00

Ordering: Order desired switch or switches and add necessary accessories for your particular application.

SERIES SF Front Mount Assembly — SERIES SR Rear Mount Assembly

Part No	ACCESSORIES Description	Price	Part No	ACCESSORIES Description	Price
SF EP	End Plate (Pair)	$.50	SR EP	End Plates (pair)	$.50
SF DP	Divider Plate (each)	40	SR DP	Divider Plate (each)	40
SF BB	Blank Body (each)	40	SR BB	Blank Body (each)	40
SF HB	Half Body (each)	40	SR HB	Half Body (each)	40

8 POSITION ROTARY SWITCH

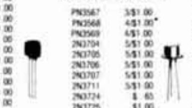
2 FOR $1.00

These switches are a 7 position, one position open, rotary switch enclosed in a TO-5 can. They have a standard 8 pin configuration and will mount perfectly on printed circuit board.

ZENERS—DIODES—RECTIFIERS

TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
IN746	3.3	400m	4/1.00	1N4005*	600 PIV	1 AMP	10/1.00
1N751A	5.1	400m	4/1.00	1N4006*	800 PIV	1 AMP	10/1.00
1N752	5.6	400m	4/1.00	1N4007*	1000 PIV	1 AMP	10/1.00
1N753	6.2	400m	4/1.00	1N3600	50	200m	6/1.00
1N754	6.8	400m	4/1.00	1N4148	75	10m	15/1.00
1N965B	15	400m	4/1.00	1N4154	35	10m	12/1.00
1N5232	5.6	500m	28	1N4734	5.6	1w	28
1N5234	6.2	500m	28	1N4735	6.2	1w	28
1N5235	6.8	500m	28	1N4736	6.8	1w	28
1N5236	7.5	500m	28	1N4738	8.2	1w	28
1N456	25	40m	6/1.00	1N4742	12	1w	28
1N458	150	7m	6/1.00	1N4744	15	1w	28
IN485A	180	10m	5/1.00	1N1183	50 PIV	35 AMP	1.60
1N4001	50 PIV	1 AMP	12/1.00	1N1184	100 PIV	35 AMP	1.70
1N4002	100 PIV	1 AMP	12/1.00	IN1186	200 PIV	35 AMP	1.80
1N4003	200 PIV	1 AMP	12/1.00	1N1188	400 PIV	35 AMP	3.00
1N4004	400 PIV	1 AMP	12/1.00				

TRANSISTORS



MPS A05	5/$1.00			PN4249	4/$1.00
MPS A06	5/$1.00	PN3567	3/$1.00	PN4250	4/$1.00
2N2219A	3/$1.00	PN3568	4/$1.00*	2N4400	4/$1.00
2N2221	4/$1.00	PN3569	4/$1.00	2N4401	4/$1.00
2N2222A	5/$1.00	2N3704	5/$1.00	2N4402	4/$1.00
2N2369	5/$1.00	2N3705	5/$1.00	2N4403	4/$1.00
2N2369A	4/$1.00	2N3706	5/$1.00	2N4409	5/$1.00
FN2415	5/$1.00	2N3707	5/$1.00	2N5086	4/$1.00
2N2484	4/$1.00	2N3711	5/$1.00	2N5087	4/$1.00
2N2906A	4/$1.00	2N3724	$.65	2N5088	4/$1.00
2N2907A	5/$1.00	2N3725	$1.00	2N5089	4/$1.00
2N2925	5/$1.00	2N3903	5/$1.00	2N5129	5/$1.00
2N3053	2/$1.00	2N3904	4/$1.00	2N5138	5/$1.00
2N3055	$.89	2N3905	4/$1.00	2N5139	5/$1.00
MJE3055	$1.00	2N3906	4/$1.00	2N5209	5/$1.00
MJE2955	$1.25	2N4013	3/$1.00	2N5951	5/$1.00
2N3392	5/$1.00	2N4014	3/$1.00	C106B1SCR	2/$1.00
2N3398	5/$1.00			2N6432	$2.00

CAPACITOR CORNER

	1-9	10-49	50-100		1-9	10-49	50-100
50 VOLT CERAMIC DISC CAPACITORS							
	05	04	03	001μF	05	04	035
	05	04	03	0047μF	05	04	035
	05	04	03	01μF	05	04	035
100 pf	05	04	03	022μF	06	05	04
220 pf	05	04	03	047μF	06	05	04
470 pf	05	04	035	1μF	12	09	075
100 VOLT MYLAR FILM CAPACITORS							
001mf	12	10	07	022mf	13	11	08
0022	12	10	07	047mf	21	17	13
0047mf	12	10	07	1mf	27	23	17
.01mf	12	10	07	22mf	33	27	22
+ 20% DIPPED TANTALUMS (SOLID) CAPACITORS							
1 35V	28	.23	.17	1.5 35V	30	26	21
15 35V	28	.23	.17	2.2 25V	31	27	22
22 35V	28	.23	.17	3.3 25V	31	27	22
33 35V	28	.23	.17	4.7 25V	32	28	23
47 35V	28	.23	.17	6.8 25V	36	31	25
68 35V	28	.23	.17	10 25V	40	35	29
1.0 35V	28	.23	.17	15 25V	63	50	40
MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS							
Axial Lead				**Radial Lead**			
47 50V	15	13	10	47 25V	15	13	10
1.0 50V	16	14	11	47 50V	16	14	11
3.3 50V	15	13	10	1.0 16V	15	13	10
4.7 25V	16	14	12	1.0 25V	16	14	11
10 25V	15	13	10	1.0 50V	16	14	11
10 50V	16	14	12	4.7 16V	15	13	10
22 25V	17	15	12	4.7 25V	15	13	10
22 50V	24	20	18	4.7 50V	16	14	11
47 25V	19	17	15	10 16V	14	12	09
47 50V	25	21	19	10 25V	15	13	10
100 25V	24	20	18	10 50V	16	14	12
100 50V	35	30	28	47 50V	24	21	19
220 25V	32	28	25	100 16V	19	15	14
220 50V	45	41	38	100 25V	24	20	18
470 25V	33	29	27	100 50V	35	30	28
1000 16V	55	50	45	220 16V	23	17	16
2200 16V	70	62	55	470 25V	31	28	26

flea market

HAMFAIR-76, TACOMA, Washington. August 21 & 22. The Radio Club of Tacoma (W7DK) presents HAMFAIR-76 at Pierce County Fairgrounds, 11 miles south of Puyallup, Washington. Contests, flea market, seminars, women & children's activities, contests, Saturday evening dinner and Sunday morning loggers breakfast. Free camping with electrical hookups. First prize ICOM IC-230. Contact W7GPR, 3421 E. 138th St., Tacoma, WA 98446. Phone: 531-3821.

RADIO EXPO '76, September 18, 19, near Chicago. Exhibits, seminars, giant flea market open Friday night. Campers welcome. Advance ticket $1.50. Box 1014, Arlington Heights, Ill. 60006.

VIRGINIA. Shenandoah Valley Amateur Radio Club Hamfest, August 1, 1976. Clarke County Ruritan Fairgrounds, Berryville, VA (8 mi. east of Winchester on Route #7). New and better factilities. Exhibitors no charge. Further info from Neil Woods, W4LOG.

MELBOURNE, FL., SEPT. 11-12. The 11th Annual Melbourne, Florida Hamfest will be held Saturday and Sunday, September 11-12, 1976, from 9 a.m. to 5 p.m. each day in the air conditioned Melbourne Civic Auditorium located on Hibiscus Boulevard. Donation is $2.50 per adult. Full program includes forums, meetings, auction, swap tables, commercial exhibits, awards, prizes, etc. Talk in on 25/85 and 52/52. Sponsored by Platinum Coast Amateur Radio Society. For more info write P. O. Box 1004, Melbourne, FL 32901.

FLORIDA. Melbourne Hamfest, September 11 & 12, Melbourne Auditorium. Contact Mike Waters, VE3BYO/W4, 965 Golden Beach Blvd., Indian Harbour Beach, FL 32937.

MEMPHIS, TN. Hamfest sponsored by Mid-South Amateur Radio Association, Delta Amateur Radio Club and Mid-South VHF Club. October 2 & 3 at State Technical Institute, 5983 Macon Cove, Memphis. Large flea market, no charge for exhibitors. Contact D. H. Powell, 4662 Crossover Lane, Memphis, TN 38117 for details.

MISSOURI. SCARC Hamfest, August 22 at Diermann Lake (located about 15 miles southwest of St. Charles, Missouri on State Highway K). Admission $1. Large flea market (no charge for set up). For more info contact St. Charles Amateur Radio Club, 3032 Mockingbird Dr., St. Charles, MO 63301.

WELCOME — Bicentennial Edition Golden Spread Hamfest. Quality Inn, Amarillo, Tex. Aug. 14, 15, 1976. Full info - write to 4408 Mesa Circle, Amarillo, Tex. 79109.

HAMFEST — Springfield, Illinois! First. Sangamon Valley Radio Club invites everyone. Sunday, September 26th. Rain-shine. Sangamon County Fairgrounds, New Berlin, Illinois on U.S. 36. Write K9HDZ, 622 Magnolia, Rochester, Illinois 62561.

ELMIRA, N. Y. HAMFEST: Sept. 25, 1976, Chemung County Fairgrounds. Flea market, dealer displays, technical talks. Talk in 10/70-146.52. $2.00 advance sale - $2.50 at gate. For further information, WA2SMM, 320 W. Ave., Elmira, N. Y. 14904.

CINCINNATI HAMFEST: 40th Anniversary Hamfest - Sunday, September 19, 1976 at the New Improved Stricker's Grove on State Route 128, one mile west of Ross (Venice) Ohio. Flea market, exhibits, contests, model aircraft flying, food and beverages all day. Advance ticket sales $7.00 - Tickets at the gate $8.00 - covers everything. For further information: Lillian Abbott, 1424 Main Street, Cincinnati, Ohio 45210.

K2DEL. Knight Raiders VHF Club's auction and flea market will be held on Saturday, August 14th, at St. Joseph's Church of East Rutherford, Hoboken Road, East Rutherford. Free admission, free parking, refreshments available. Talk-in will be on 146.52. Doors will open 10 AM. Flea market tables: $6.00 for a full table, $3.50 for a half table. Reserve your tables in advance by writing to The Knight Raiders VHF Club, K2DEL, P. O. Box 1054, Passaic, New Jersey 07055.

CAST YOUR BALLOT Nov. 6th and 7th for Southern Hospitality at the Florida Section Convention on the beach at the Sheraton Sand Key hotel, Clearwater Beach. Technical sessions on the latest advances in Amateur Radio and orientation for the newcomer. Family attractions nearby, including Disney World and Busch Gardens. Early snowbirds can catch season activity at its best. Bonus gift for early registration of $3.00. Saturday night banquet with ARRL President Dannals and others at $9.00 per. Write Florida Gulf Coast Amateur Radio Council - FGARC, Inc., P. O. Box 157, Clearwater, Florida 33517.

More Details? CHECK—OFF Page 126

More Details? CHECK—OFF Page 126

flea market

OHIO: 34th Annual Findlay Hamfest - Sept. 12, Riverside Park, Findlay, Ohio. Talk-in 146.52. For advanced tickets and/or info write: (SASE please for under 5 tickets) Clark Foltz, W8UN, 122 W. Hobart St., Findlay, Ohio 45840.

LA PORTE, INDIANA — The combined La Porte County Amateur Radio Clubs will hold their Fall Hamfest on Sunday, August 29th, 1976 at the La Porte County Fairgrounds in La Porte, beginning at 7 a.m. Chicago time. Overnight camping available. Indoors in case of rain. No table or set-up charge. Paved midway, good food and drink. $2.00 donation at the gate. For info write, P. O. Box 30, La Porte, IN 46350. Talk-in on .01-.61 & 94 Simplex.

MICHIGAN: The Fourth Annual L'Anse Creuse ARC Swap & Shop will be held on September 19, 1976 at the L'Anse Creuse High School in Mount Clemens, Michigan. Doors will be open from 0900 to 1500 EDST. First prize $200.00 cash. Talk-in on 146.52 and 146.94. Admission $1.50 at door, $1.00 in advance. For tickets enclose $1.00 and S.A.S.E. and send to, Robert Harder, WB8ILI, 51769 Base, New Baltimore, Mich. 48047.

MICHIGAN: The Jewish Community Center Amateur Radio Club of Metropolitan Detroit's fourth Annual Swap 'n Shop is Sunday, August 22, from 9:00 a.m. to 3:00 p.m. at the new Center at Maple and Drake Rds., W. Bloomfield. Talk in on 146.94. For more information, contact Bob, W8DGR at 6600 W. Maple Rd., W. Bloomfield, Mi. 48033.

SOUTH JERSEY RADIO ASSN., 28th Annual Hamfest: Sept. 12, 1976, 10-5 p.m. at Molia Farms, Malaga, N. J. Lake, picnic grounds and food available. Tailgate sales, swap shop and door prizes. Family tickets: advance sales - $2.50, gate sales - $3.50. Advance sales send S.A.S.E. to Jack Koch, Box 103, Cherry Hill, N. J. 08002. Talk in 146.52.

AURORA, ILLINOIS, August 22, 1976. The Fox River Radio League - W9CEQ Hamfest will be held August 22, 1976 at beautiful Phillips Park, east edge of Aurora, U.S. Hwy. Rt. #30. Talk in on 146.94. All day family fun, picnic, zoo, lake and flowers. Same old price. $1.00 advanced with S.A.S.E. to F.R.R.L., P. O. Box 443, Aurora, Ill. 60507.

1976 DELTA QSO PARTY. All amateurs are invited to participate. Contacts must take place from 2000Z Sept. 25 to 0200Z Sept. 27. No time or power restrictions. Amateurs outside Delta Division will attempt to contact as many amateurs inside Delta Division (Ark-La-Miss-Tenn) as possible. Delta Division Amateurs will attempt to contact as many amateurs as possible both inside and outside of Delta Division. For rules and complete details, send SASE to Malcolm P. Keown, W5RUB, 213 Moonmist, Vicksburg, MS 39180.

BLOSSOMLAND SWAP-SHOP, October 3rd, Berrien County Fair Grounds, Berrien Springs, Michigan. Greatly expanded facilities, 150 tables, entertainment, refreshments. Advance ticket donation $1.50, tables $2.00. Write: John Sullivan, P. O. Box 345, St. Joseph, Michigan 49085. Make checks payable to: Blossomland Hamfest.

Stolen Equipment

ICOM IC-22A transceiver, s/n1216, 16/76, 31/91, 01/61 xtals plus standard Icom xtals. Motorola type microphone. Realistic PRO-11 scanner, s/n 08370930. Xtals for 155.79 & 155.685 MHz in the unit. Sanyo FT-867 AM/FM 8 trk in dash car radio, sn/87661611. Realistic MPA-10 PA amplifier, non-stock mike w/PA amp. Name and SSN (214-68-9618) engraved on each unit. Any info contact Steve Martin, WA3SAD (phone 301-627-4933).

UNITMETRICS ULTRA-COM 25, SN 080213, stolen from locked auto parked in residence driveway about 5 AM, May 12. Unit engraved N. C. driver license #2067134. Contact Greensboro, N. C. Police Department or W4DWR.

ICOM IC230 two meter FM transceiver with mount and 32 xtal, ser. no. 2835. TPL model 1002 two meter power amplifier ser. no. 0426. Regency 10 channel scanner, model ACTR 10HLU with all crystals and antenna junction box, ser. no. 185A88279. If located, advise San Diego Police Dept., Burglary Div. at 236-6281, Case #76-33350 or Zane Sprague, K6WK (714) 481-0594.

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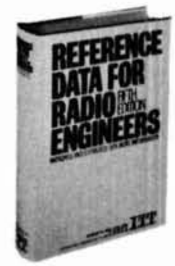

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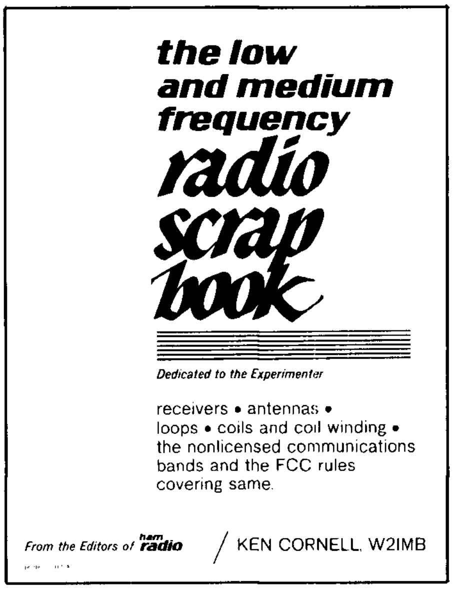

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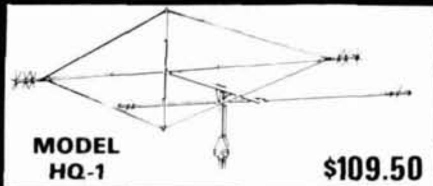

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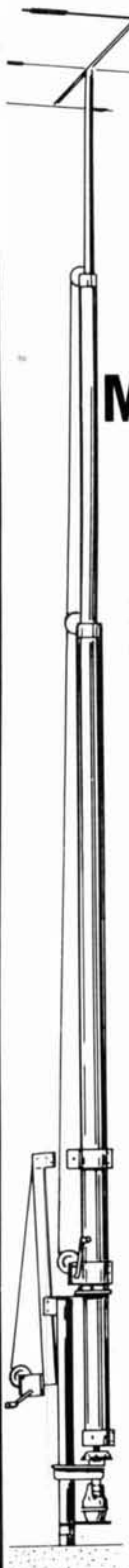

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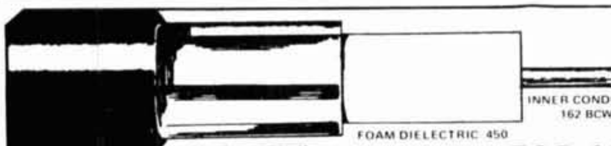

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August 1976

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