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magazine

JUNE 1977

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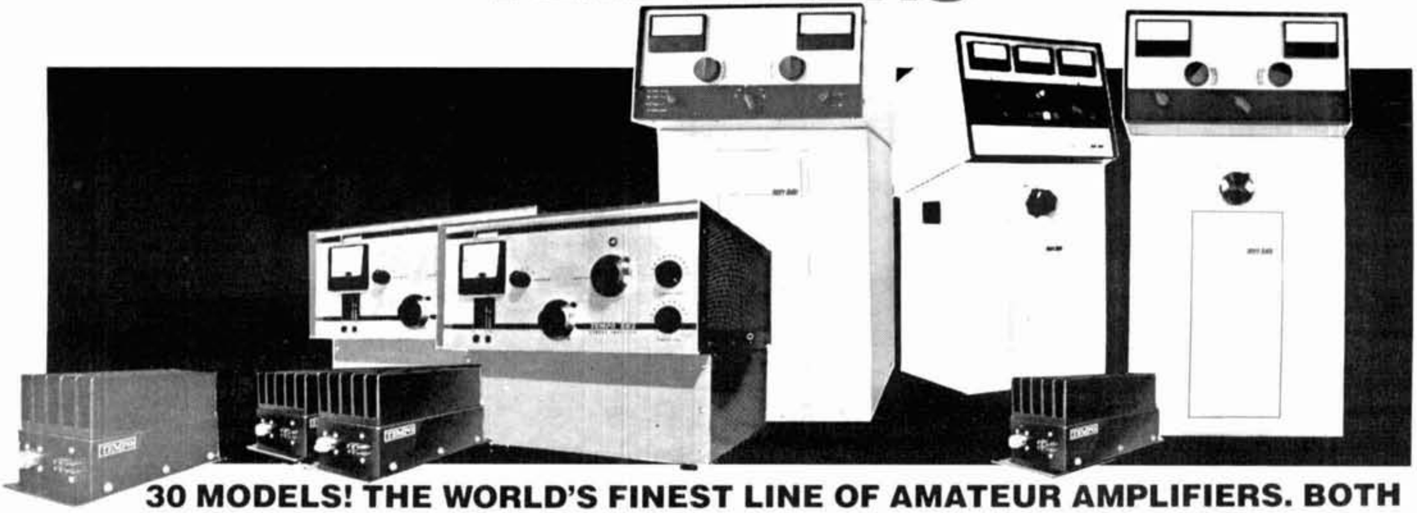

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a second look

by Jim Fisk

In early March the FCC released new rules which could have a substantial effect on the whole future of amateur radio. The new rules, spelled out in Docket 20777, became effective on April 15th, and require that the spurious emissions from *all high-frequency transmitters, transceivers, and amplifiers must be at least 40 dB below the mean power* of the fundamental, without exceeding 50 milliwatts. The rules for vhf transmitters operating between 30 MHz and 235 MHz are even more stringent: for transmitters of 25 watts or more, spurious emissions must be down a minimum of 60 dB. Note that this is not just a proposal — it is the law, and it applies to *all* amateur equipment: new, used, or presently in use!

Furthermore, as the law is now written, there is no provision for the use of external filters to reduce spurious emissions to the required level; the rules imply that the required purity of emissions must be measured at the transmitter or amplifier output connector. Recent tests by the ARRL Technical Department with a Hewlett-Packard spectrum analyzer indicate that *most* of the commercial high-frequency transmitters now on the market meet the new requirements.

In most cases, when the Commission adopts new rules, it makes an effort to minimize hardships and harmful economic effects on the licensees — usually by providing enough lead time so equipment can be brought into line with the new requirements. That was not done in this case. Most amateurs, in fact, probably weren't even aware of the new restriction until several weeks after it went into effect, and few of those who did know about it had any way of measuring the spectral purity of their transmitters!

All amateur equipment sold after April 15th must comply with the new restrictions on purity of emissions; but what do you do if you are using equipment which was manufactured before April 15, 1977? The law doesn't tell you *how* to comply, only that you must. The best way is with a spectrum analyzer, but a new one costs as much as a small house, and good used ones cost as much as a complete transceiver. You might consider the homebuilt analyzer described in this issue (and we'll have others in the future) or perhaps the members of your club can be persuaded to pool their assets to buy one. There are also commercial test labs which can make the measurements for you, but no matter which approach you take, the price isn't going to be cheap.

The ARRL has petitioned the FCC to stay the effective date of the new rule for nine months, and they have also petitioned for a reconsideration of the action, but at this point it's impossible to tell what the outcome wvllnbe. In the meantime, all amateur transmitters must comply with the new restrictions.

The new rules on spectral purity are just one of a series of restrictions on amateur equipment which have been brought about by unscrupulous CB dealers and manufacturers who are peddling amateur transceivers and illegal broadband "amateur" linears to CBers. The FCC district offices are besieged with complaints about RFI from these devices; in some areas television interference from CB is extremely severe, and there's a strong feeling in Washington that the CB situation is now so far out of hand that drastic action is necessary.

One way the FCC feels they can control this is to place an outright ban on the manufacture of linear amplifiers which operate between 24 and 35 MHz, and to require type acceptance of all commercial amateur transmitters. Neither of these actions would probably have much effect on the illegal CB operators (any more than gun control removes guns from the hands of criminals), but it will surely increase the prices you'll have to pay for your amateur equipment in the future. It will also mean that the manufacturers won't be able to offer you new circuits and technology as fast as they have in the past — each improvement they make in their equipment will require a whole new round of type acceptance. In-line production improvements will cease, and the natural evolution of modern amateur gear will grind to a halt.

In a statement that accompanied Docket 20117, the amplifier ban, FCC Chairman Richard Wiley expressed the hope that "the comments we receive will suggest other and better alternatives to the Commission's proposals." One alternative, proposed by the San Antonio Repeater Organization (SARO) would place legal responsibilities on the seller and buyer of amateur transmitting equipment. The R. L. Drake company has strongly endorsed the SARO proposal, but has worked out a modification which would remove the paperwork burden from the FCC and still provide the Commission with traceability and accountability for enforcement.

The Drake plan, which is based on presentation of a valid amateur license when purchasing transmitting equipment, is especially recommended. Under this plan your callsign would be recorded on the sales invoice along with the equipment serial number. If and when you sold the equipment you would keep a record of the amateur you sold it to. This plan would not add to the cost of amateur equipment because it's based largely on records which are already being kept as part of good business practice.

If the FCC located amateur equipment in the hands of an unlicensed operator, it would be a simple matter to trace the equipment, by serial number, from the manufacturer to the point where it crossed over from amateur to CB. The violator would be subject to fines up to $500 per day during the time which the offense occurred.

This proposal requires your support — the whole future of amateur radio demands it. Let's place the burden of the CB problem where it belongs: on the shoulders of the unscrupulous dealers and illegal operators.

Jim Fisk, W1HR
editor-in-chief

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SPECIFICATIONS

FREQUENCY COVERAGE	*144.00 to 148.00MHz
MODES	FM (F3) *SSB (A3J), CW (A1)
SUPPLY VOLTAGE	DC 13.8V ±15%
SIZE (mm)	90H × 155W × 235D
WEIGHT	2.7
TRANSMITTER TX OUTPUT	F3 10W *A3J 10W (PEP), A1 10W
CARRIER SUPPRESSION	40 dB OR BETTER
SPURIOUS RADIATION	-60 dB BELOW CARRIER
MAXIMUM FREQUENCY DEVIATION	±5KHz
MICROPHONE IMPEDANCE	600 OHMS
RECEIVER SENSITIVITY	*A3J, A1 0.5 MICROVOLT INPUT GIVES 10 dB S+N/N F3 0.6 MICROVOLT OR LESS FOR 20 dB QUIETING S+N+D/N AT 1 MICROVOLT INPUT 30 dB
SQUELCH THRESHOLD	-8 dB OR LESS (F3)
SPURIOUS RESPONSE	60 dB OR BETTER
SYNTHESIZER FREQUENCY RANGE	144 MHz to 148 MHz
STEP SIZE	5 KHz for FM *100 Hz or 5 KHz for SSB
STABILITY	PER C IN THE RANGE OF -10 TO +60C, ±0.000145%

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More Details? CHECK—OFF Page 126

presstop

A PROPOSAL TO LIMIT TRANSMITTER SALES to Amateur licensees was filed with the FCC by the San Antonio Repeater Organization in response to an FCC request for workable alternative solutions to the problem of non-Amateur use of Amateur linear amplifiers in the CB bands (May HR). The SARO proposal, assigned RM-2839, drew a tremendous — possibly record-breaking — number of Comments filed in its support.

R.L. Drake And ARMA were both among those filing strong supporting Comments on RM-2839. Drake's offering even described a procedure for implementing a "Proof of License" program which would require little investment of time or money by the FCC.

FCC/AMATEUR DIALOGUE was severely curtailed as a result of a court decision handed down recently in Washington. In its ruling in the case of Home Box Office vs the FCC, the U.S. Court of Appeals of the District of Columbia stated:

"Once a Notice of Proposed Rule Making has been issued...any agency official or employee who is or may reasonably be expected to be involved in the decisional process of the rule-making proceeding, should refuse to discuss matters relative to the disposal of the Rule-Making proceeding with any interested private party or any attorney or agent for any such party prior to the agency's decision."

This Prohibition Is Being interpreted to extend to 60 days after the final Report and Order on an NPRM becomes effective, to include the period in which a Petition for Reconsideration may be filed. An appeal of the decision by the FCC is considered likely but could in itself take years.

CONFUSION BETWEEN "TYPE ACCEPTANCE" and "Type Approval" seems to be contributing lots of heat but little light on much of the current discussion of Docket 21117, the FCC's proposal to require commercially-made Amateur transmitters to be Type Accepted. Type Acceptance requires only that the manufacturer submit certain performance data on his product and certify that what he markets will meet or exceed the performance he claims for it. Type Approval is a much different process, with the Commission itself performing elaborate tests following extensive testing by the maker.

As One High-Ranking FCC Spokesman said, "The FCC Type Acceptance procedure doesn't ask for any data that a manufacturer shouldn't have developed for himself, long before he was ready to put his product into production." This isn't to say that the FCC could not, or would not, require additional testing of suspect equipment in its own labs — that's what happened to CB, after FCC found that practically no CB sets met claimed specifications.

JACK ANDERSON SLAMMED AMATEUR RADIO, and the many dedicated FCC people who are also Amateurs, in his nationally-syndicated column that appeared in many papers Monday, April 4th. Thrust of the piece, which cited confusing comparisons from a confidential report prepared for U.S. Representative Elliott Levitas (D-Ga.) was that letting Commission Amateurs make CB policy was letting "the wolf guard the flock" and implied ARRL membership might be a conflict of interest for FCC personnel.

The Column Turned Out to have provided one of the best PR opportunities Amateur Radio has had in recent years. Copies of many letters to the editor, a good number of them already published, have been received by HR — and Pete O'Dell, ARRL's Public Information Officer, says the League has a pile of pro-Amateur Radio clippings about four inches high with more coming in every day. In Michigan WB8VBP even managed to turn Anderson's blast into invitations to two radio talk shows, where she was able to push the positive side of the Amateur Service very strongly.

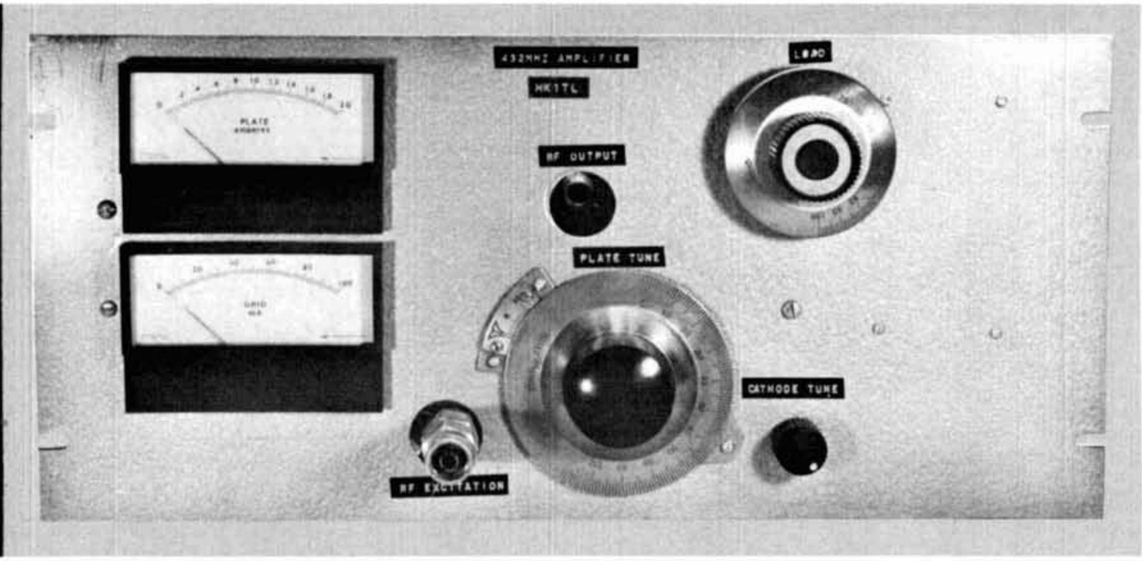
"HOW TO RESOLVE RADIO-TV INTERFERENCE PROBLEMS" is a new 35-page booklet by the FCC due out sometime in June. Coverage includes interference to telephones and a variety of electronic equipment, with useful tips for both homeowners and service technicians. Cost is about $1.50 from the Government Printing Office.

RUSSIAN COMMERCIAL AMATEUR GEAR was to have been produced for the first time under a DOSAAF (a somewhat MARS-like organization) five-year plan begun last year. Transmitters and receivers are already supposed to be out, with transceivers, keyers and other accessories to be introduced this year.

THE U.S. AMATEUR POPULATION passed 300,000 for the first time during March, with 300,372 licensed U.S. Amateurs at the month's end. The total at the beginning of the month was 296,967, and 6,693 new licensees during March more than doubled the 3,288 who let their licenses expire during the month. A year ago we numbered only 265,528, so we increased almost 35,000 in one year!

ALIEN AMATEURS WHO become citizens are no longer eligible to hold reciprocal licenses, even if their non-U.S. licenses are still valid. The new citizen must take the U.S. Amateur examination and receive a U.S. license if he wishes to remain on the air, according to a recent FCC release.

More Details? CHECK — OFF Page 126



432-MHz power amplifier

using stripline techniques

Design and construction details for building the rf amplifier used in the EME expedition to South America in 1976

This article describes a 2-kW peak-envelope power (PEP) amplifier for 432-MHz using the Eimac 8938 triode in a stripline configuration. The design evolved from work previously done by K2RIW[1] and W3CCX. This amplifier was used during the "Pack Rat's" earth-moon-earth (EME) expedition to South America in July and August of 1976. The success of the expedition was partly due to the reliability of the equipment, which included this amplifier.

Kilowatt amplifiers for 432-MHz aren't available at your favorite hobby shop. Amateurs who like to work EME either build amplifiers from scratch or from kits. A popular solution to the kW amplifier for EME is the design described by K2RIW. His stripline technique was adapted to the amplifier design described here.

background

Tests of a 432-MHz kW amplifier loaned to the South America expedition members before the trip indicated that power output suffered from poor plate-current efficiency. Worse, maximum output power was marginal for EME work, leaving no reserve in the event of low driver output or low supply voltage. This situation often occurs in a remote location. We decided to proceed with a new design for an amplifier for our South America operation.

design approach

Trouble-free operation during the planned 12-hours-per-day operating schedule, using slow CW (high duty cycle), dictated a military-type approach to amplifier design. Such an approach is to use a device rated at twice

By Tony Souza, W3HMU, Post Office Box 169, Ottsville, Pennsylvania 18942

the expected requirements to ensure reliability. An amplifier used by W3CCX uses an Eimac 8938 triode in a grounded-grid, cathode-driven configuration. This tube is rated to 500 MHz with 1500 watts plate dissipation. The tube is a coaxial-base version of the popular 8877 triode.

Advantages of the 8938 amplifier at W3CCX were trouble-free operation and easy drive requirements using only a straightforward power supply; a disadvantage of the W3CCX amplifier was that the cavity construction required metal-working facilities. Also, output-loading adjustment was difficult and time consuming. The only way to couple output from a cavity is through an inductive link. Those who have experimented with link coupling in hf gear know the frustration involved. These frustrations are even worse in vhf cavity designs because of the limited number of adjustments possible.

Recently published vhf amplifier designs have used stripline techniques with excellent results. Striplines have been used from 50 through 1296 MHz. The ease of tuning and loading the K2RIW 432-MHz stripline amplifier convinced me that this was the way to go; what remained was to adapt the technique to the 8938 coaxial-based triode.

description

The plate circuit is a half-wave stripline with flapper capacitor tuning and loading controls. The grounded-grid triode is cathode driven using a half-wave stripline cathode circuit tuned by a movable capacitor disc. Matching to the driver transmission line is through a variable capacitor also of a movable-disc type. The amplifier enclosure consists of two aluminum sheet-metal chassis boxes similar to the K2RIW amplifier with flat-plate top, middle, and bottom plates. The amplifier is easily built using simple tools and lends itself to simple disassembly for inspection and parts replacement. The only purchased parts were a blower, some brass shim stock, and a few nuts and bolts. The remainder of the parts were either adapted from the junk box or scrounged from friends.

design and construction

The amplifier schematic is shown in **fig. 1**. It is a straightforward grounded-grid triode which is cathode driven using zener bias. The operating voltages required are B+ and filament — period! The plate stripline is located in the upper box 12 inches long x 7 inches wide x 4 inches deep (306x179x102mm), and the cathode circuit is located in the lower box 7 inches long x 5 inches wide x 3 inches deep (179x128x77mm). A 1/8 inch thick (3mm) aluminum center plate divides the two chambers and provides mounting points for the tube socket and the stripline standoff pillars.

Plate circuit. The plate circuit is a half-wavelength stripline made from a 5 x 8 inch (128x204mm), double-clad, 1/8-inch-thick (3mm), glass-epoxy PC board. The corners are rounded to minimize voltage discontinuities. The stripline is located approximately midway between the compartment baseplate and cover, producing a transmission line with a characteristic impedance, Z_o, of about 56 ohms. Capacitor C3 (about 0.5 pF at resonance) tunes the line. The dimensions of C3 are 3 inches (77mm) wide and 3¼ inches (83mm) long. The 1/32-inch-thick (1mm) brass shim stock capacitor overlaps the outer ½ inch (12.5mm) of the plate line.

Capacitor C4 is a beryllium copper flapper ½ inch (12.5mm) by 2 inches (51mm) long, which is soldered at one end to the center contact of the type-N output

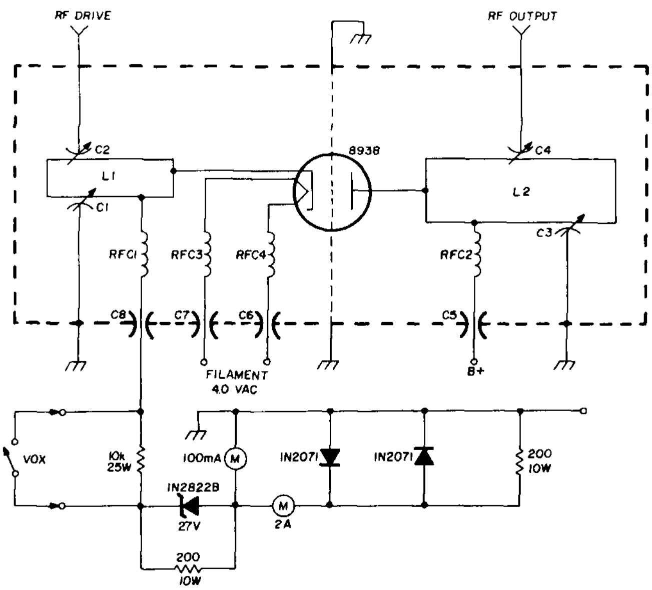


fig. 1. 432-MHz amplifier schematic using stripline construction. Nominal 13-dB gain is provided with this design, which emphasizes reliability for operation under adverse line voltage and temperature conditions.

C1	1 inch (25.5mm) disc on piston tuner
C2	1 inch (25.5mm) disc mounted on transmission line center conductor
C3	Brass shim stock flapper (see text)
C4	Beryllium copper flapper (see text)
C5-C8	Erie 1000 pF feedthrough bypass capacitors
L1	1/16-inch-thick (1.5mm) brass sheet cathode line.
L2	Double-sided PC board plate line
RFC1-RFC4	Rf chokes (see text)

connector. A ¼-inch-thick (6.5mm) plexiglass block bolted to the plate compartment side wall prevents rotation of the output coupling capacitor and type-N connector center pin. Capacitor C4 overlaps the plate circuit by 1 inch (25.5mm).

Both C3 and C4 are positioned by dial-cord strings. The string for C4 passes straight down from the capacitor through a clearance hole in the base plate to a ¼-inch-diameter (6.5mm) steel rod, which is rotated by the plate tuning knob. This control is a National *Velvet Vernier* planetary drive reduction unit. The output capacitor string leads directly upward from the capacitor, passes over a standoff-mounted fairlead, and exits the compartment at the end. It then wraps around a ¼-inch-diameter (6.5mm) steel rod, which is also driven by a National *Velvet Vernier* knob. The capacitors are retracted by the pull of the strings and return due to spring action when the string tension is relaxed. Action is smooth and positive. A string with a good dielectric must be used. Some fly fishing line was tried but melted in the strong rf field!

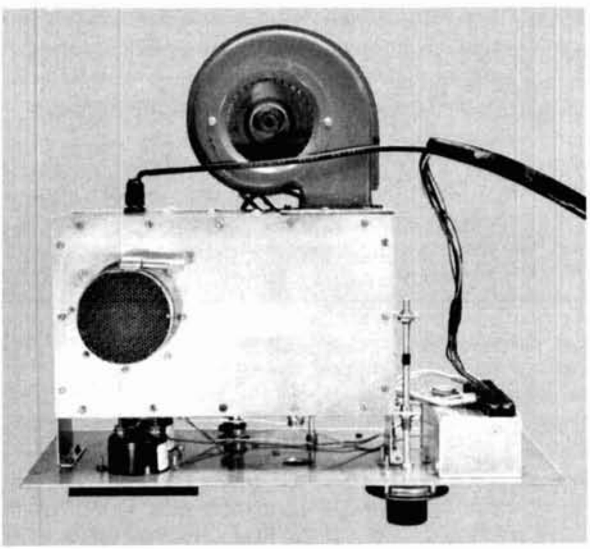
Cathode circuit. The cathode circuit is a half-wavelength stripline counterpoised against one wall of the cathode compartment; it is tuned to resonance by a disc capacitor to ground at the open end of the line. Near this same end of the line, another disc capacitor couples drive power into the cathode.

The cathode line is made of 1/16-inch-thick (1.5mm) brass sheet 4½ inches (115mm) long by 1¼ inch (32mm) wide, which is soldered at one end to the tube socket cathode terminal. A teflon pillar, 3/4 inch (19mm) diameter by 1¼ inch (32mm) long, supports the cathode line at about its physical midpoint.

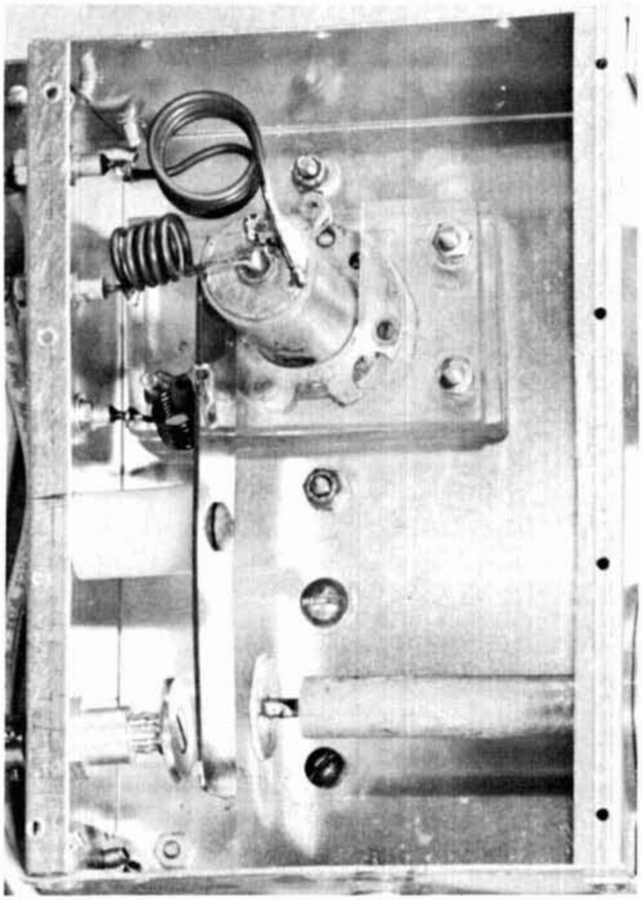
Cathode bias is connected through a feedthrough bypass capacitor and rf choke to the low rf point of the line, which is 1 inch (25.5mm) from the tube socket. The rf choke, RFC2, is made of 5½ turns of number 18 AWG (1mm) enameled copper wire. Diameter of the choke is 3/8 inch (9.5mm). Two large ferrite beads were slipped over RFC2 as added insurance against vhf parasitics.

The tuning capacitor, C1, is a 1-inch-diameter (25.5mm) brass disc mounted on the end of a surplus piston tuner. The tuner has finger-stock contacts to ensure good rf grounding of the capacitor rotating section. The stator is the end of the cathode line. Capacitor C1 is actuated by a panel-mounted knob through a long flexible shaft coupling.

The coupling capacitor, C2, is a 1-inch-diameter (25.5mm) disc of 1/8-inch-thick (3mm) double-clad PC board, which is soldered on both sides to the center conductor of a section of ½-inch-diameter (12.5mm) foam-filled, semirigid coaxial transmission line. The



Amplifier top view showing blower and air exit.



Cathode-tuning box with cover removed (upper left). Coils are rf chokes, which are described in the text.

inner conductor of the hard line projects 3/8 inch (9.5mm) from the outer conductor of the hard line. The hard line passes into the cathode chamber through a flange mount, which is made of 5/8-inch-diameter (16mm), 0.049-inch (1.2mm) wall copper tubing. The tubing is soldered to a brass disc 1/8 inch (3mm) thick, 2 inches (51mm) in diameter. Saw cuts in the end of the copper tube allow the hard line to be clamped into position with a small hose clamp. A second clamp on the hard line prevents the transmission line from sliding in too far and closing the capacitor completely. The coupling capacitor is adjusted for minimum vswr on the drive line with the cathode circuit at resonance.

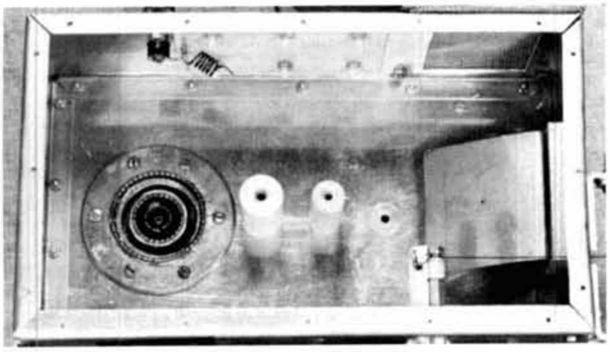
Tube socket. The tube socket was home brewed from a surplus 2C39 socket. Looking down into the plate compartment, the outer ring of the tube socket is the grounded-grid ring. It is made of Instrument Specialties type 97-135 finger stock. The finger stock is soldered into a 2-3/16-inch-diameter (51.5mm) hole in a brass sheet, which is 1/8 inch (3mm) thick and 3-5/8 inches (93mm) in diameter.

The grid ring assembly is bolted to the plate-compartment base. The grid ring mates over a 1-3/4-inch-diameter (45mm) hole in the base plate. The remaining rings in the tube socket (moving inward in order) are the cathode ring, outer filament ring, and center cathode pin collet. The central rings are supported by the tube socket, which is mounted between two layers of ¼-inch-thick (6.5mm) plexiglass. These plexiglass layers insulate cathode and filament from ground.

The 2C39 tube socket adapted for this project has a coaxial set of brass tubes mutually insulated by a mylar sleeve. The outer tube is connected to the cathode ring and cathode line. The inner tube is connected to the outer filament ring. The center pin passes through a fused-glass bead up through the center of the socket. The center pin terminates in the filament-pin collet.

The cathode ring is a 2C39 plate ring (Instrument Specialties part no. 90-70), which is soldered inside a 1-3/8-inch (35mm) OD length of copper tubing. The tubing is shimmed to fit with copper flashing material. The filament ring is made from a short length of 5/8-inch (16mm) OD copper tubing, 0.049 inch (1.2mm) wall, which was slotted with a hacksaw. The tube filament socket fits over the filament ring.

A 4½-turn close-spaced coil 7/8 inch (22mm) in diameter connects to the filament ring, and a 5-turn close-spaced coil, ½ inch (12.5mm) in diameter, connects filament voltage to the center pin. Both coils are made of number 10 AWG (2.6mm) enameled copper wire, and both coils are connected to 0.001-μF feed-through bypass capacitors mounted on the compartment side wall.



Amplifier showing plate-circuit components but with plate line and tube removed. Note tube socket and bypass capacitor arrangement.

Blower requirements. The blower is mounted at the plate-tuning end of the plate-circuit compartment. The air-inlet hole is covered with aluminum insect screen. Air is forced into the plate-circuit compartment by the blower and is restricted to exiting through the anode fins by the plate line and a mylar chimney, which is connected between the anode outer diameter and the top-plate air exit. A Dayton type 2C610 blower is used.

Measurement of the differential pressure across the tube anode, with all air-system components in place, indicated ¼ inch (0.5mm) water-pressure drop. According to the Eimac data sheet for the 8938 tube, this pressure drop corresponds to an air flow of 28 cfm (0.79cmm), which is sufficient for safe operation at 1-kW plate dissipation at sea level.

The mylar chimney is 3½ inches (89mm) in diameter – and 2½ inches (64mm) long. It is formed from 5-mil (0.03mm) mylar. I found that mounting the plate-circuit finger stock upside down created a convenient cup for one end of the mylar chimney.

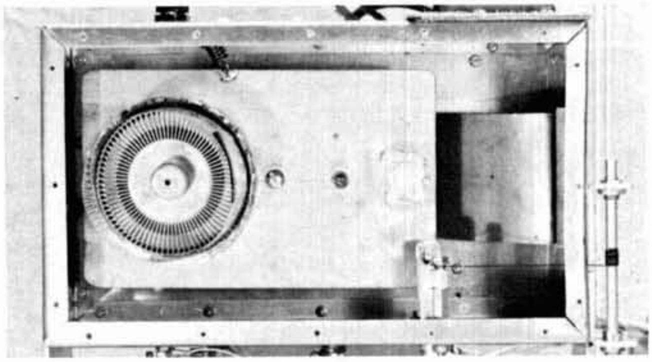


Plate-circuit compartment showing plate line, mylar chimney, and air exit.

The tube was installed into the socket first, then the plate line was pushed down over the tube anode and the plate line was screwed onto the ceramic standoff insulators. The mylar chimney was then placed over the tube anode and carefully fitted into the finger stock cup. The top plate was aligned with the chimney and screwed down, using self-tapping screws. Blind fasteners can be used for an rf tight seal. Remember to solder the plate rf choke to the plate line before securing the cover!

The top cover is a 1/16-inch-thick (1.5mm) aluminum sheet, 1 inch (25.5mm) wide, which is fitted with an air outlet assembly. This assembly is made from a 1/16-inch-thick (1.5mm), 1-inch-wide (25mm) sheet-metal strip, which was bent around the tube anode for sizing, then cut. Several fingers were cut into the strip with a fine-blade saw, then each finger was bent up.

A ½-inch-thick (12.5mm) piece of aluminum honey-comb cut into a disc fits into the outlet assembly. A large stainless-steel hose clamp secures the honeycomb disc inside the outlet assembly. The outlet assembly is clamped to the top cover by a 4½ inch (115mm) square sheet of 1/16-inch-thick (1.5mm) aluminum with a 3-3/8-inch-diameter (86mm) hole cut in the center. The mylar chimney fits inside the air outlet, forming an effective air- and rf-tight enclosure.

power supply requirements

Two-kW PEP input requires a supply nominally rated at 2500 volts at one ampere. The actual plate current at 2 kW input will depend on the value of the zener diode used for cathode bias. A 27-V zener (1N2822B) was used here. With 2200 volts, the amplifier idles at about 50-70 mA and whistles up to 800 mA on SSB or key-down CW tune position. The 1760 watts input yields over a kW output. Operation is just on the "A" side of class B. On-the-air reports are of good quality SSB.

The tube manufacturer recommends 4.0 Vac on the filament for 432-MHz operation. This can be supplied by a 5-V transformer with *Variac* adjustment of the primary voltage.

The power supply used for the South American operation was built by W3HQT. It uses *Variac* control on both the B+ and filament supplies. It is detailed schematically in **fig. 2**. The B+ primary power circuit is circuit-breaker protected. The blower comes on with the filament switch.

bias and metering

The bias and metering circuits are pretty standard for grounded-grid amplifiers.[2] A 10k, 25-watt resistor in the cathode bias lead provides cutoff bias during standby periods. In the operate mode, this resistor is shorted by a set of contacts tripped by the ssb vox circuit. Operating bias is provided by the 27-volt zener connected in series with the cathode and B–.

The grid current is measured by a 0-100 mA meter in the cathode-to-ground (grid) lead. Plate current is

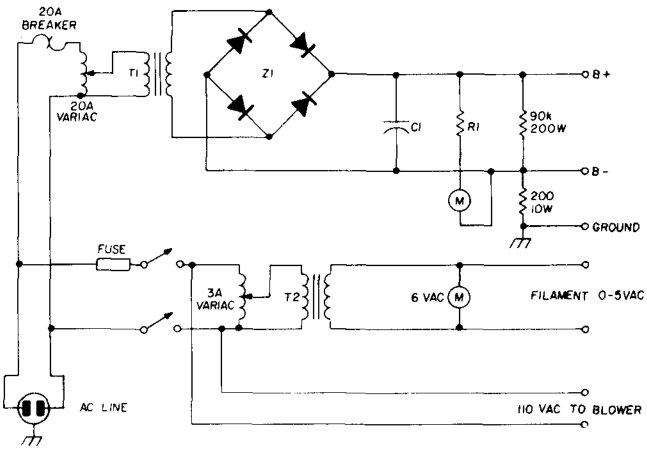


fig. 2. Power supply for the 432-MHz rf power amplifier. The 10 megohm resistor for the B+ meter is made from ten series-connected 1 megohm, 2 watt resistors.

measured in the B– lead by a 0-2 A meter. Two 0-200 μA API meters were used with appropriate shunt resistors. A 200-ohm, 10-watt resistor provides B– reference to ground should the grid meter open up. A second 200 ohm, 10-watt resistor prevents the cathode from rising to a high potential in the event the zener burns open.

B+ bypass capacitor

The B+ lead to the plate compartment is bypassed by a homemade capacitor consisting of a 0.015-inch-thick (0.38mm) piece of Teflon sheet sandwiched between a single-sided piece of PC board and the compartment wall. The PC board is 3-1/2 inches x 4-1/2 inches (89x115mm) with a flat ground lug soldered to the copper foil at one end. This lug is connected to the high-voltage connector. The foil side of the board is in contact with the teflon dielectric, which overlaps the PC board by 1/4 inch (6.5mm) all around. Six nylon screws bolt the PC board to the compartment side wall.

The plate rf choke is 5 turns of number 16 AWG (1.3mm) enameled copper wire connected between the high voltage connector and soldered to the plate line at the low voltage point, which is located approximately at the inboard edge of the tube anode.

operation and adjustment

Never operate the 8938 tube with rf drive but without B+ voltage. In operation, both the B+ and filament-voltage *Variacs* are turned to zero before turning on the power switches. The filament switch turns on the blower, after which the filament voltage is increased to 4.0 volts with its *Variac*.

Turn on the B+ supply and set the voltage to 1000 volts. Apply operating bias by shorting the standby resistor and observe idling current level (it should be about 35 mA with 1000 V B+ and 50-70 mA with 2000 V). Apply about 25-50 watts of drive and resonate the plate and cathode tuning capacitors, C1 and C2.

Load the amplifier by adjusting output-coupling capacitor C4 a small increment at a time while re-resonating the tuning capacitor each time the load capacitor is adjusted. Adjust for maximum output. Now adjust the rf-drive coupling capacitor, C2, in small increments and re-tune the cathode tuning capacitor each time until minimum vswr occurs on the drive line. Minimum vswr should easily be below 1.5:1.

With all controls adjusted, increase B+ to the operating level and readjust all controls for the new operating conditions. Adjust C3 and C4 for maximum output. Adjust C2 for minimum input vswr when C2 is tuned for maximum grid current.

When shutting down, remove B+ and turn down filament voltage to zero. After a few minutes cooling-off time, the blower may be switched off.

performance

Two-kW PEP input is achieved at 60% plate efficiency with about 13 dB power gain as a minimum. Since the calibration of power meters, even those of good repute, is often in doubt, this performance figure is conservative.

No difficulties were experienced with this amplifier during the South America EME expedition. It ran reliably and well during long days of moonbounce schedules under adverse line voltage and temperature conditions.

acknowledgement

I wish to thank Walt Bohlman, K3BPP, Bill Olson, W3HQT, and Bill Wenner, W3IVL for their help with the design and checkout of the amplifier, and Dick Knadle, K2RIW, who reviewed the manuscript.

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1. R. T. Knadle, K2RIW, "A Strip-Line Kilowatt Amplifier for 432 MHz," *QST*, April and May, 1972.
2. R. I. Southerland, W6UOV, "High-Performance 144-MHz Power Amplifier," *ham radio*, August, 1971.

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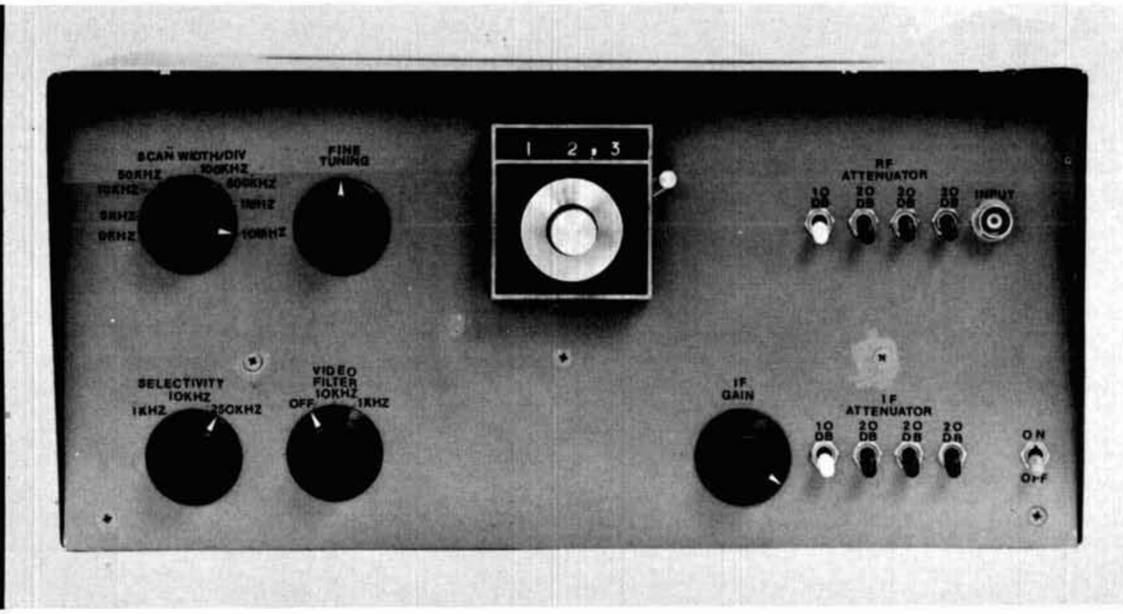


high-performance spectrum analyzer

These instruments have been limited mainly to professional labs — now you can build your own spectrum analyzer from the information furnished here

A spectrum analyzer is a radio receiver with a swept local oscillator that allows continuous tuning over a specified frequency range. The received signals are displayed on a conventional oscilloscope as pips. **Fig. 1** shows the radio spectrum between 100 kHz and 100 MHz in the San Francisco area as displayed on the spectrum analyzer described in this article.

Because of their cost and complexity, good spectrum analyzers have been limited primarily to research and development activity. Some military instruments have shown up on the surplus market, but poor sensitivity and selectivity, images, spurious responses, and poor dynamic range have limited their usefulness.

Considerable time was spent in the design of the spectrum analyzer described here. Even more time was required to assure circuit reproducibility, minimize the variety of parts, and select the least-expensive components. No PC boards are used in the design nor are any planned. PC boards in the rf and i-f strips, unless very carefully made, would probably degrade the performance of the instrument. Complete design, construction, and testing details are included for those wishing to build the spectrum analyzer.

applications

The spectrum analyzer can be used to observe harmonics, parasitic oscillations, and sidebands of CW, a-m, fm or ssb signals. Propagation conditions in terms

By **Wayne C. Ryder, W6URH**, 115 Hedge Road, Menlo Park, California 94025

of station activity can be observed by looking at the number of signals on an amateur band. This suggests another use for DX and contest operating: You can immediately see where the pileups are without cranking the receiver tuning dial back and forth. The spectrum analyzer described in these pages was built to look for the source of birdies in an experimental general-coverage radio receiver.

The conventional way to observe radio signals is in the time domain on an oscilloscope. This is a good method when no harmonics, parasitic oscillations, or other signals are present. However, amplifiers can oscillate, modulators or mixers might not reject signals being modulated, oscillators may be oscillating on more than one frequency, or detectors may be passing the signal being detected. These signals might appear on an oscilloscope in the time domain as confusing pictures, as shown in **fig. 2A**. **Fig. 2B** shows a signal and behind it the second harmonic. This is the frequency domain. If a signal fundamental is on 5 MHz and the second harmonic in on 10 MHz, these appear on a spectrum analyzer as pips at 5 and 10 MHz. Instead of seeing a complex waveform in the time domain as on an oscilloscope, the signal is viewed in the frequency domain. The amplitude and frequency of each component can be seen (**fig. 2C**).

three kinds of spectrum analyzers

Fig. 3 shows a real-time analyzer. The incoming signal is connected to a series of filters followed by detectors. A scan or sweep generator drives the horizontal plates of an oscilloscope and controls an electronic switch, which selects the proper detector output. The frequency range is limited by the number of filters and their bandwidth. This type of analyzer is quite expensive because of the large number of filters required to cover a spectrum. Its main application is in the audible or subaudible range.

Another type is the tuned radio-frequency analyzer shown in **fig. 4**. The input filter is a tunable bandpass

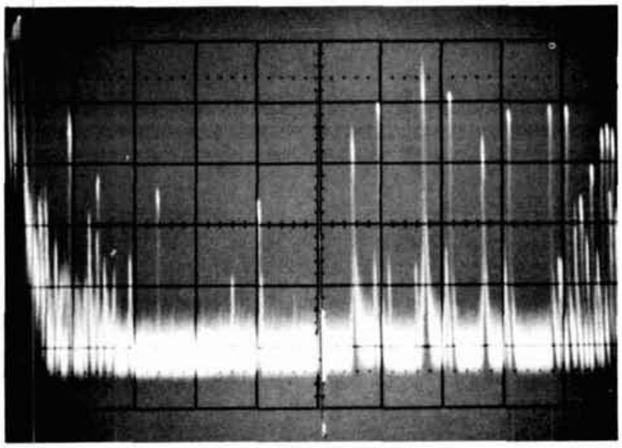


fig. 1. Display showing the San Francisco-area radio spectrum on the spectrum analyzer. A TV antenna was used to receive the signals on the right half of the display, and the TV antenna lead was used as a long-wire antenna to receive signals on the left half.

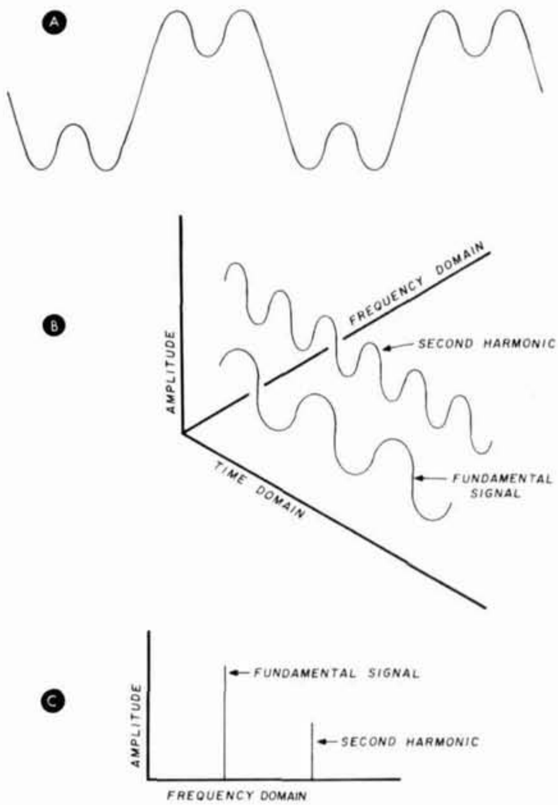


fig. 2. A fundamental-frequency signal and its second harmonic as shown on an oscilloscope in the time domain, A. The same signal and its second harmonic are shown in B in the frequency domain to illustrate their relationships. The spectrum analyzer displays a fundamental signal and its second harmonic in the frequency domain as shown in C. Amplitude and frequency of each can be seen.

filter, which is tuned by the scan generator. The TRF analyzer lacks resolution and sensitivity. Also, tunable filters usually don't have constant bandwidth. The TRF analyzer is used mainly for microwave analysis. There are other types of analyzers, such as a digital processor that performs Fourier analysis, but these are very complex.

The most common type of spectrum analyzer is the superheterodyne, as shown in **fig. 5**. It is similar to a standard superheterodyne receiver with the following exceptions: There is no tuned circuit in the front end, because it would be difficult to make one continuously tunable from 100 kHz to 100 MHz, and the local oscillator is electronically tuned by the scan generator from the oscilloscope. The superheterodyne analyzer has the advantage of a wide tuning range and controlled bandwidth. Also the sensitivity can be made uniform. Image problems are minimized by triple conversion.

Fig. 6 shows the spectrum analyzer described here. No more than 1 volt rms should be applied to the input attenuator. The output from the attenuator should be limited to about 10 mV rms maximum. The function of each assembly is discussed with reference to the block diagram. Later, theory of operation, circuit design, construction, and checkout are described for each assembly.

The input attenuator is followed by an LC 130 MHz lowpass filter. Its purpose is to prevent incoming signal harmonics from mixing with the first local oscillator and producing spurious responses. The incoming signal, 0.1 - 100 MHz, is mixed with the vco to produce a 200-MHz first i-f. The vco is driven by a sweep shaper that predistorts the sawtooth wave from the oscilloscope to compensate for tuning nonlinearities of the varactor tuning diode in the vco.

The 200-MHz i-f output from the first mixer (all mixers are double balanced) goes to an amplifier, which compensates for the mixer loss. Next is a 200-MHz bandpass filter followed by the second mixer. The 200-MHz i-f signal mixes with the second local oscillator output to provide a 50-MHz i-f signal. This signal is amplified and applied to a 50-MHz bandpass filter. The output of this bandpass filter is mixed with 39.3 MHz to produce a third i-f of 10.7 MHz. The 10.7 MHz output from the third mixer is then amplified and fed through a 250-kHz bandpass ceramic filter. This stage is followed by an i-f attenuator.

Next are two identical crystal filters to provide 1- and 10-kHz bandwidths. These feed a gain equalizer, which compensates for the differences in gain between the various bandwidths. Completing the i-f section is an amplifier that drives a second 250-kHz bandpass ceramic filter.

The logarithmic amplifier compresses the output signal so that a large-amplitude signal can be displayed on an oscilloscope. Two divisions on the scope correspond to a change of 10:1, or 20 dB. Six divisions represent a change of 1000:1, or 60 dB. Without the log amp, a CRT about 30 inches (76cm) tall would be required to display the same information with the sensitivity available at the top of the display. A video filter is included to remove high-frequency noise in narrow-bandwidth operation. The performance specifications for the spectrum analyzer are listed in **table 1.**

table 1. Performance specifications of the W6URH spectrum analyzer.

frequency range:	0.1 MHz - 100 MHz.
sensitivity:	10 μV or less in the 1-kHz position.
selectivity:	Approximately 1, 10, or 250 kHz (switchable).
sweep width:	0, 5, 10, 100, 500, 1000, and 10,000 division. In the 10,000 kHz or 10 MHz/division position, the center frequency is set to 50 MHz. Otherwise, frequency is controlled by the center control on the front panel.
first LO frequency jitter:	≈ 5 kHz
signal separation:	10 kHz for 40-dB resolution in 1-kHz position.
input impedance:	50 ohms
response 100 kHz - 100 MHz:	±2 dB
frequency accuracy:	±3 MHz.
video filter:	10 and 1 kHz.
shape factor 3 - 60 dB:	250 kHz 2.5:1 10 kHz 50:1 1 kHz 30:1

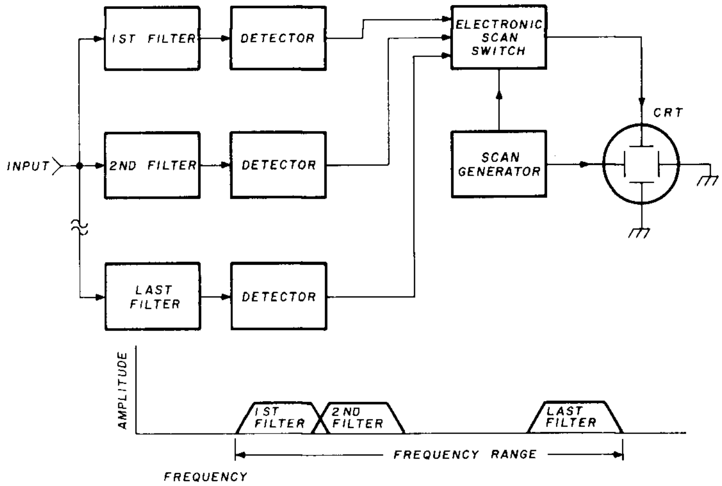


fig. 3. A real-time spectrum analyzer. Filters are spaced to provide continuous coverage of the frequency spectrum.

The display scope for the spectrum analyzer should be dc coupled, have 0.1 V/division vertical sensitivity, and have six divisions of vertical calibration by ten divisions of horizontal calibration. Also approximately 6V of horizontal sawtooth voltage should be available to drive the spectrum analyzer from the scope. Scope input impedance should be 1 megohm and input capacitance 10 - 40 pF.

circuit description

This section presents the theory of operation and design details of each of the major circuits in the spectrum analyzer. It is recommended that the following text be read before attempting to build the circuit modules as much information is provided that will be helpful when construction is started. Especially important are the suggestions on decoupling, shielding, and bypassing.

Sweep shaper. The sweep-shaper schematic is shown in **fig. 7.** The resistors in the input circuit, which are switched by S601A, determine the sweep width. In the 10-MHz/division position, the sweep center frequency is set to 50 MHz by R101. In all other positions the sweep center frequency is set by R601. R104 sets the low-frequency limit, 0.1 MHz, and R102 sets the high-frequency limit, 100 MHz. R103 is a fine frequency adjustment with about 500 kHz of range. U101 provides isolation for the input attenuator and some gain. U102 provides isolation for the frequency control. The outputs of U101 and U102 are resistively combined and drive U103, which is a nonlinear amplifier that complements the nonlinearities of the tuning diode, CR203, in the vco.

As the frequency is increased, more voltage is required. If a change of 1 volt is required to go from 260 to 270 MHz, a change of 1.3 volts may be required to go from 270 to 280 MHz. **Fig. 8** gives an idea of the distorted waveform available from U103 and the actual waveform required to compensate for nonlinearities of the tuning diode in the vco. Compensation is usually required between 270 and 300 MHz. When the voltage

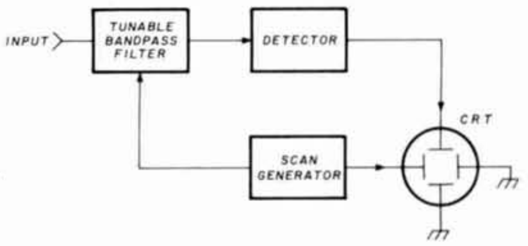


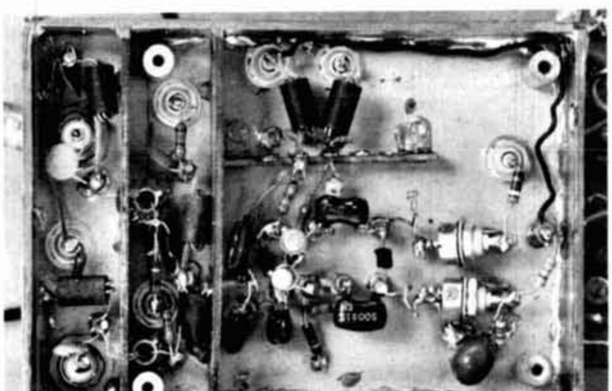
fig. 4. Tuned radio frequency (TRF) spectrum analyzer. Tuning is accomplished by adjusting the bandpass-filter frequency.

on pin 3 of U103 reaches the voltage on the right side of R115, gain will be increased by the setting of R115, and so on up the line, to R106. R115 sets the nonlinearity at 10 MHz and R106 at 100 MHz. The tuning diode used in the vco I built was linear up to about 280 MHz.

Care must be taken so that no low-frequency components, including power-supply hum, are added to the incoming sawtooth wave by the sweep shaper. Added low-frequency components will result in first local-oscillator instability. This shows up as jitter in the display when looking at narrow scan widths.

Voltage-controlled oscillator. Q201 and Q202 together with L201, L202, and CR203 form a 200-300 MHz voltage-controlled oscillator (vco – **fig. 9**). Q203 is a buffer amplifier. Q204 and Q205 provide outputs to the first mixer and a second output.

Several compromises were made in the design of the vco. To achieve high frequency stability, the oscillator should have a high C/L ratio; however, to tune it with a varactor, a low C/L ratio is required. Varactor circuits with reasonably high Q at these frequencies have a relatively small tuning range. When considering tuning range, linearity, and Q the Motorola Epicap tuning diodes seem to be about the best. Careful design of the oscillator and amplifiers was necessary to provide a constant output level to the first mixer. The 2N2369A did not have sufficient gain-bandwidth product to provide constant output when used as an oscillator transistor, but because of the low gain requirements of the amplifiers, the 2N2369A is satisfactory in these circuits.



Closeup view of the voltage-controlled oscillator. Main circuit is in the compartment to the right; connectors for output to the first mixer and tracking generator are in compartment on left.

Rf section. The rf section (**fig. 10**) contains the first two i-f stages operating at 200 and 50 MHz. The third i-f, operating at 10.7 MHz, is contained in the i-f section.

The input signal is applied to the first mixer, MX301, through an rf attenuator and a lowpass filter. The attenuator is required to keep the input to the first mixer below approximately 10 mV. MX301 combines the 0.1 - 100 MHz input with the 200 - 300 MHz signal from the vco to produce the first i-f of 200 MHz. Q301 provides gain to compensate for the loss through MX301. L305 through L307 form a 200-MHz bandpass filter. Output from the 150-MHz local oscillator, Q305, mixes with the 200-MHz signal in MX302 to produce a 50-MHz i-f. The 50-MHz signal is amplified by Q302 and Q304 then applied to the third mixer, MX303, through a four-section bandpass filter. The signal is then mixed with the 39.3-MHz output from Q306 to produce 10.7 MHz.

Several steps were taken to minimize spurious and image responses. The input to the pass filter attenuates harmonics that would otherwise mix with the first local oscillator signal, producing spurious signals. Using a first i-f twice, the highest input frequency minimizes basic image problems. Intermediate frequencies separated 5:1

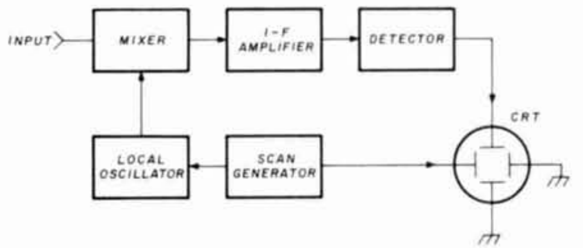


fig. 5. Simplified block diagram of a superheterodyne spectrum analyzer. Local oscillator is tuned by the scan generator.

or less from the next i-f further reduce any images. Double-balanced, four-diode mixers produce the fewest unwanted products and have relatively high local-oscillator isolation. Shielding and bypassing prevent the three local oscillators and their harmonics from mixing and producing spurious signals. For example, with inadequate shielding, the 39.3-MHz local oscillator signal would leak back through the front end. A local signal on 6 meters would be picked up by the second i-f (50 MHz). Ideally each section should be in a die-cast box with an rf gasket, but the construction with copper-clad board, described later, is adequate if care is taken to ensure an rf-tight enclosure.

A ferrite bead is used on all dc leads on each side of the feedthrough. Piston trimmers are required for the 200-MHz bandpass filter for mechanical rigidity. Links L309 and L311 should be constructed so they can be moved from 0 to ½ inch (0 to 12.5mm) away from their respective coils.

I-f section. The third-mixer output, 10.7 MHz, is amplified by Q401 (**fig. 11**). Q402 provides drive at 300 ohms to FL401. Next, an emitter-follower, Q403, provides drive at 50 ohms to the i-f attenuator. The two crystal filters are identical, and the first, Y401, is described.

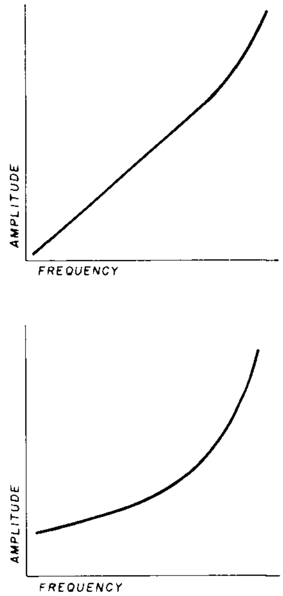


fig. 8. Top curve shows wave shape available from U103 in the sweep shaper; bottom curve shows actual wave shape required to compensate for the nonlinearity of the tuning diode, CR103, in the vco.

Q404 provides a paraphase output, which drives Y401 and neutralizes its capacitance. Q405 provides a high input impedance to the crystal-filter output. CR401 provides a bypass around Y401 for 250-kHz bandwidth. CR402 switches in R401 to broaden the crystal filter for 10-kHz bandwidth. R402 adjusts the gain in the 1-kHz bandwidth position, and R403 adjusts gain in the 10-kHz bandwidth position to the same as in the 250-kHz bandwidth positions. CR403 switches in R403 in the 10-kHz bandwidth position. Q408 and Q409 provide a gain of about 40. Q409 provides 300-ohm drive to FL501 in the log-amp section.

FL401 and FL501 provide 250-kHz bandwidth. After manufacture, these filters are separated into groups and are color coded as to their actual center frequency, which are around 10.7 MHz. Because of spurious resonances above the crystal-filter frequencies, *orange and only orange* color-coded filters maybe used. With the orange-colored ceramic filters, crystal-filter spurious responses are down -50 to -60 dB. With opposite-end ceramic filters, they may be down only 15 dB.

One of the more challenging aspects of designing a spectrum analyzer is selectivity. In its widest position, the spectrum analyzer should have a bandwidth to observe signals using a sweep width of 100 MHz. In its narrowest position, it should be able to resolve signals close together, such as a carrier and its sidebands. Spectrum analyzers with elaborate sets of carefully matched crystal filters can achieve bandwidths as low as 10 Hz at the 3-dB points. These crystals are especially ground to

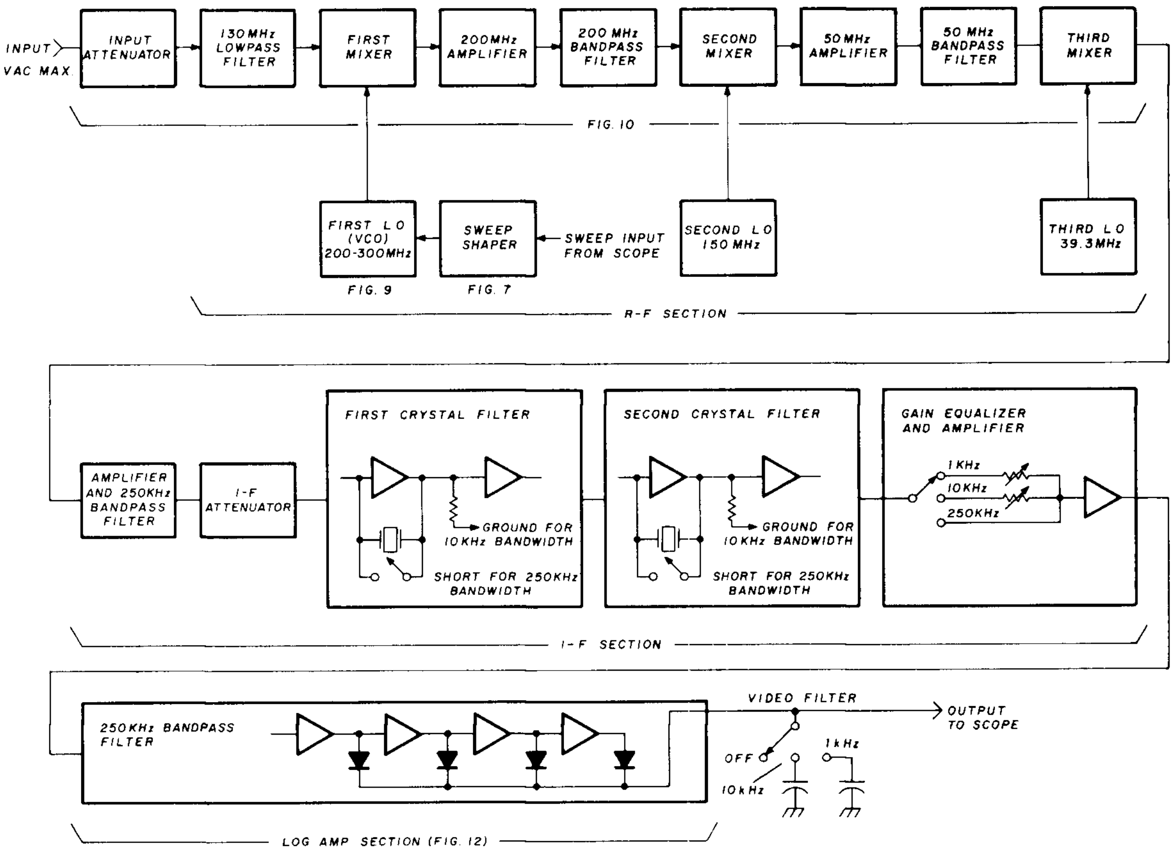


fig. 6. Complete superheterodyne spectrum analyzer described here. Each block represents an individually shielded section or assembly.

minimize resonances other than those desired for crystal filters.

Another important consideration is shape factor and ringing. Communications receivers often have a 3-60 dB shape factor of 2:1. Unfortunately, these filters with steep skirts have phase discontinuities at their band edges, and therefore ringing occurs when signals are rapidly swept through them. This phenomenon was demonstrated using a crystal filter from an fm transceiver in the spectrum analyzer. Shape factor of narrowband filters used in spectrum analyzers cannot be as low as that found in communications receivers.

The ceramic filter used in the 250-kHz bandwidth position has a shape factor of about 2.5:1. These filters are used mainly in fm automobile radios.

The crystal filters were made using off-the-shelf crystals from the local Heathkit store. The filters exhibit some resonance, −50 to −60 dB, slightly above the filter frequency. With a shape factor of about 30:1 two signals about 10-kHz apart have about 40 dB of resolution. The filters are not sufficiently narrow to look at 1- or 2-kHz sidebands of a transmitter. To observe signals which are separated by only 1 kHz would require a 500-Hz filter with a shape factor of 2:1, a sweep speed of 5 sec/cm, a

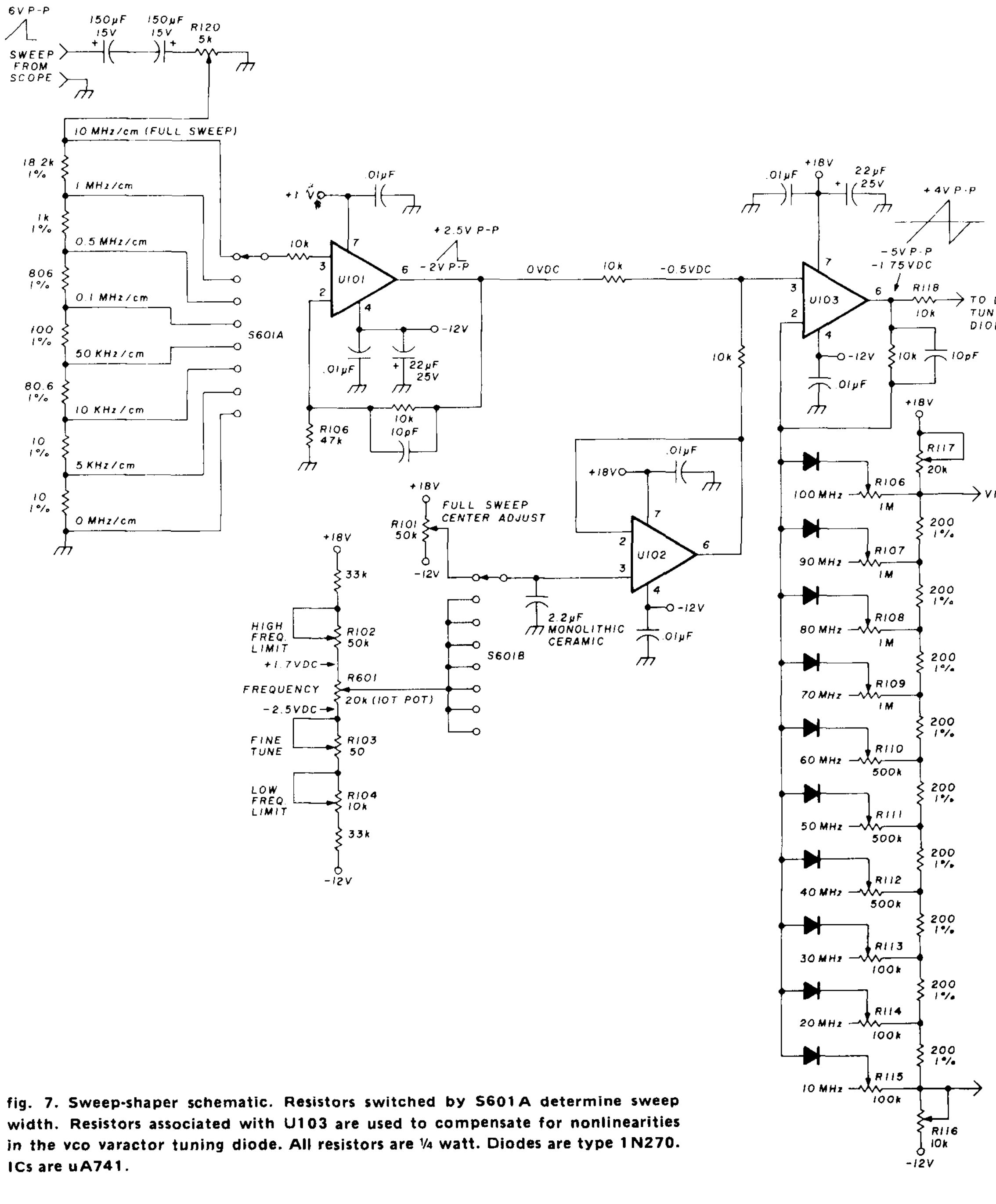


fig. 7. Sweep-shaper schematic. Resistors switched by S601A determine sweep width. Resistors associated with U103 are used to compensate for nonlinearities in the vco varactor tuning diode. All resistors are ¼ watt. Diodes are type 1N270. ICs are uA741.

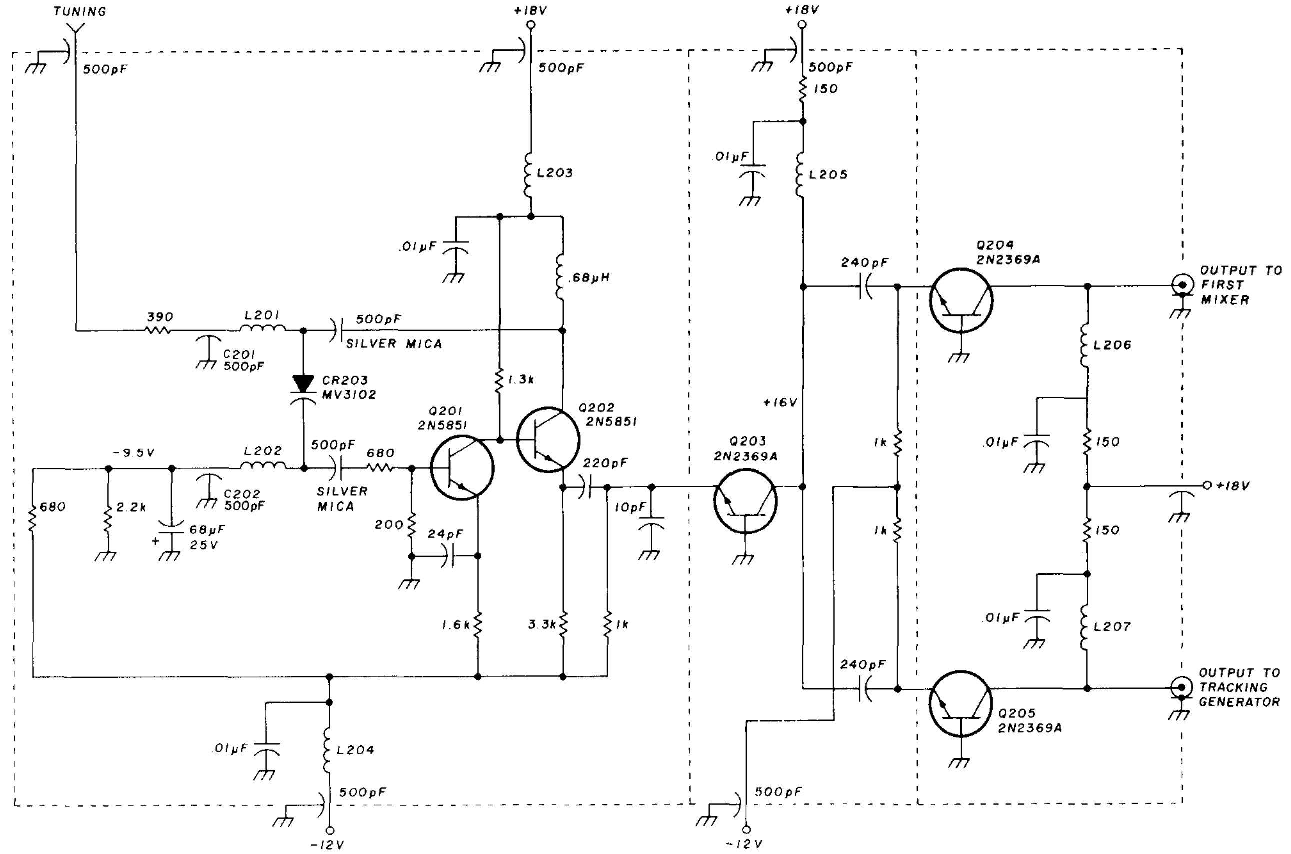


fig. 9. VCO schematic. Range is 200-300 MHz. CR203 is a Motorola MV3102. C201 and C202 are 500-pF feedthrough bypass capacitors. L201, L202: 1 turn no. 18 AWG (1mm) wire. L203-L207: 5½-turn chokes.

phase-locked vco to remove vco fitter, and a storage scope because of the very slow sweep speed.

Log amplifier. The log amp schematic is shown in **fig. 12.** FL501 provides selectivity in the 250-kHz position. Emitter-follower Q501 provides a low-impedance input for Q502. Q502 through Q505 are a series of four wideband resistance-coupled amplifiers. Each stage has a gain of 6 for a total gain of 1296, or slightly over 60 dB.

The detected output of each stage is summed through each diode at the output. As the signal level at the output of Q505 increases, CR505 conducts with a logarithmic-shaped curve. Q505 then saturates and CR504 starts conducting, and so on down the line. CR502 and CR503 are in series to provide a voltage greater than that of CR503 and CR504. R501 through R503 set the output level from each stage, and R504 sets the total output level. The video filter reduces high-frequency noise. It can be used only for narrow sweep widths. The video filter is described in detail in the section on operation.

construction

All assemblies except the sweep shaper are built in boxes made of 1/16-inch thick (1.5mm) single-sided copper-clad board. The rf assembly has double walls, as explained later, and individual covers. All other assemblies have single-wall separators and one cover for the entire assembly. All boxes are 3/4 inch (19mm) high (inside measurement), and the covers are secured with 3/4-inch long (19mm) 4.40 (M3) metal spacers. The following parts had to be obtained from the sources shown. All other parts came from the junk box and the sources are unknown.

component	source
ceramic filters (FL401, FL501) Vernitron FM-4	Vernitron, 232 Forbes Road, Bedford, Ohio 44146
crystals (Y401, Y402) Heathkit 404-39	local Heathkit store
2N2369A, J309 Siliconix	James Electronics, P.O. Box 882, Belmont, CA 94002
one-hole beads (Amidon FB-43-101) six-hole bead (Amidon FB-64-5111)	Order from Amidon Assoc., 12203 Otsego St., North Hollywood, CA 91607

Sweep shaper. This assembly is built on copper-clad boards without shielding or feedthrough bypassing.

vco. The vco is built in a copper-clad box measuring 4½x3x¾ inches (114x76x19mm). The buffer section and output amplifiers are shielded from the oscillator. Paper-thin copper is wrapped over the outside edges where the cover attaches.

rf section. The rf section has double walls between compartments, which ensures good shielding. The walls are spaced 1/16 inch (1.5mm) apart, and paper-thin copper (available from hobby shops) is laid over the walls. The covers for the two bandpass filters are cut as shown in **fig. 13** (a continuous shield will create a shorted turn, which will seriously detune the filters). All assemblies have 500-pF feedthrough capacitors for dc voltages and 0.01-μF capacitors across the feedthrough caps inside the compartments. Dimensions of the rf section are shown in **fig. 14**.

The construction of the three mixers is identical. The *transformers for each unit use four trifilar-wound turns.* All windings are wound at the same time for a total of 12 turns. **Fig. 15** shows four turns through a ferrite bead. The four turns are counted from the inside — *not* the outside. Also shown in **fig. 15** are the transformer connections and how they are installed in the mixer. All *ports have baluns consisting of two bifilar-wound turns* on ferrite beads. The same type bead is used for the transformers and baluns.

I-f section. Dimensions for this assembly are shown in **fig. 16**. It has one large cover rather than individual covers as used in the rf section.

power supply. The power supply is located beneath the chassis at the rear. Two separate supplies are required (**fig. 17**). Transformer T701 should provide 23-29 Vdc at approximately 200 mA and T702 should provide 18-24 Vdc at 100 mA. Both regulators are commonly available three-terminal ICs. The supplies must not go out of regulation above about 100 Vac line voltage.

checkout

Before any tests are made, all dc voltages should be checked against those shown in the schematics. In many instances, each stage of an assembly is checked. In each case, the generator should be terminated and coupled through a 0.01-μF capacitor. The existing input to the stage should be removed during these checks. The generator should then be carefully tuned to the frequency that produces maximum deflection on the display scope. Stage-by-stage checkout is important because it is otherwise difficult to determine which stage is at fault if performance is not satisfactory. Assembly checkout is followed by a final alignment.

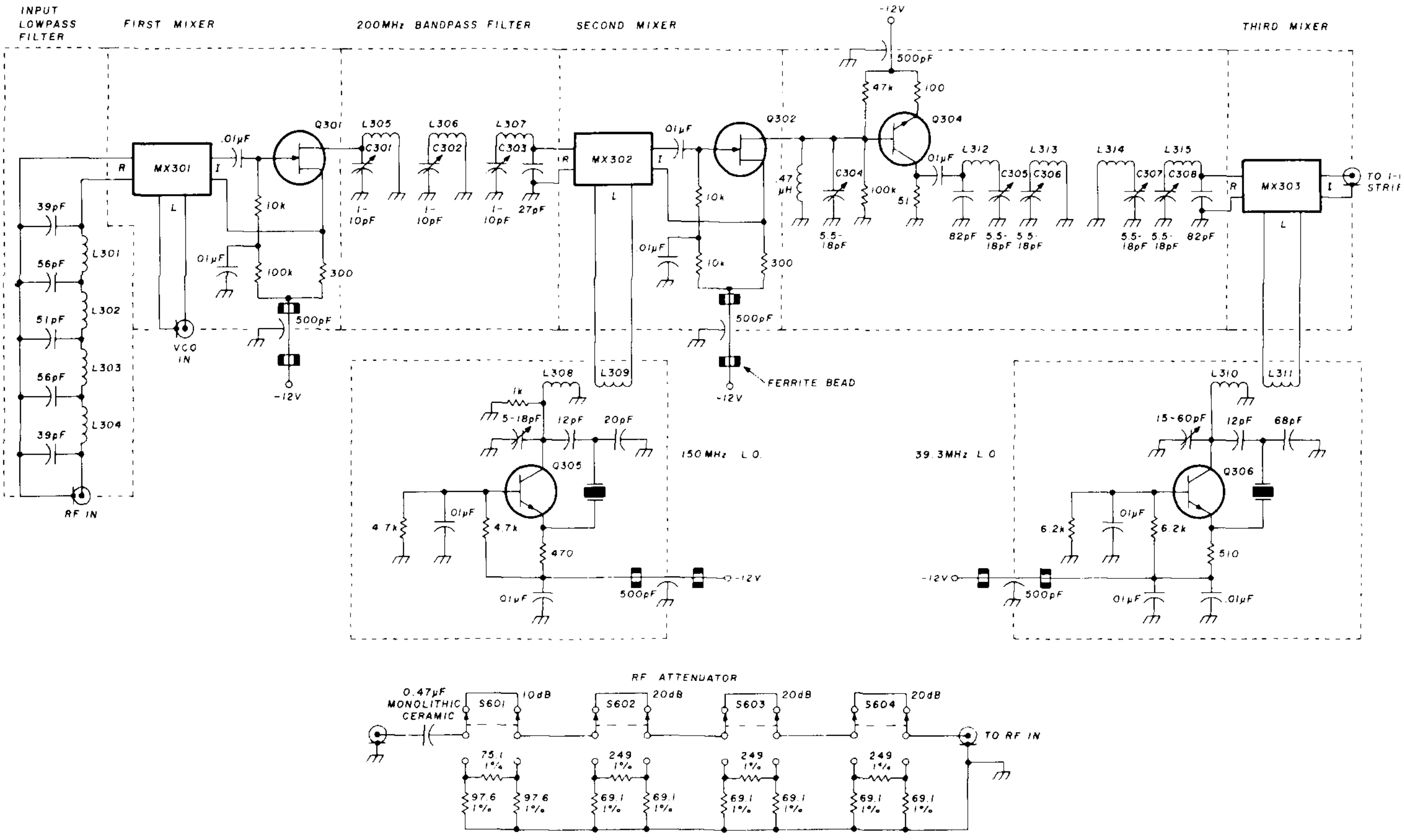


fig. 10. Rf-section schematic. All bipolar transistors are 2N2369As; fets are J309 or J310. Feedthrough caps are 500 pF.

L301-L304	4 turns no. 18 AWG (1.0mm), 3/16" (5.0mm) diameter, 1/4" (6.5mm) long
L305-L307	4 turns no. 22 AWG (0.6mm), 1/4" (6.5mm) diameter, 3/16" (5.0mm) long. Spacing between coils 1/4" (6.5mm). All wound on one 1/4" (6.5mm) diameter plastic form.
L308	3 turns no. 18 AWG (1.0mm), 1/4" (6.5mm) diameter, 1/4" (6.5mm) long
L309	2 turns no. 22 AWG (0.6mm), 1/4" (6.5mm) diameter, close wound
L310	10 turns no. 18 AWG (1.0mm), 5/16" (8.0mm) diameter, close wound
L311	2 turns no. 22 AWG (0.6mm), 1/4" (6.5mm) diameter, close wound
L312-L315	7 turns no. 26 AWG (0.3mm), 1/4" (6.5mm) diameter, close wound. Spacing between coils is 1/2" (12.5mm)

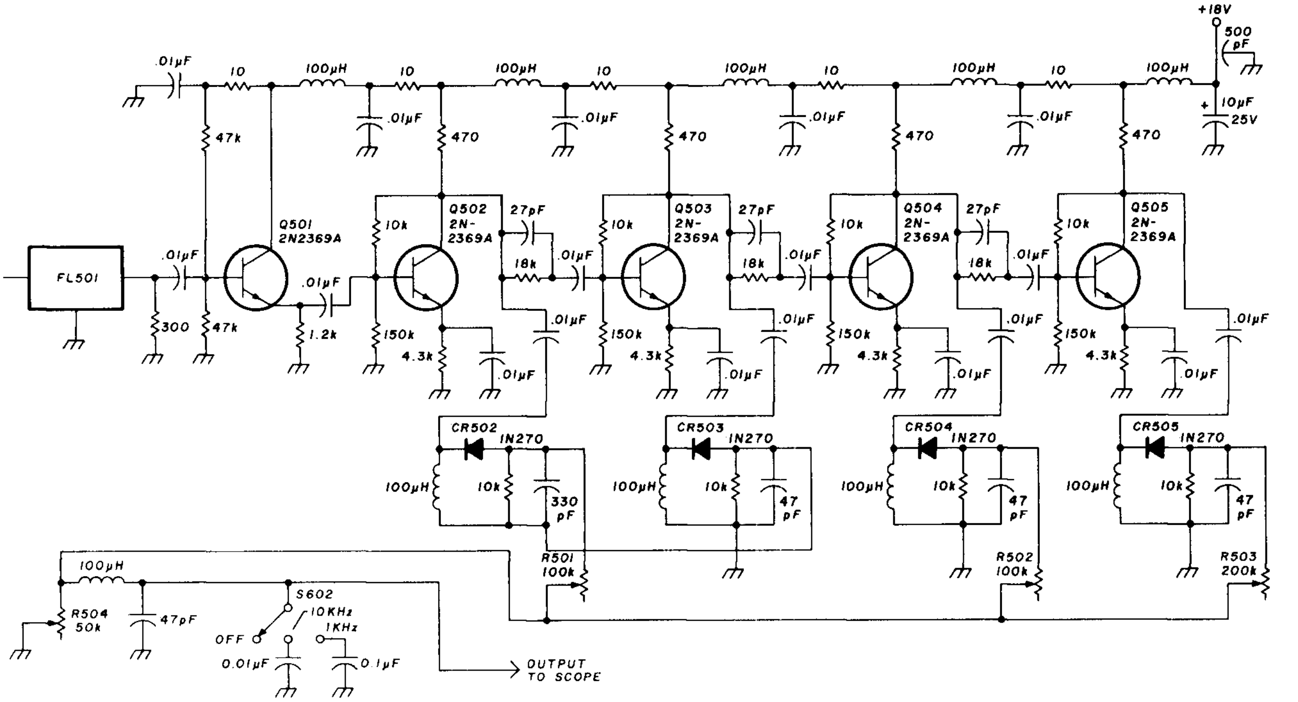


fig. 12. Log amplifier schematic.

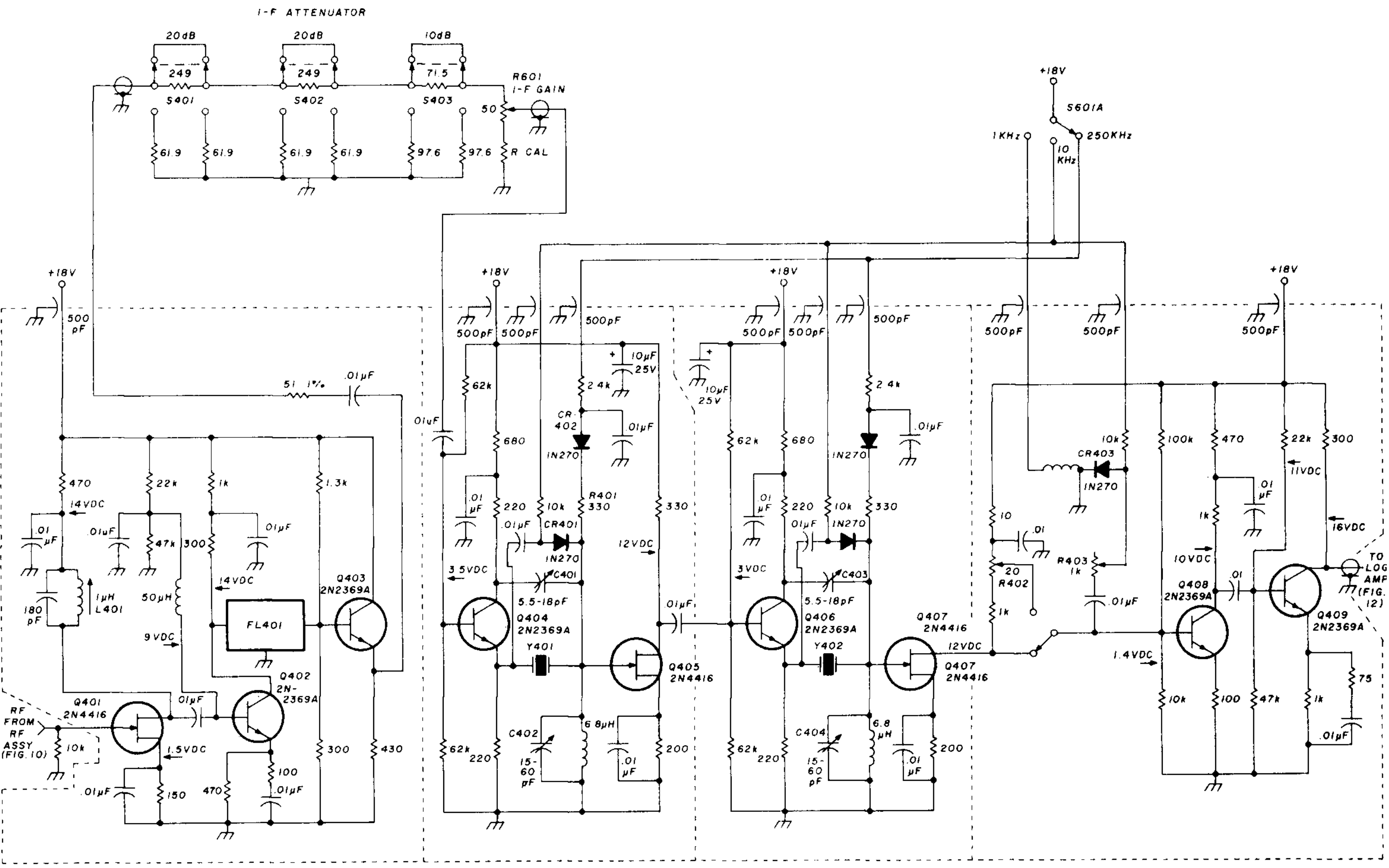


fig. 11. I-f section schematic. All resistors in the attenuator must be 1% tolerance. Resistors must be added in series or parallel to obtain 51 ohms ±1% with R601. Feedthrough caps are 500 pF.

test equipment required

1. Rf signal generator, 10-300 MHz, with calibrated attenuator 1 μV to 0.1V rms (Measurement model 80 or HP 608 or equivalent).

2. Grid-dip oscillator, 10 MHz - 200 MHz (Heath GD1 or equivalent).

3. Rf signal generator, 100 kHz - 10 MHz. Calibrated attenuator 1 μV - 0.1 V desirable (General Radio 605 or equivalent).

4. Volt-ohmmeter (Triplet 630 or equivalent).

5. Comb generator. (Prescalar described in June, 1975, *ham radio*.[1] not absolutely necessary but convenient for adjusting vco linearity).

6. Frequency counter capable of counting to 300 MHz.

Sweep shaper and vco. Verify the voltages at the output of U101 and U103 against those shown in the schematic, **fig. 7**. Using a 300-MHz counter, determine what tuning voltages in the vco produce 200, 250, and 300 MHz. (Further checkout of the vco and sweep shaper is described in the section on final alignment).

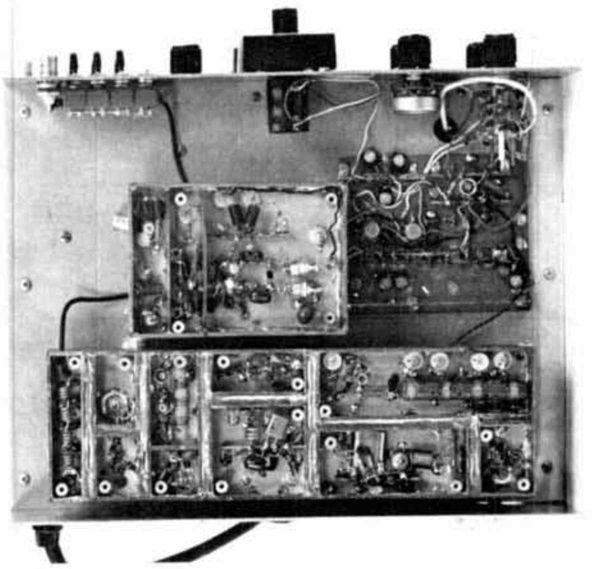
Log amplifier. Check dc voltages as shown in **fig. 12**. Ensure that 2-3 volts are between collector and emitter of Q502-Q505.

1. Set all pots to center range.

2. Connect generator to log-amp input.

3. Set display scope to 0.1 V per division.

4. Set generator for 300 μV output and tune generator for peak on display. Verify according to the chart shown in the next column.



Top view of spectrum analyzer chassis showing the rf assembly (along rear of chassis, bottom), voltage-controlled oscillator (center), and sweep shaper (top right).

300 μV	$\approx$ 1 division
1,000 μV	$\approx$ 2 divisions
3,000 μV	$\approx$ 3 divisions
10,000 μV	$\approx$ 4 divisions
30,000 μV	$\approx$ 5 divisions
100,000 μV	$\approx$ 6 divisions

5. Set R504 so that 100,000 μV equals six divisions.

Log amplifier tracking is described under final alignment.

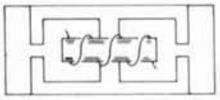


fig. 13. Foil cutout for the 200- and 50-MHz bandpass-filter covers used in the rf section.

I-f section. When checking dc voltages, verify voltages from S601 to the i-f section in all bandwidth positions.

1. Set bandwidth to 250 kHz.

2. Turn off i-f attenuator and turn i-f gain to maximum.

Because of noise or interfering signals, the baseline during some tests of the i-f and rf sections will shift up one or two divisions. If checkout shows that 100 μV should provide one division of deflection and the baseline is already up 1.8 divisions due to noise or external signals, this 100 μV should move the baseline up one to 2.8 divisions. Fifteen microvolts to the base of Q408 and Q404 should provide one division of deflection.

3. Connect generator to Q401 base.

4. Peak L401. Two microvolts should provide one division of deflection.

The crystal filter alignment is described in the section on final alignment.

Rf section. Using a grid-dip meter verify that the 39- and 150-MHz oscillators are oscillating and are crystal controlled. If the oscillators are free running, some adjustment of the two capacitors on the collector side of the crystal can be made. Turn power off and on several times to ensure that oscillators continue to run.

Perform the following steps:

1. Set bandwidth to 250 kHz and video filter to off.

2. Connect generator to input of MX303 (at lead that goes to L315).

3. Tune generator to a peak at 50 MHz and adjust L311 for maximum signal consistent with minimum coupling to L310. Ten microvolts should provide one division of deflection.

The 50-MHz bandpass filter is tuned next. The windings are critically coupled and require careful step-by-step tuning.

1. Connect generator to Q304 base and set display for about four divisions.

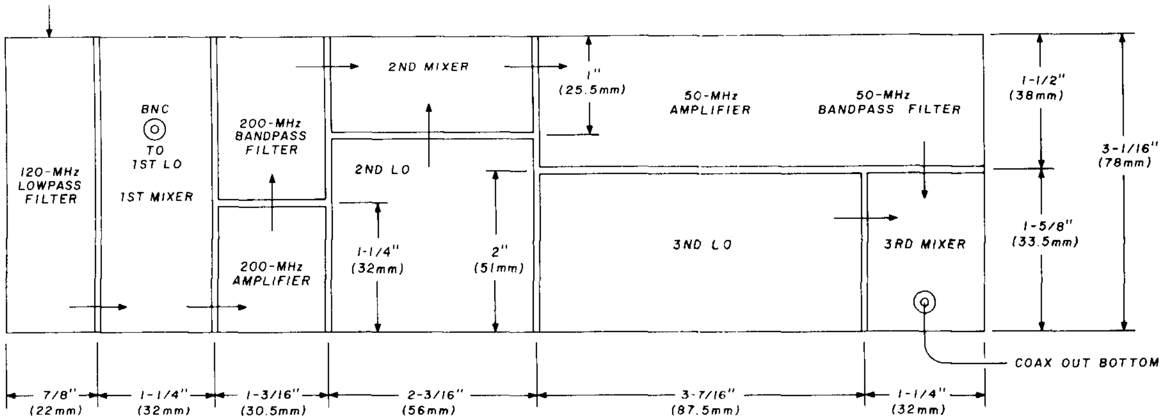


fig. 14. Rf-section dimensions.

2. Short L314 and adjust C306 and C308 for maximum deflection.

3. Short L313 and adjust C305 and C307 for maximum deflection.

4. Make slight adjustment of C305 through C308. Some interaction exists and several attempts will be required, but only a small adjustment will be necessary. Later, when the cover is installed, a slight readjustment will be required. Fifteen microvolts here should provide one division of deflection.

5. Connect generator to MX202 input and tune to a peak around 200 MHz. (At this frequency, signal generators sometimes drift and some readjustment may be necessary).

6. Adjust C304 for maximum deflection.

7. Set generator for exactly three divisions.

8. Install cover on 50-MHz bandpass filter and make slight adjustments of C304-C308. Twenty microvolts should produce one division of deflection.

As with the 50-MHz bandpass filter, the 200-MHz bandpass filter is critically coupled and needs careful adjustment. Proceed as follows:

1. Set all three piston trimmers to the center of their range.

2. Connect generator to Q301 gate.

3. Short L306 and peak C301 and C303.

4. Peak C302.

5. Make slight adjustments of C301-C303 for maximum deflection. Twenty microvolts here should provide one division of deflection.

6. Install covers. (Fine adjustment of the 200-MHz bandpass filter is covered under the section of final alignment).

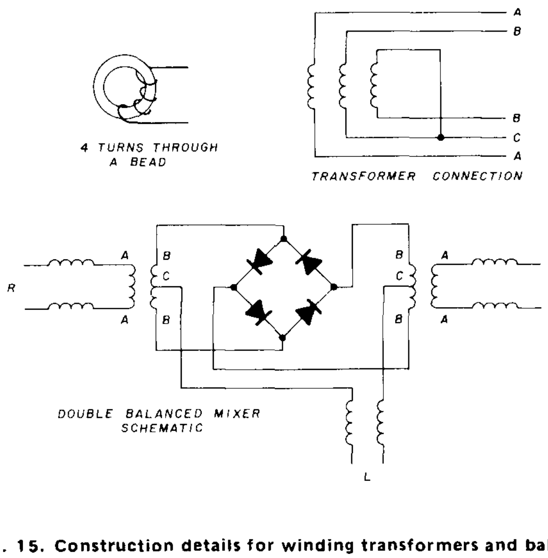


fig. 15. Construction details for winding transformers and baluns used in the r-f section.

final alignment

This section provides final alignment information on the sweep shaper and vco, crystal filter, gain equalizer, tuned circuits, and the log amplifier. Since the sweep shaper and vco may not be familiar, some background information is given on how these circuits work together.

Sweep shaper and vco. The first step is to provide a sweep signal between 0.1-100 MHz. Two conditions must be met to accomplish this: The sawtooth amplitude must be adjusted so that the varactor tuning diode can tune the vco from 200 to 300 MHz. Also, the dc level must be set so that the voltage *one-half way up the sawtooth* tunes the vco to 250 MHz.

The bottom of the varactor diode, CR203, is returned to about −9 volts. The sawtooth amplitude required to tune the vco from 200 to 300 MHz is about 8 volts p-p. The voltage level required to tune the vco to 250 MHz is about −2 volts. Therefore, the sawtooth should start at −6 volts, the half-way point should be at −2 volts, and the peak should end at +2 volts. R101 sets the dc level and R120 sets the sawtooth amplitude (see **fig. 7**).

When the vco was first built, a relatively nonlinear tuning diode (1N5140) was used. It required frequency correction at most frequencies between 240 and 300 MHz to provide linear change in frequency with linear change in dc voltage. However, the MV3102 required

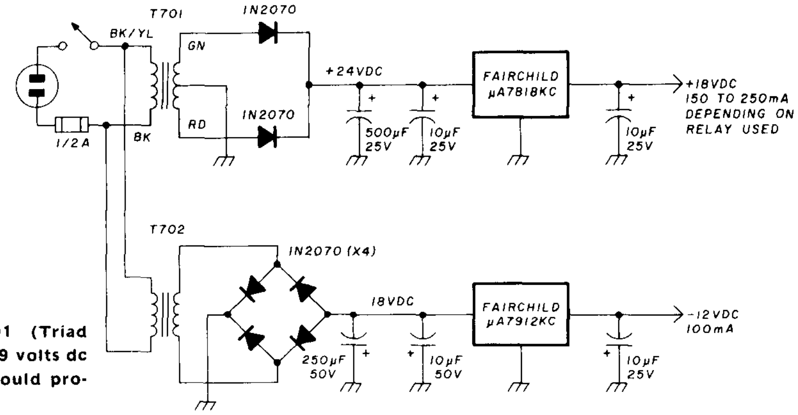


fig. 17. Power-supply schematic. T701 (Triad F90X or equivalent) should provide 23-29 volts dc at approximately 200 mA, and T702 should provide 18-24 volts dc at 100 mA.

correction only at 290 and 300 MHz, therefore R108 through R115 (**fig. 7**) were removed. Several MV3102s were tried with the same results; however, the entire correction circuit is included in the vco schematic.

The settings of R116 and R117 determine at what voltage, or frequency, the correction is made. If R107 is shorted, R117 can be adjusted by setting it so that a 90-MHz signal on the display is moved to the left. R106 through R115 determine the amount of correction, and R116 and R117 determine the position of correction.

A signal source that simultaneously puts out 10-100 MHz in 10-MHz steps makes vco alignment easier. The comb generator in the test equipment list fulfills this requirement. A comb generator can be made by feeding 100 MHz into the prescalar described in reference 1. The ECL logic has fast switching times and generates many harmonics, which provide the comb signal. Possibly other ECL prescalers would provide the comb signal. The frequencies generated will not be of equal amplitude.

In all scan widths other than 10 MHz/division, the center frequency is controlled by R601. R102 and R103 adjust the voltage to R601, so the oscillator is at 200 MHz when the dial reads 0 MHz and at 300 MHz when the dial reads 100 MHz. Errors of at least ±3 MHz are normal for the dial calibration.

Super heterodyne spectrum analyzers produce a signal when the local oscillator is at the i-f or 0 MHz. At 0 MHz, the local oscillator is at 200 MHz, the i-f, and energy from the local oscillator leaks through the first mixer, producing a 0-MHz marker. This is both normal and handy for calibration.

Sweep shaper and vco alignment. Set controls and test equipment as follows:

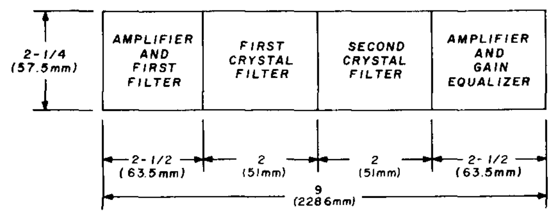


fig. 16. I-f section dimensions.

Bandwidth: 250 kHz.

Scan: 10 MHz/division.

Generator frequency: 50 MHz.

Generator output: ≈ 5 divisions.

Video filter: off.

1. Turn R106 through R115 to maximum resistance.

2. Adjust R101 so the 50 MHz signal appears at the center, fifth division, of the display.

3. Adjust R120 so the 0-MHz marker is at the first division. Some interaction occurs between R101 and R120. Readjust them so that the 0-MHz marker is at the first division and the 50 MHz signal is at the fifth division.

4. Look at the sawtooth with a scope and determine the dc value of the sawtooth at 0 MHz. Set R116 so this dc voltage is present on R115.

5. Look again at the sawtooth at 100 MHz and similarly set R117. Because of interaction, R116 and R117 must be adjusted several times. Final adjustment of R116 and R117 is as follows.

6. Connect comb generator.

7. Turn R108 and check that the 80-, 90-, and 100-MHz signals move to the left. Readjust R117 so only these signals move to the left.

8. Turn R108 to maximum resistance and adjust R114. All signals 20 to 100 MHz should move to the left. Adjust R116 so this condition can be met.

9. Alternately adjust R116 and R117 so that amplitude corrections are made at the proper frequencies.

10. Turn R106 through R115 to maximum resistance.

11. Determine the lowest frequency needing correction and adjust the appropriately labeled pot. Always start adjustment at the lowest frequency because all frequencies above are affected.

Fig. 18 shows final alignment. The camera lens did not have a wide enough angle to display 100 MHz. Note that 10 through 40 MHz are slightly high and 60 through

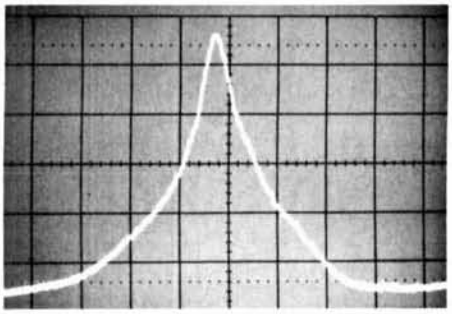


fig. 18. 10-kHz bandwidth position after alignment. Horizontal scale: 10 kHz per division.

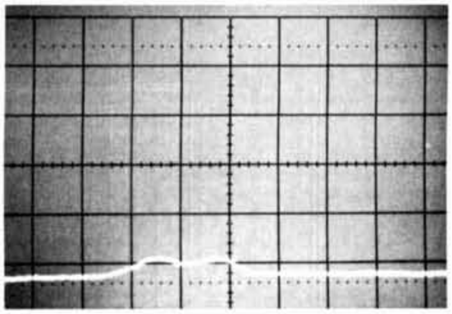


fig. 19. Setup as shown in fig. 24. Turn on video filter and note reduction of noise.

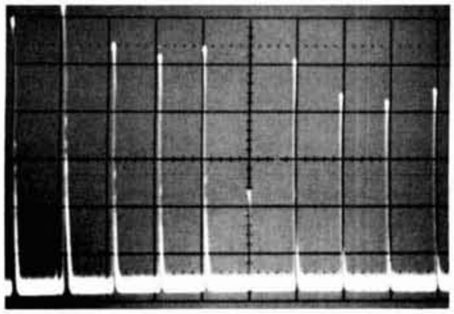


fig. 20. Output from comb generator, 10-100 MHz. Scope camera did not have enough width to display 100 MHz. Maximum errors are about 2 MHz.

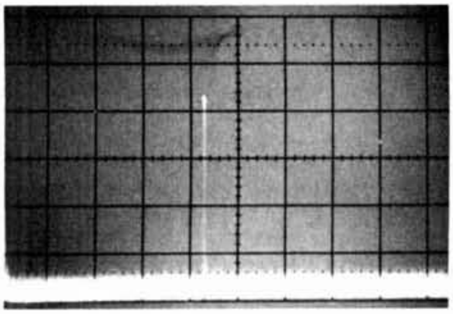


fig. 21. Setup is as shown in fig. 27. I-f gain has been reduced to take advantage of the full dynamic range.

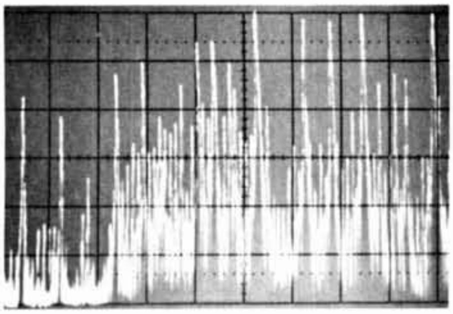


fig. 22. After connecting antenna to inputs, turn BW to 1 kHz and note how stations in the broadcast band can be resolved.

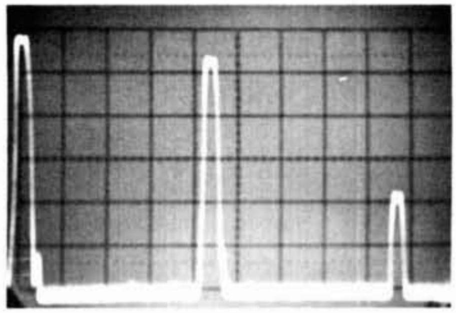


fig. 23. Second harmonic of transmitter operating at about 4 MHz. Second harmonic is down only about 32 dB.

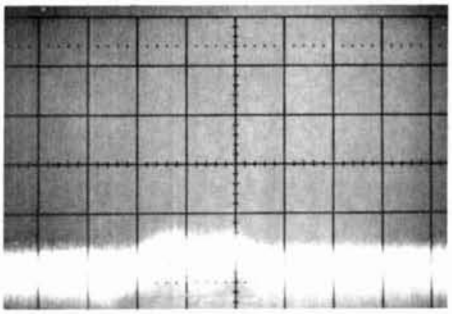


fig. 24. BW 250 kHz, Scan 100 kHz/division and generator frequency about 40 MHz. Increase generator output to see a signal in the noise with the video filter off.

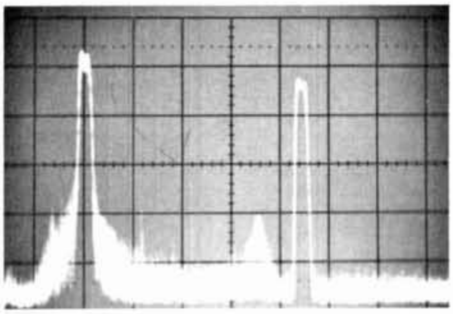


fig. 25. Setup as shown in fig. 27. Set scan to 1 MHz/division. Note video carrier, color information about 3.6 MHz above video carrier, and sound carrier 4.5 MHz above video carrier.

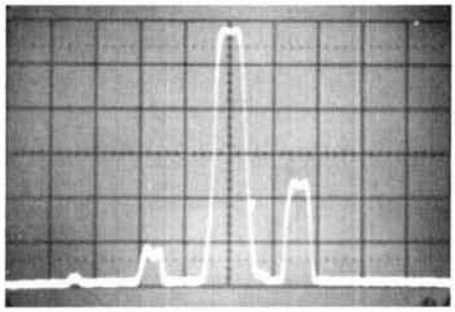


fig. 26. Output of a commercially made ssb transmitter into a 50-ohm load with carrier inserted at about 3.8 MHz. Scan 500 kHz/division. Note spurious signals at about 3050 kHz and 4550 kHz.

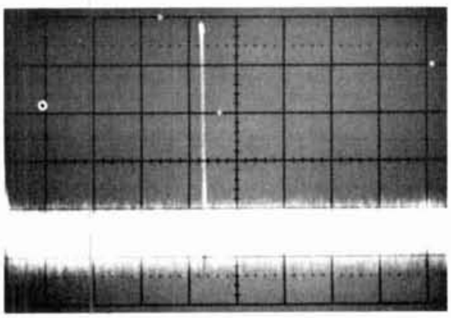


fig. 27. I-f gain adjustment. BW 250 kHz, Scan 10 MHz/division, Generator frequency 40 MHz; generator output about 4 divisions. The i-f gain is set too high, which shifts the baseline up two divisions. This results in a loss of about 20 dB of dynamic range.

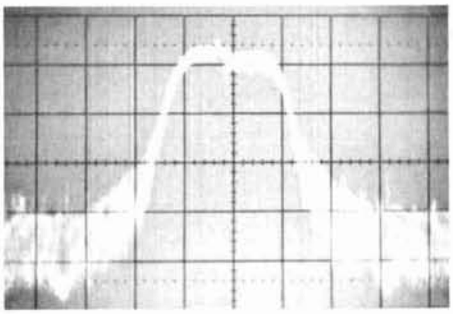


fig. 28. TV station. BW 250 kHz, Scan 100 kHz/division. Adjust frequency control to the video carrier of any TV station between channel 2 and channel 6. Connect 50- or 75-ohm TV antenna to input. Line lock the scope and note video sidebands and vertical sync pulse, which will drift through the top of the signal.

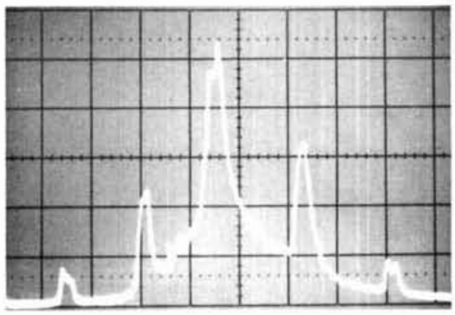


fig. 29. Nonlinear modulation. BW 1 kHz, Scan 10 kHz/division, generator frequency about 40 MHz, and about 50% modulation with a 15 kHz signal. The generator used here was not designed to be modulated at a 15-kHz rate. Note unsymmetrical sidebands and harmonics of the modulated frequency. The harmonics were generated in the modulation process.

90 MHz are slightly low. Maximum error shown here is about 2 MHz.

Crystal-filter alignment. Set controls and test equipment as follows:

Bandwidth: 1 kHz.

Scan: 50 kHz/division

Generator frequency: 20 MHz (not critical).

Generator output: 5½ divisions on display.

Sweep speed: 20 ms/divisions.

Video filter: 10 kHz.

Video gain: maximum.

The crystal filters are aligned one at a time. The filter not being aligned should have a 0.01-μF disc capacitor soldered across the crystal. Always maintain about 5½ divisions on the display but never more than six divisions.

1. Adjust C402 for a narrow peak at the crystal-filter frequency.

2. Go between 1 kHz and 10 kHz bandwidth, adjusting C401 for narrowest bandwidth without regard to amplitude. Do not adjust C402 in the 10-kHz position.

3. Adjust the second crystal filter as you did the first one.

4. Increase input level to 10,000 μV.

5. Turn i-f gain down to 5½ divisions and remove both .01-μF capacitors.

6. Alternately adjust C401 and C403 for minimum width and symmetrical skirts at the base of the signal in both the 1-kHz and 10-kHz positions. Possibly there will be some spurious resonances slightly above the filter frequency after alignment in the 10 kHz position.

Gain compensation. Set controls and test equipment as follows:

Bandwidth: 250 kHz.

Scan: 50 kHz/division.

Generator frequency ≈ 20 MHz.

Generator output: 5 divisions on display.

Sweep speed: 50 ms/division.

1. Go to 10 kHz bandwidth and adjust R403 for 5 divisions.

2. Go to 1 kHz bandwidth and adjust R402 for 5 divisions.

3. Drill access hole for L401 in i-f amplifier cover and install cover.

Tuned-circuit alignment. All tuned circuits are peaked during this procedure. Even though most have been adjusted, this is a good check to see if all are peaked at full operation. Again, only slight adjustments of the 50- and 200-MHz bandpass filters should be needed. If major adjustment is required, or if one coil won't tune, go back to the procedure in the rf section that describes initial alignment. Set controls and test equipment as follows:

Bandwidth: 250 kHz.

Scan: ≈ 500 kHz/division

Generator frequency: ≈ 20 MHz.

Generator output: ≈ 4 divisions on display.

1. Peak C301, C302, and C303 for maximum signal.

2. Peak C304 through C308 for maximum signal consistent with a flat top across signal.

3. Peak L401 for maximum signal consistent with a flat top across signal. There may be about a 1-dB dip in the center of the signal.

Log amplifier. Set controls and test equipment as follows:

Bandwidth: 10 kHz.

Scan width: 10 kHz/division.

Generator freuqency: ≈ 20 MHz.

Sweep speed: 10 ms/division.

Video filter: 10 kHz.

1. Set the generator output to 30 μV.

2. Set i-f gain for one division. Note that each pot adjusts the level for two divisions and the best compromise should be achieved. The tolerance is ±0.2 division. Always start from one division when performing this alignment.

divisions	output from generator (microvolts)	alignment pot
1	30	R503
2	100	R503
3	300	R502
4	1000	R502
5	3000	R501
6	10,000	R501

operation

This section provides an explanation of the function of the controls and concludes with some experiments to demonstrate operation.

Rf attenuator. The purpose of this circuit is to reduce the amplitude of the incoming signal to a convenient level for display on the spectrum analyzer. The maximum level to the attenuator should be no greater than 1 volt rms, and the maximum level to the first mixer should be no greater than 10 mV rms. One of the most common errors in the operation of a spectrum analyzer is to use too much i-f attenuation and feed an excessively high level to its input. This results in overloading the rf section, which generates spurious signals and causes possible damage to the first mixer. Never overload the front end.

Frequency control and fine tuning. The frequency control tunes the spectrum analyzer to the desired operating frequency. In the 10 MHz/division position, the center frequency is set to 50 MHz and the frequency control is inoperative. The fine tuning control has a range of about 500 kHz.

Scan width. After setting the center frequency with the frequency control, the scan width determines the frequency width that will be displayed. For example, if the frequency control is set to 20 MHz and the scan width to 100 kHz/division, the spectrum analyzer will sweep from 19.5 to 20.5 MHz.

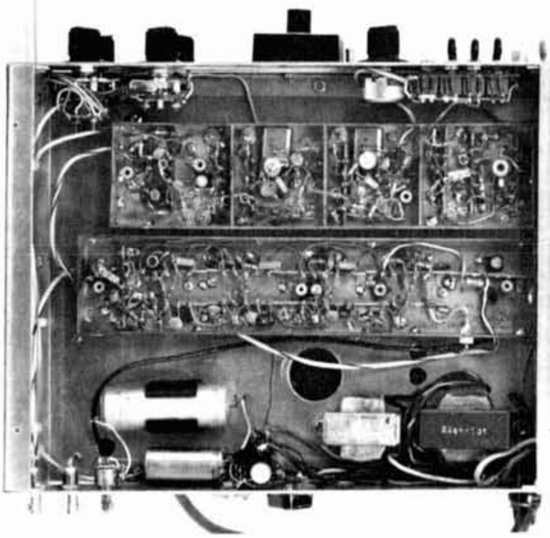
Bandwidth. As in a conventional radio receiver, this control determines the selectivity of the analyzer. Narrow bandwidth is required to separate signals relatively close together and wide bandwidth is required for high sweep speeds.

I-f attenuator and i-f gain. The i-f attenuator is used for limited tests where discrete steps of attenuation are required. The i-f gain is used to provide sufficient gain with minimum noise at the baseline. The noise at the baseline rises as the bandwidth is increased and should be maintained at between one-half and one division to avoid degrading the dynamic range. See **figs. 22** and **23**.

Video filter. The video filter is used to remove high-frequency noise from the signal when looking at small scan widths with slow sweep speeds. See **table 2** for video filter operation. **Figs. 24** and **25** are examples of its effect in the 250-kHz bandwidth position.

Resolution. If the sweep speed is too high, the scan (or sweep) width too great, or the bandwidth too narrow, the display will lose amplitude or smear. **Figs. 26** and **27** show displays of a broadcast signal at the 250-kHz bandwidth setting. Further examples are shown in **figs. 28** and **29**, which are displays of a local television station.

A quick check on resolution is to decrease sweep speed and look for narrowing or increased amplitude of the displayed signal. The sweep speed should never exceed 2 ms/division. A P2 and P7 scope tube would be advantageous but is not absolutely required. **Table 2** gives maximum sweep speeds for some scan widths, bandwidths, and settings of the video filter.



Bottom view of spectrum analyzer chassis showing the i-f assembly (along front of chassis, top) and logarithmic amplifier (center). Power supply components are located along the rear of the chassis, bottom.

table 2. Maximum sweep speeds for resolution at spectrum-analyzer bandwidth settings.

bandwidth (kHz)	scan width (MHz/division)	maximum sweep width (per division)	video filter (kHz)
250	10	10 ms	Off
250	1	5 ms	10
250	0.1	2 ms	10
250	0.01	not usable	—
10	10	not usable	—
10	1	0.2 sec	10
10	0.1	50 ms	10
10	0.01	20 ms	1
1	10	not usable	—
1	1	not usable	—
1	0.1	0.2 sec	10
1	0.01	50 ms	1

400-500 MHz operation. If the input lowpass filter is disconnected, the 200- to 300-MHz first local oscillator will mix with 500 to 400 MHz, producing the required 200 MHz first i-f. This was tried and some strong local low-frequency signals leaked through. Sensitivity seemed similar to normal operation, but signals were displayed in reverse.

Use as a radio receiver. The output can be connected to an audio amplifier with 1-megohm input impedance and used to monitor radio signals. However, because it is a spectrum analyzer, the instrument has several limitations when used as a radio receiver. Dial calibration is accurate to only ±3 MHz. Because of its high frequency, the local oscillator drifts and some fine tuning might be required. There is no tuned circuit at the front end, therefore a strong signal can cross-modulate the analyzer. Despite these shortcomings, WWV, BBC, Australian Broadcasting, Radio Netherlands, Radio Moscow and many other stations have been received using a TV feedline as an antenna.

As with most anything else, there is more than one way to design a spectrum analyzer. If you find an easier or better way to improve this spectrum analyzer, without degrading its performance, or find any obvious errors, I'd appreciate hearing from you.

reference

1. Wayne C. Ryder, W6URH, "500-MHz Decode Prescaler," *ham radio*, June, 1975, page 32.

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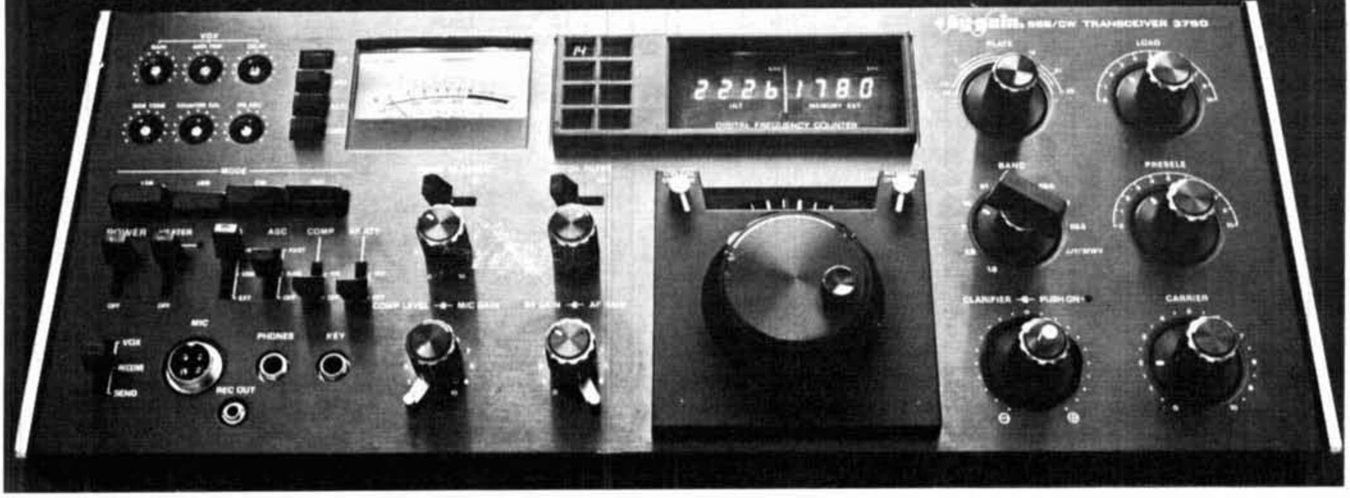
2. *Spectrum Analyzer Measurements*, (Tektronix part no. 062-1070-00), Tektronix, Inc., P.O. Box 500, Beaverton, Oregon 97077.
3. *Spectrum Analyzer Circuits*, (Tektronix part no. 062-1055-00), Tektronix, Inc., P.O. Box 500, Beaverton, Oregon 97077.

ham radio

HOW TO ENJOY GLOBAL COMMUNICATIONS WITHOUT A PHONE BILL



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It covers all amateur band 1.8-30MHz (160-10 meters) with uncanny accuracy. It has advanced Phase-Lock-Loop circuitry. Dual gate MOS FET's at all critical RF amplifier and mixer stages. Electronic frequency counter with digital readout. Digital memory display. And stability that's truly remarkable by any standards.

The Hy-Gain is simply the finest amateur transceiver you can buy. There is nothing else like it.

Read our specs. Then see your amateur radio dealer or write for complete details on the features and performance of the incomparable 3750.

SPECIFICATIONS

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

AUDIO FREQUENCY CHARACTERISTICS
300 - 2700Hz (-6dB)

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

MODES OF OPERATION
SSB (LSB or USB)
CW

INPUT POWER
200 W PEP

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 (including 23 FET)	98
IC	43
Diode	120
Tube	3
Digital Ind.	1

WEIGHT
44 lbs. 6 ozs. (23kg)

DIMENSIONS
16¾" x 7" x 13⅝"
(420 x 172 x 340mm)

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WE KEEP PEOPLE TALKING.

RTTY tape editor

This tape editor is designed for use with a transmitter-distributor (or TD as it is more commonly known), a keyboard send-receive teletype (KSR) and a reperforator. Any serious RTTY operator who owns this equipment has surely longed for an easy way to delete or insert material into a tape. This need could arise from errors in reception when punching tape off the air, or it might be merely a desire to modify a tape to suit your particular needs. In any case the arrangement described here should fit the bill.

tape editing

The usual method of tape editing is to insert the original tape into the TD and, while operating on local loop, make a new tape from the original tape on the reperforator. Editing is done by stopping the tape in the TD, entering corrections from the keyboard, moving the tape manually through the TD past the incorrect material, then restarting the TD to continue until reaching the next point for correction. The result of this exercise is monitored on the page printer. The problem with this procedure is getting the original tape moved to just the exact spot from which it should be restarted. This may sound simple enough, but if you've ever tried this method you know how easy it is to end a character or so off from where you intended to restart, with another error to edit.

examples

With the tape-edit circuit described here, the TD is stepped one character for every character entered from the keyboard. Thus, if the tape reads "shop," and you wish to change it to "ship," you can stop the TD after "sh," put the *EDIT* switch to *ON*, and type "i." The TD will advance one character when the "i" is entered so that when the *EDIT* switch is thrown to *OFF*, the TD will start again and print the letter "p" thus completing the correction. Alternatively, the tape could have been stopped just before the word "shop," the *EDIT* switch placed to *ON* and the entire word "ship" typed in from the keyboard. The TD would have been stepped along one character for each character typed so that the net result would be the same.

Suppose the word "ship" had been garbled in the original tape, but from the message context you were able to determine that "ship" was the correct word. Suppose it read "shxmp." Stop the tape as before and turn the *EDIT* switch to *ON*. Enter "i" from the keyboard then strike the *LETTERS* key, which will move the TD one more character past the extraneous "m." Turn the *EDIT* switch to *OFF*, which will restart the TD with the letter "p," and the correction will be complete. With these corrections, editing can be done on the air after a very little practice.

The circuit is shown in **fig. 1**. The objective is to advance the TD one character for each character entered from the keyboard without having the TD affect the print (or the new tape being punched by the reperforator).

A miniature dpdt toggle switch was mounted on the front of the TD just below the main *ON/OFF* switch and slightly to the left. This required drilling a ¼-inch (6.5mm) hole in the TD casting — an easy job. One section of this switch, S1A, was placed in parallel with the TD signal line so that a *MARK* condition exists at all times when the *EDIT* switch is in the *ON* position. The other section, S1B, is connected in series with the *TAPE-OUT* and the *TAUT-TAPE* switch and is closed when the *EDIT* switch is *OFF*. A pair of leads parallel S1B and run to the contacts on a polar relay. When the *EDIT* switch is *ON*, S1B is open so the only way that current can get to the *RUN MAGNET* on the TD is through the polar-relay contacts.

The polar relay coil is connected in series with the keyboard. Thus loop current flows through the coil until a key is depressed. This action holds the relay in the *MARK* position.

When a key is depressed, the *START* pulse moves the polar relay to the *SPACE* position, which closes the circuit to the TD run magnet, allowing the TD to start. As soon as the *STOP* pulse is transmitted for whatever character was typed in, the polar relay will return to the *MARK* position, which opens the run-magnet circuit, stopping the TD after it has advanced one character. Thus the TD advances the tape one character for each character sent from the keyboard. Since the first section of the *EDIT* switch has shorted the TD signal line, the TD has no effect on the transmission.

tape changes

Earlier in the article I discussed error corrections. Sometimes it's desired to make changes in a tape. For example I have a Christmas-greeting tape, which was received from the State Council of Civil Defense. I wanted to use this tape with my name in the box where the original read "State Council of Civil Defense." **Fig. 2** shows the print before and after editing.

cautionary measures

Remember that nonprinting functions must be taken into account when editing. For example, if W3EAG is to be replaced by "Tommy," this is not just a one-for-one substitution as it appears. The figure function preceding the 3, and the letters function following it, must be accounted for since these are two additional characters

By Tom Gibson, W3EAG, R. D. 1 Evansburg Road, Collegeville, Pennsylvania 19426

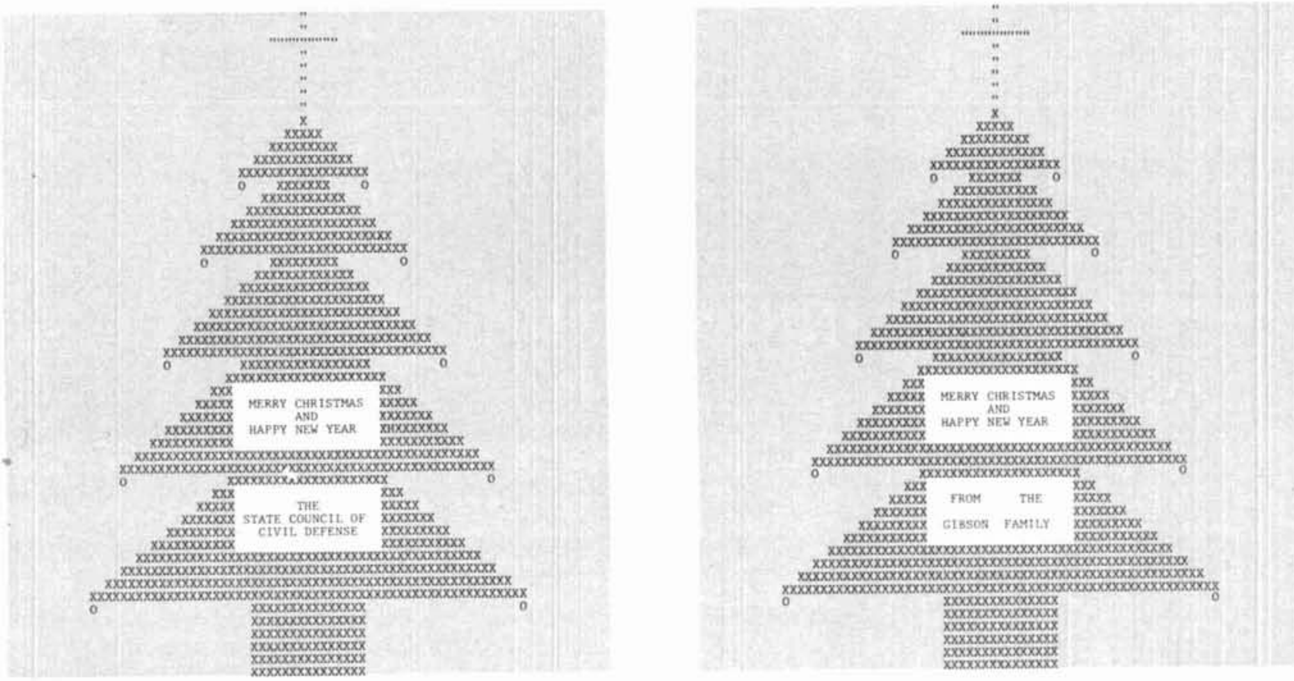


fig. 2. A Christmas message before and after editing.

on the original tape that are not required for this particular substitution. This can be accommodated by striking the *LETTERS* key twice, which will advance the tape but not the printer. However, if W3EAG were being replaced with K2XXX this would be a one-for-one substitution and no *LETTERS* need be struck.

Always stop the TD with the *TAUT-TAPE* switch, then turn the *EDIT* switch to *ON*. If you attempt to stop the TD by turning the *EDIT* switch to *ON*, you may garble a character since the edit switch shorts the TD signal line.

The polar relay used in my rig switches ac to the run magnet. Some operators may be using dc, in which case some attention to spark suppression at the relay contacts would be advisable.

ham radio

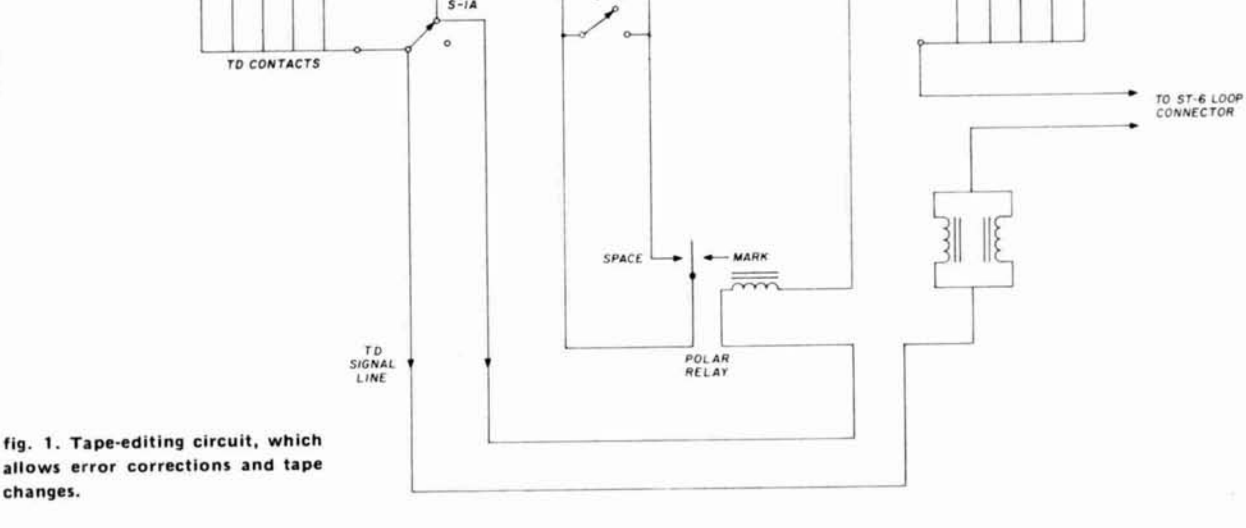


fig. 1. Tape-editing circuit, which allows error corrections and tape changes.

top-coupled bandpass filter

a chebyshev design

Practical design and construction data for 3-, 4- and 5-section Chebyshev bandpass filters for amateur applications

The top-coupled bandpass filter is the easiest of its type to construct. It appears to be several parallel resonant circuits connected in sequence with small coupling capacitors. All inductors are the same value. In actuality, this type of bandpass filter is one of many configurations based on a lowpass filter prototype. Presented here are the necessary component formulas to enable design and construction of 3-, 4-, and 5-section "1-dB ripple" Chebyshev bandpass filters along with passband and skirt selectivity response, input impedance, and insertion loss.

why Chebyshev?

The most common filter design is the Butterworth with its flat passband response. A Chebyshev filter has ripple in the passband, defined by dB difference of peak-to-valley; design bandwidth refers to the dB-down points on the edges rather than the 3 dB points of the Butterworth. The Chebyshev response also has better skirt attenuation than the Butterworth.

The names refer to the mathematical operations used in determining component values. The math involved is lengthy and may be found in many textbooks. Here, the only math will be that required for one type of Chebyshev ripple design.

Both Chebyshev and Butterworth designs for equal load and source impedances suffer shape distortion and increased insertion loss as a function of finite Q. This will be seen on the response curves.

Schematics and individual value formulas are shown in **fig. 1**. The formulas are derived from lossless-element designs and feature equal end resistances. Following are definitions common to all:

f_L = *Lowest frequency of passband, MHz*
f_H = *Highest frequency of passband, MHz*
f_o = *Geometric center frequency* = $\sqrt{f_L f_H}$
f_b = *Bandwidth* = $f_H - f_L$
p = *Fractional bandwidth* = f_b / f_o
C_o = *Resonating capacitance* = $\dfrac{25330.3}{f_o^2 L}$

with L equal to $L3$, $L4$, or $L5$, depending on filter

Note that all terms are scaled to megahertz, microhenries, picofarads, and kilohms. Practical designs require high end resistances. Matching to lower values is given later.

how degrading can you get?

Circuit Q has little effect on skirt attenuation but does distort the passband. Skirt response is shown in **figs. 2** and **3** while passband response is shown in **figs. 4, 5**, and **6**.

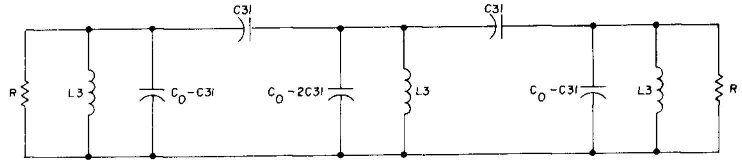
The parallel L-C combined Q will determine the passband shape and insertion loss. It can be shown that

$$Q = \textit{combined } Q_C \textit{ and } Q_L = \frac{Q_L Q_C}{Q_L + Q_C}$$

A reference to Q will generally refer to the combination. Very high Q will still result in distortion, as shown on the passband curves. The curves are *not* ideal – they have been deliberately chosen to represent degraded performance in choosing coils and coupling capacitors as will be explained in the design example.

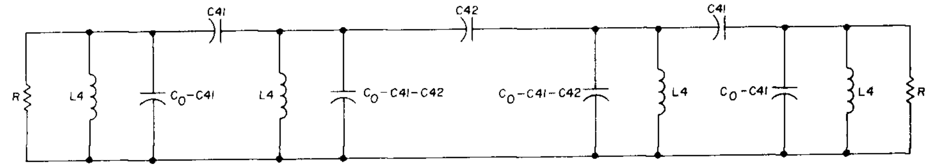
This unorthodox presentation is the "worst-case" design situation. Selection of component values closer to calculated values will improve shape, insertion-loss, and skirt attenuation. Curves of **figs. 2** through **6** were taken from the design example where the inductor was higher in value than given by the formula, the coupling

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



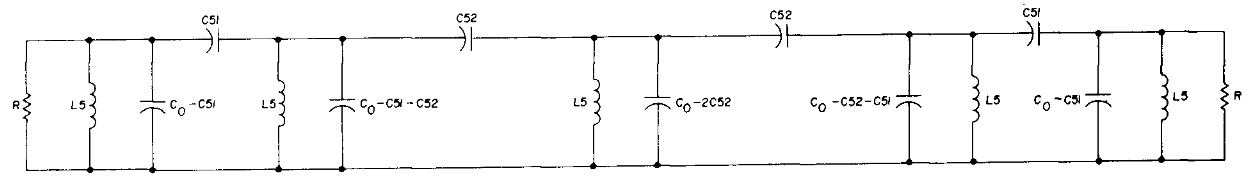
$$L3 = \frac{786\ p\ R}{F_0} \qquad C31 = \frac{227}{F_0\ R}$$

3 SECTION



$$L4 = \frac{765\ p\ R}{F_0} \qquad C41 = \frac{225}{F_0\ R} \qquad C42 = \frac{206}{F_0\ R}$$

4 SECTION



$$L5 = \frac{745\ p\ R}{F_0} \qquad C51 = \frac{223}{F_0\ R} \qquad C52 = \frac{188}{F_0\ R}$$

5 SECTION

fig. 1. Basic top-coupled bandpass filters of Chebyshev design. All values are in MHz, microhenries, pF and kilohms. The value p is the fractional bandwidth, as discussed in the text.

general philosophy of design

The intention of the design data presented here is to enable the nonspecialist to obtain working values as quickly as possible in a wide area of applications. The "symmetric" design of equal source and load resistance meets this criterion even though the component calculations are for lossless elements. The trade-off of increased insertion loss and passband distortion was considered valid for a non-commercial application when compared with ease of calculation and impedance matching.

The filter specialist may prefer some more "meat," but this requires reference to heartier texts and references. It would have been nice to present a table of filter elements – this is possible by using Craig's excellent *Design of Lossy Filters* handbook (see bibliography). It was considered for this article, then shelved for several reasons. First, there is the choice of bands; which frequencies to choose besides the obvious 160- to 6-meter bands, such as for intermediate frequencies and the like. Secondly, what would be the expected impedance levels? The real world is not always 50 ohms. Last, such designs require a specific unloaded Q of resonators which can be achieved by resistor loading of higher-Q components; the passband takes on strange shapes when this is not done.

An optimum shape and loss design takes at least twice the effort to obtain working hardware, even with a table of values to start with. The prime consideration was the amount of time available to the Amateur – calculations can be done during your spare time, lunch breaks, etc., but working with hardware requires workshop time.

capacitor lower. This is the general rule in component selection. The Butterworth 3-section passband of **fig. 7** is based on *ideal* values and is useful for comparison.

Decreasing Q will narrow the bandwidth. This does not mean that a low Q is unusable – simply that the design bandwidth must be changed if it does not fit a particular situation. This can be used to advantage since *required* unloaded Q is inversely proportional to bandwidth.

scaling

The horizontal scale of **figs. 2** through **7** may be read in two ways: Units of bandwidth plus or minus center frequency as a universal value, *or* fractional frequency as a function of center frequency for a 5 per cent bandwidth. The curves may be used with bandwidths from 3 to 8 per cent of center frequency with reasonable accuracy.

The inverse relationship of Q versus bandwidth can be seen by examining required inductance value for a given load resistance. Inductance must increase with bandwidth because of the p term in the numerator of the formula. Since the loss from a finite Q appears as a resistance in parallel with the shunt L and C at a value of Q times the reactance of either at center frequency,

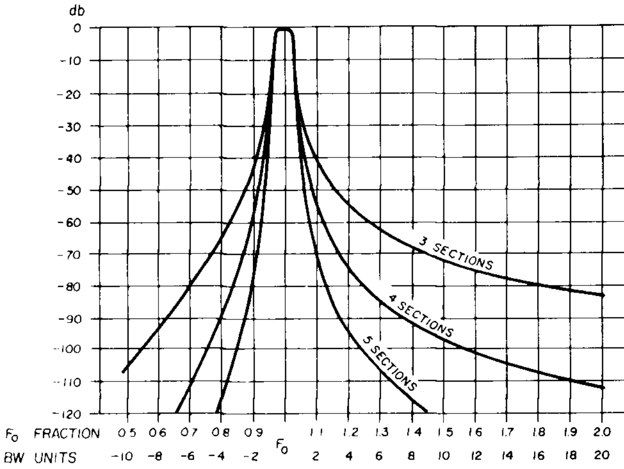


fig. 2. Skirt response of 1-dB ripple Chebyshev filters with 5% bandwidth, worst case example.

increasing the reactance will also increase the equivalent resistance. It is the ratio of equivalent resistance to the source and load resistances that determines insertion loss and shape change.

If the available Q is 160 and the percentage bandwidth only 2.5, the $Q=80$ curve of the 5 per cent bandwidth example will be the response. Similarly, if the bandwidth is to be 10 per cent, the $Q=320$ curve would be used. Other Qs and bandwidths may be interpolated.

Variations in Q will have very little affect on skirt response below −30 dB; skirt response beyond this can be taken directly from the curves since the reactances and end resistances control here.

example

A bandwidth of 0.5 MHz with a center frequency of 10 MHz is desired. Source and load resistance will be 5 kilohms. For the 3-section filter,

$$L3 = 78.6 \times 0.05 \times 5/10 = 1965 \mu H$$
$$C31 = 227/(10 \times 5) = 4.54\ pF$$

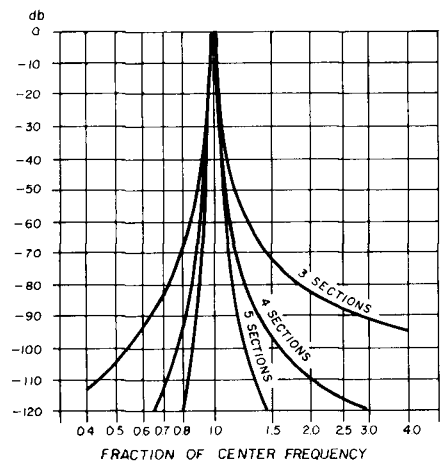


fig. 3. Skirt response of 1-dB ripple Chebyshev filters with 5% bandwidth, worst case example. This is the same as fig. 2 except that frequency is plotted on logarithmic scale.

Resonating capacitance C_o is determined solely by L3, L4, or L5 at the center frequency:

$$C_o = 25330.3/(10 \times 10 \times 1.965) = 128.907\ pF$$

The actual parallel circuit capacitance will be C_o minus the adjacent coupling capacitance values, so the end values will be $128.907 - 4.54 = 124.367\ pF$, and the middle value will be $128.907 - 2(4.54) = 119.827\ pF$.

For the 4-section filter:

$L4 = 1.9125\ \mu H$
$C_o = 123.446\ pF$
$C41 = 4.5\ pF$
$C42 = 4.12\ pF$

End parallel capacitors = $127.946\ pF$
Middle parallel capacitors = $123.826\ pF$

For the 5-section filter:

$L5 = 1.8625\ \mu H$
$C_o = 136.002\ pF$
$C51 = 4.46\ pF$
$C52 = 3.76\ pF$

End parallel capacitors = $131.542\ pF$
Next-to-middle capacitors = $127.782\ pF$
Middle parallel capacitor = $128.482\ pF$

All of these designs look very nice on paper but are not practical because none of the component values are commercially available. To obtain practical values, it was decided to select the nearest higher value of L in the 10% tolerance value of $2.2\ \mu H$. Coupling capacitance was selected as the nearest lower 10% value, $3.9\ pF$ for C31, C41, C51, and $3.3\ pF$ for C42 and C52. C_o now becomes $115.138\ pF$ and the actual circuit parallels chosen by the adjacent coupling capacitor subtraction rule.

These values are extreme, particularly for the inductor, but serve to show "worst-case" variations and are the values calculated for the curves of **figs. 2** through **6**.

As a very general statement, lowering coupling capacitance will improve skirt attenuation with little affect on passband ripple; increasing inductance reduces ripple and skirt attenuation. Lowering coupling capacitance and increasing inductance together has the effect of approaching Butterworth response. A simultaneous change in the opposite direction will increase both ripple and skirt attenuation.

A low-Q situation always results in a rounded passband, whether Butterworth or Chebyshev. Skirt attenuation is governed by reactance only at frequencies well removed from center. This can be seen by the converging curves of **figs. 3, 4, 5,** and **6** at 20 *dB* down.

insertion loss and input impedance

Using lossless elements, the input impedance of all odd-number section filters will be equal to the load resistance at the center frequency. The impedance seen by a current source (transistor collector or tube plate) feeding the input will be half the load resistance for symmetric load and source resistances.

Input impedance will vary over the passband with magnitudes changing as much as 2:1. Magnitude is maximum where ripple in the passband is at its peak. Odd-section filters have a peak in the middle whereas even-sections have a dip. The four-section filter will have a lower input impedance at the middle but will be equal to load resistance at the peaks. Because of this variation, input impedance must be considered to be a *mean* value over the passband.

Insertion loss, as used here, is the ratio of input voltage magnitude (including the source resistance) to output voltage magnitude as compared to the peak of lossless element response. Insertion loss is commonly the input/output ratio alone but the ripple response of Chebyshev designs requires a different definition.

Fig. 8 shows the variation of input impedance and insertion loss for ideal values but with finite values of Q. This may be used as a general guide.

response of the filter

For an interstage application, the driver will have a load equal to source-resistance in parallel with input impedance of the filter. Insertion-loss data assumes use of lossless elements so the driver load will be half the design resistance. For the 5k example, the driver load is 2.5k when computing overall driver and filter gain.

As an example, take the 3-section filter as graphed in **fig. 4** with a Q of 80. Assume the driver stage has a 20 dB gain with a 2.5k load. At the center frequency, filter insertion loss is 5 dB so the overall gain is $20 - 5 = 15$ dB. From **fig. 2**, the skirt attenuation at $0.8f_o$ is $67\ dB$, so the overall gain is $20 - 67 = -47\ dB$. Relative difference between f_o and $0.8f_o$ is $-62\ dB$.

Any driver resistance, such as $1/h_{oe}$ of a transistor, must be part of the load resistance with the actual load adjusted to fit the design resistance. Stray capacitance becomes part of the end section capacitance.

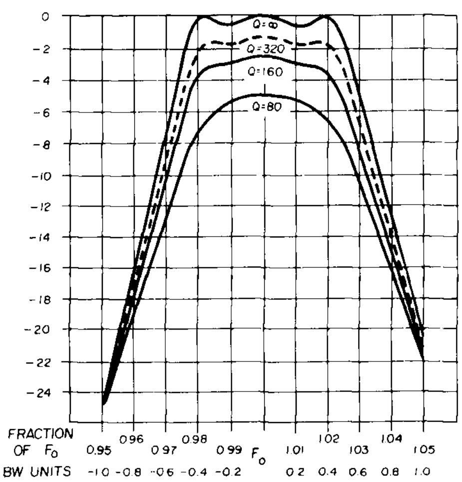


fig. 4. Passband response of 3-section, 1-dB Chebyshev filter vs Q. Worst case example.

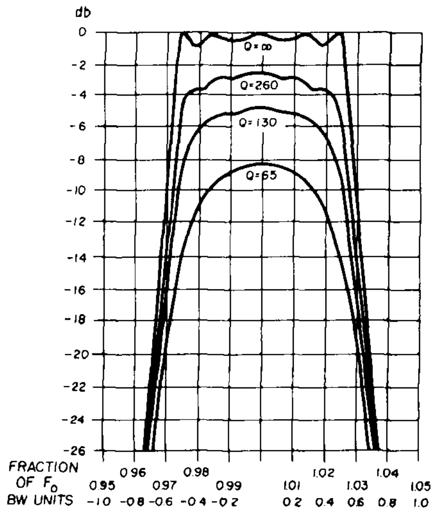


fig. 5. Passband response of 4-section, 1-dB Chebyshev filter vs Q. Worst case example.

Fig. 9 shows three different matching methods. They apply to either end and assume matching to a pure resistance. Filter end resistance varies with section Q and is shown in **fig. 8** as a fractional value of design resistance. The following definitions apply to all methods:

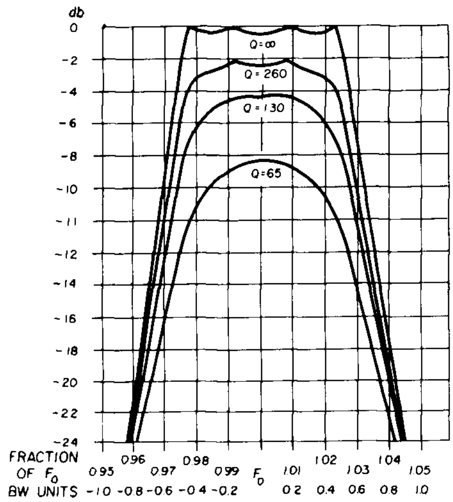


fig. 6. Passband response of 5-section, 1-dB Chebyshev filter vs Q. Worst case example.

R_i = Actual input resistance; obtain by multiplying design R by factor shown in **fig. 8**

R_g = Low resistance to be matched

X_i = Inductor reactance at center frequency less reactance of the adjacent coupling capacitor at the center frequency.

f_o = Center frequency of filter

The inductive-tap method assumes a unity coupling factor and should work if toroids are used with windings spaced evenly over the entire form. Lower permeability toroids such as Amidon types 2 and 6 should have at least 75 percent of the toroid form filled with wire. Note that wide spacing and placing a gap between start and finish will lower the coupling factor as well as Q.

Best results in overall performance of capacitive matching is achieved by making C_s and C_a fixed values with C_p and C_b trimmable. The end inductor must still resonate at f_o. All of the matching methods will work at either end.

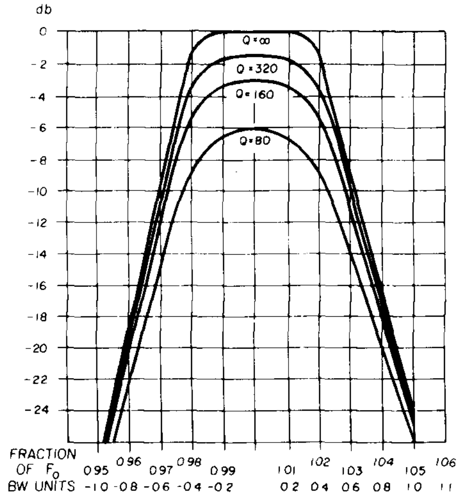


fig. 7. Passband response of 3-section ideal Butterworth filter vs Q.

The impedance looking into the filter will not be uniform with any of the methods. The inductive-tap and C_a/C_b arrangements will present low impedances outside of the passband; the C_s/C_p type will present high impedances outside of the passband. When used with antenna-inputs, a small tee or pi resistance pad should be used to prevent unusual responses due to mismatch.

table 1. 50-ohm attenuation pads using standard 10% resistors. R_{in} assumes 50-ohm load; worst vswr is with output open or shorted.

	TEE-pads			worst
dB	R1	R2	R_{in}	vswr
-3.52	10	120	50.00	2.60
-4.24	12	100	50.27	2.24
-5.29	15	82	51.26	1.94
-6.68	18	56	48.71	1.58
-8.14	22	47	50.44	1.38
-10.47	27	33	50.10	1.20
-13.64	33	22	50.39	1.10
-18.43	39	12	49.57	1.04

	PI-pads			worst
dB	R3	R4	R_{in}	vswr
-1.74	470	10	49.39	5.11
-2.05	470	12	50.99	4.76
-2.58	330	15	49.63	3.48
-3.75	220	22	48.82	2.50
-4.56	180	27	48.36	2.13
-6.19	150	39	50.66	1.67
-9.66	100	68	50.33	1.25
-12.51	82	100	50.44	1.13
-15.85	68	150	49.27	1.07

A table of attenuation pads using 10-percent tolerance resistors is shown in **table 1.**

Voltage gain is achieved when matching is only at the input; voltage loss occurs when matching is on the output. This is true for all three types. Gain or loss is equal to the square-root of R_i/R_g. Filter loss is still present and the curves of **fig. 8** should be used for the overall gain or loss.

To illustrate this, assume the 3-section example already given with a $Q = 160$. From **fig. 8**, the loss of the filter alone is 2.6 dB and $R_i = 3.74k\ (0.718 \times 5k)$. The input R_g is $50\ ohms$ so $R_i/R_g = 7.48$ with a voltage loss of $1/2.735$ or $-8.74\ dB$. The net voltage gain is $18.74 - 2.6 - 8.74 = +7.4\ dB$.

capacitive matching examples

Before beginning, the following reactance formulas apply with all values in ohms, pF, μH, and MHz:

$$X_L = 6.28319 f_o L$$

$$X_C = \frac{159155}{f_o C}$$

Take the 3-section example with $Q = 160$ and match to 50 ohms. R_i is found to be 3.74k from the **fig. 8** curve and R_g will be 50 ohms. L_i is *not* the inductor value used but is modified by the coupling capacitance immediately adjacent, plus any distributed capacitance of the inductor. The latter can be discounted since a 2.2 μH inductor will have little distributed capacitance. A properly-adjusted filter will cancel all reactance at the center frequency since each coil will resonate with the combination of parallel capacitance and adjacent coupling capacitance. All of the matching component formulas take this into account.

Consider the C_a/C_b arrangement, sometimes called the "capacitive voltage divider" match. To find X_i, the reactance of the inductor, X_L, must be found (138.230 ohms at 10 MHz for 2.2 μH); X_{Cc} represents the coupling capacitance of 3.9 pF which will be 4080.90 ohms. Therefore,

$$X_i = \frac{X_{Cc} X_L}{X_{Cc} - X_L} = \frac{4080.9\ (138.23)}{4080.9 - 138.23} = 143.08\ ohms$$

For this type of match, intermediate terms are calculated to simplify computation:

$$b = \frac{R_i X_i}{R_i^2 + X_i^2} = \frac{3740\ (143.08)}{(3740)^2 + (143.08)^2} = 0.0382$$

$$d = bX_i = 0.0382\ (143.08) = 5.465$$

Then:

$$X_{Ca} = R_g \sqrt{\frac{d}{R_g - d}} = 50\sqrt{\frac{5.465}{50 - 5.465}} = 17.516\ ohms$$

$$X_{Cb} = bR_i - \sqrt{d(R_g - d)}$$
$$= 0.0382(3740) - 5.465(50 - 5.465)$$
$$= 142.867 - 243.403 = 127.266\ ohms$$

From the reactance formulas, $C_a = 908.6\ pF$ and $C_b = 127.3\ pF$. The nearest 5% tolerance value for C_a is 910 pF but any combination within 5 percent is adequate.

The C_b value is from the exact solution and would be a trimmable value in actual construction. C_b could also be found from

$$C_b = \frac{C_r C_a}{C_a - C_r}$$

where C_r is the required resonating value. In this case $C_r = 115.1 - 3.9 = 111.2\,pF$. If C_a was fixed at 910 pF, $C_b = 126.7\,pF$. The C_b formula above is approximate but is very close to actual.

The C_s/C_p matching arrangement is simpler to calculate. For the same example

$$X_{Cs} = \sqrt{R_g(R_i - R_g)} = \sqrt{50(3740 - 50)} = 429.5\ ohms$$

From the reactance formula $C_s = 37.1\ pF$ but a fixed 39 pF value will work well. The simple computation of C_p results in values very close to actual and is given by

$$C_p = C_r - C_s = 111.2 - 39 = 72.2\,pF$$

using the fixed value of C_s as 39 pF. As in the other case, C_p would be trimmable. If the C_a/C_b matching system is used in interstages, supply voltage bypass capacitors should be at least a hundred times larger than C_a.

construction

Each section of the filter, the parallel component group, should be shielded from every other section. Even if toroids are used the shielding must be used since they can couple slightly by the magnetic field, more so by electrostatic coupling. Only the coupling capacitors should be common to adjacent sections.

Double foil, one-ounce printed-circuit board material is suitable if all the joints are completely soldered. One large piece can be used as a baseplate with strips forming the sides and interstage shields. Threaded spacer rods can be used in the corners of section compartments, soldered to the board material. With this construction, section tops can be made removable; copper tape should be used on removable tops to insure 100 dB shielding.

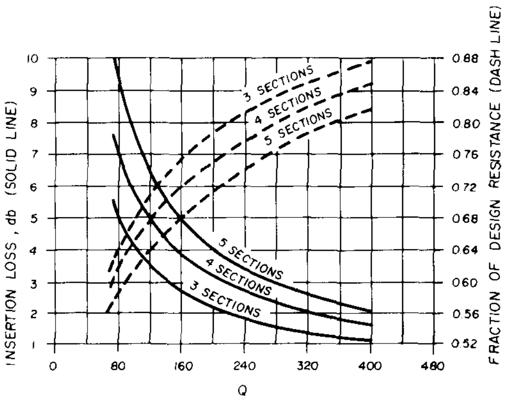


fig. 8. Ideal value insertion loss and end resistance for 3-, 4-, and 5-section, 1-dB Chebyshev bandpass filters.

tuning methods

Fig. 10 shows a setup for tuning each filter section. This may be done as the filter is built, section by section. The loose coupling for end sections must be *very* loose, particularly at high frequencies, to avoid stray capacitance that can cause mistuning.

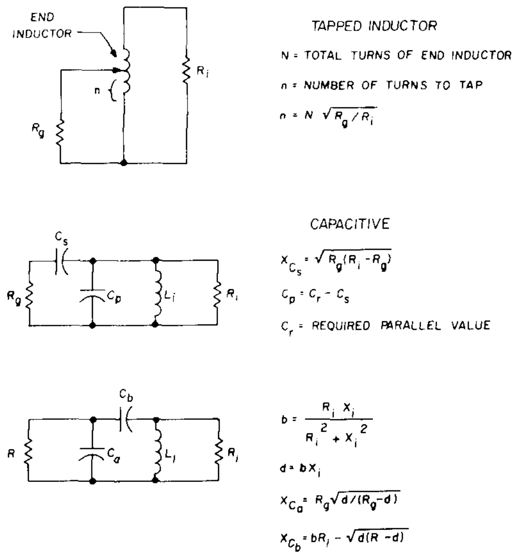


fig. 9. Impedance matching methods for bandpass filters. Complete information on calculating component values is given in the text.

An easier method can be used by the fact of passband rounding at low values of Q. This is simply loading of each section by a 1/4-watt carbon composition resistor of about the same value as the design resistance. All the trimmers are then peaked at the center frequency. Several passes should be made since there is some interaction until all capacitances are balanced.

Fig. 10C shows the setup for coupling the high impedance ends (500 ohms or higher) with the simple tuning method. This is the "Norton Theorem" transformation from a current source to a voltage source with the 10- and 39-ohm resistors serving to isolate the generator and receiver impedances. For 50-ohm end matches, connect directly but use at least 10 dB of attenuation at each end for isolation.

When the simpler method of tuning is completed, the loading resistors are removed and replaced by megohm-range carbon-composition resistors of the same wattage and brand. The reason for replacement is that most all carbon resistors have about 1 or 2 pF of shunt capacitance but the variation between different resistance values of the same wattage and brand is very small. This prevents mistuning by the added capacitance. Film-type resistors have been found to vary greatly with frequency, and some even exhibit inductance instead of capacitance. Where the parallel circuit capacitance is about 400 pF or greater, replacement is not needed since the tuning change is very small.

component Q

Manufacturer's data on toroid Q is quite reliable. In many instances the required inductance needs fewer turns than table data. To keep Q at its highest value, *always use the entire form with even spacing and a* minimum gap between the winding start and finish. Use the largest wire size that will fit the form.

Coating of the finished coil is recommended and ordinary exterior varnish is very good. A double light coating will reduce Q by 7 to 8 per cent. Polyurethane varnish can result in a Q reduction of about 12 per cent and is not recommended. Acrylics and *Q-dope* are generally useless because, in time, moisture can cause lifting of the adhesion area. Acrylics look pretty and are fast-drying but need porous or laquer-compatible surfaces for long life. Q reduction is about the same as varnish.

Dipped-mica and mica-compression trimmer capaci-

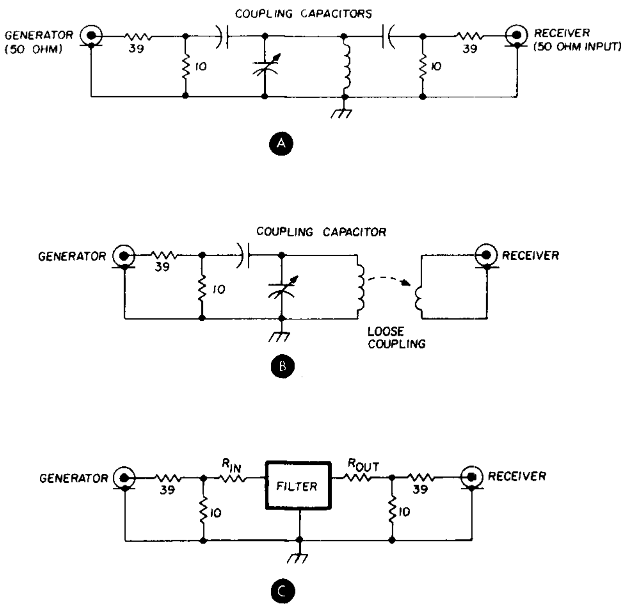


fig. 10. Test setups for tuning the bandpass filters for the desired passband response.

tors have *Qs* of 1000 or more in the 3- to 30-MHz range. Silver micas are good at low frequencies but a few have been found to have a Q as low as 600 at 30 MHz. However, using a Q value of 1000 for dipped-mica types is a good rule. Paralleling mica capacitors will not generally affect Q of the total.

If the capacitor is above about 1000 pF, Q may be lower and may also vary with different types. It is best to measure the higher capacitance values.

Shielding will affect inductor Q. Data on shielding affects of single-layer solenoid coils can be found in many texts. Toroids have been found to exhibit negligible Q reduction if kept at least one form-thickness away from the shielding on any surface. Foam polystyrene used as packing material is a good support in this case.

test setups for Q measurements

Fig. 10A can be used for parallel resonance and Q tests if the coupling capacitors are replaced with 1 pF units above 3 MHz, about 10 pF below. A 1-pF capacitor can be made from a 1/4-inch (6.5mm) square of double-side G10 fiberglass circuit board material. In fact, small-value coupling capacitors can be made this way for higher-frequency filters as part of the interstage shielding but the thickness should be measured and dielectric constant known; keep edges at least 1/8th inch (3mm) from the ground surface and make certain the grounded area of the shield completely surrounds it.

A requirement of this setup is the ability of the receiver to measure 3 dB differences in amplitude from the peak. The signal generator must have a counter attached for accurate frequency differences. If the generator has a variable output attenuator, accurate in dB, the receiver can be operated with manual gain control (turn off agc) and the level controlled by the generator.

To measure Q of the parallel-resonant circuit, set the generator for maximum signal at the receiver and note the frequency. Then tune the generator to either side of this frequency, tracking the receiver tuning, until the signal is 3 dB less than peak and note the frequencies. Find the difference between the two 3 dB frequencies and divide into the peak frequency. The result is the total Q of the parallel circuit.

what happens if the components aren't measured?

About the worst thing that can occur is increased insertion loss and a slight change in bandwidth. This assumes that the coils have been checked for inductance within 10 per cent of the calculated value at least with a dip-meter or inductance bridge.

The usual case is a Q reduction with toroids by turn spreading or bunching to trim it to right inductance. Another is close spacing of shielding on solenoid coils which lowers inductance as well as Q. Slug-tuned inductors can change Q by 50 per cent between slug in and slug out. Pot-core inductors are very good at frequencies below a 1 MHz but often the wrong core material is selected.

Beware of junkbox capacitors with partly legible markings. They may not be what you think they are; they may also be damaged. A multisection filter can still perform with one section grossly off resonance but the insertion loss will be dozens of dBs lower than you would expect.

A very common mistake goes all the way back to initial calculations. Double check your math. Check for section resonance at the center frequency. The best insurance against dropping decimal points is to use a pocket calculator.

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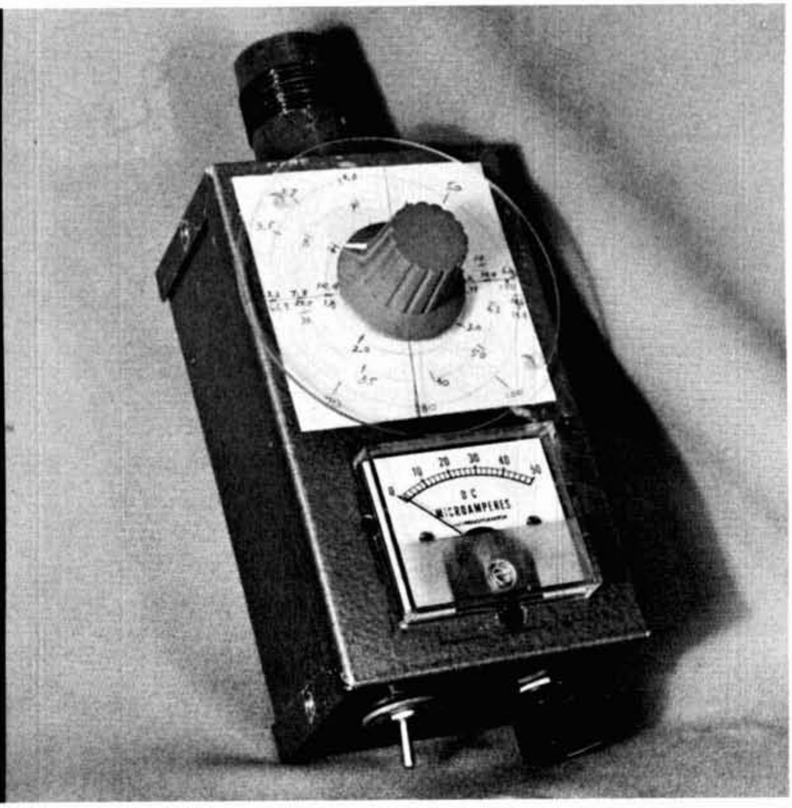


gate-dip meter

that really dips

This gate dipper covers 1.8 - 150 MHz and can be built for less than $25

The dip oscillator described here is a solid-state version of the 6C4 vacuum-tube grid-dip meter in the 1957 *ARRL Handbook*. Unlike some of the single-fet Colpitts circuits described in the literature, this Hartley configuration really performs. In fact, a dip from about 50 - 20 microamps can be obtained on most bands when the dipper is held an inch (25mm) or so away from the resonant circuit under test. Furthermore, few false dips are encountered throughout the tuning range. The circuit is straightforward and the entire project can be completed in a weekend. If all new parts are used the cost should be less than $25, which isn't bad when compared with some of the commercial dip meters available.

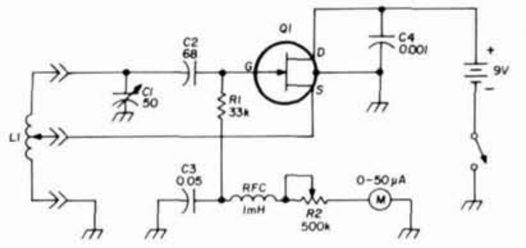
The heart of the instrument is a Siliconix 2N5398 vhf uhf jfet (**fig. 1**); however, an MPF107 (2N5486) should work just about as well. Try several of one type and use the one that yields the highest off-resonance gate current. If you're not interested in the vhf capability, the MPF102 is by far the most economical device choice. Coil data appears in **table 1**.

construction hints

A Minibox houses all parts with plenty of extra space inside as shown in the photo. Components are mounted on a one-inch-square (25mm) piece of copper-clad board, whose surface was sectionalized with a hacksaw. This board is located directly at the coil-socket terminals to minimize stray inductance; otherwise vhf operation will be limited.

At this point a few general comments are in order. My model contains a 50-pF tuning capacitor with one plate removed. If your tuning range is a little different, the reason can be attributed to a difference in tuning-capacitor value. The coil tap position becomes more critical at the higher frequencies. A point can be reached

By Charles G. Miller, W3WLX, Associate Professor, Anne Arundel Community College, Arnold, Maryland 21012



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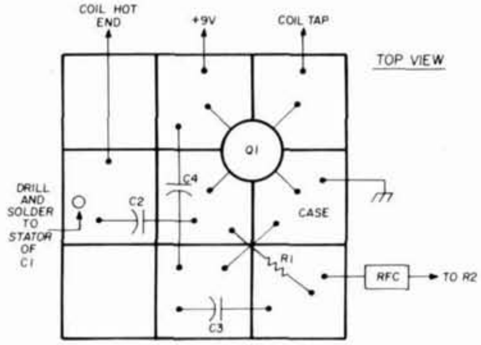


B

fig. 1. Schematic of the gate-dip oscillator, A, and suggested circuit-board layout, B (not to scale). C1, 50 pF (Hammarlund HF50 or equivalent). Q1 is a 2N5398, 2N5486, or MPF102 (see text). Coil data is shown in table 1. Power is supplied by a 9-volt transistor-radio battery.

where the dip is definitely most pronounced; therefore, some trial and error tinkering with the coil-tap point may be in order if you're really particular. The vhf coil is especially critical in this respect. Fortunately, the tap point is easily moved about on this hair-pin-shaped coil.

number of turns will change. Remember that inductance varies as the square of the number of turns.

calibration

Calibration was accomplised by coupling the dipper to a coil (about 40 turns, 1 inch (25mm) diameter), which in turn was connected to a Hewlett-Packard 5300A counter for frequencies up to 50 MHz. A 5328Z counter was used for the higher frequencies. A receiver or another calibrated dip meter can also be used, of course.

Since you may prefer to design your own dials and chassis layout, I've left this part up to the reader. The photos show how I arranged the parts and made the

Open Minibox showing component layout. The 9-volt battery is mounted in the bottom half of the Minibox. One-hand operation, easy construction and low-cost parts make this a nice piece of test equipment.

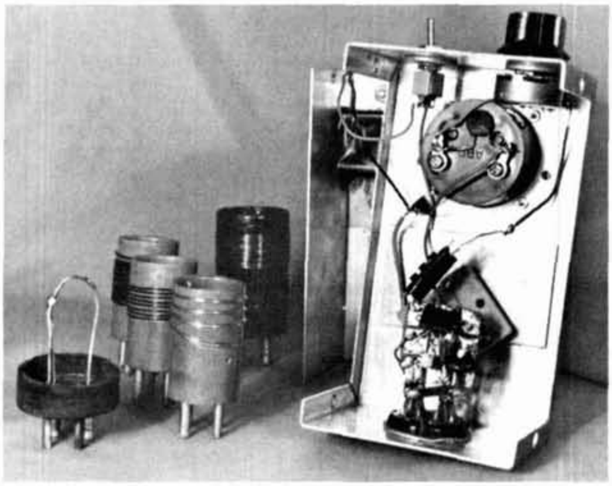
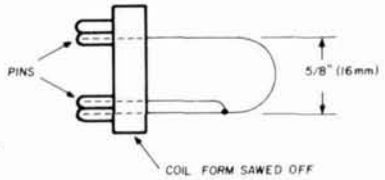


table 1. Coil data.

frequency range (MHz)	no. turns	wire size AWG	(mm)	winding length inches	(mm)	tap*	coil diameter inches	(mm)
1.8 - 3.8	82	26 enamel	(0.4)	1 9/16	(40.0)	12	1¼	(32)
3.6 - 7.3	29	26 enamel	(0.4)	9/16	(14.5)	5	1¼	(32)
7.3 - 14.4	18	22 enamel	(0.6)	3/4	(19.0)	3	1	(25)
14.4 - 32	7	22 enamel	(0.6)	1/2	(12.5)	2	1	(25)
29 - 64	3½	18 tinned	(1.0)	3/4	(19.0)	3/4	1	(25)

61 - 150 Hairpin of 16 no. AWG (1.3mm) wire, 5/8 inch (16mm) spacing, 2 3/8 inches (60mm) long including coil-form pins. Tapped at 2 inches (51mm) from ground end.



*Turns from ground-end. 1 inch (25mm) forms are Millen 45004 available from Burstein-Applebee

I made a coil-capacitor combination that resonated at about 100 MHz to adjust the vhf-coil tap for the best meter null.

The total cost of this project can be reduced somewhat if you have some old four-prong vacuum tubes lying around. Their bases serve admirably as coil forms. The diameter of these forms is different from the one-inch (25mm) forms specified in the parts list, so the

frequency dial. This dial arrangement, by the way, is a larger version of that described in the *ARRL Handbook* article and allows for convenient, one-hand operation.

acknowledgement

I wish to extend my appreciation to Professor Jim Privitera for the photographs in this article.

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the FCC, harmonics, spurious radiation, 40 dB, and you.

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toroid permeability meter

Complete design data and construction details for an instrument to determine the permeability of unknown toroid cores

Small toroidal inductors wound on cores of various permeabilities are being used today in great numbers. I obtained from a surplus source some toroid cores of different sizes and colors having an assortment of colored stripes and dots, plus house numbers of many digits, imprinted on them. To determine what kind of cores these are, in terms of permeability, you could wind a coil with a reasonable number of turns (about 10), and measure the resultant inductance using a bridge, Q meter, or grid-dip oscillator. Then the well-known formula for the inductance of a toroidal coil can be used to compute the permeability:

$$L = 0.004046\,\mu\,n^2\,h\,\log_{10}(b/a) \qquad (1)$$

where L = inductance (μH)
μ = permeability (non-dimensional)
n = number of turns
h = core height through hole* (cm)
b = outside diameter (cm)
a = inside diameter (cm)

However, it would be easier to place the unknown core into an instrument and directly determine permeability. Such a hands-off approach is what this article is all about: how to build a toroid-core permeability meter. The instrument described here uses as instrumentation only a grid-dip oscillator whose dip frequency signal is

*Height is the axial distance, not the distance measured along the coil radius, which is the thickness.

By Milton Ash, W6RJO, 455 21 Place, Santa Monica, California 90402

monitored by a communications receiver with accurate frequency readout. The permeability meter has been designed with the average experimenter in mind and is built of surplus materials.

background

In 1924 G. A. Kelsall[1,2] constructed a permeability meter for measuring the permeability of large iron toroidal windings used in power-line-frequency electrical machinery and low audio frequencies of that era. In 1953 his designs were modified by Haas, Edson, and others[3,4] of the National Bureau of Standards, for the construction of an rf toroid permeability meter. With simple modifications, both the Kelsall and Haas instruments are also capable of measuring the core dissipation factor (loss tangent) and permeability temperature coefficient as well as the permeability. Only the latter is of interest here. The references can be consulted for information on the other parameters.

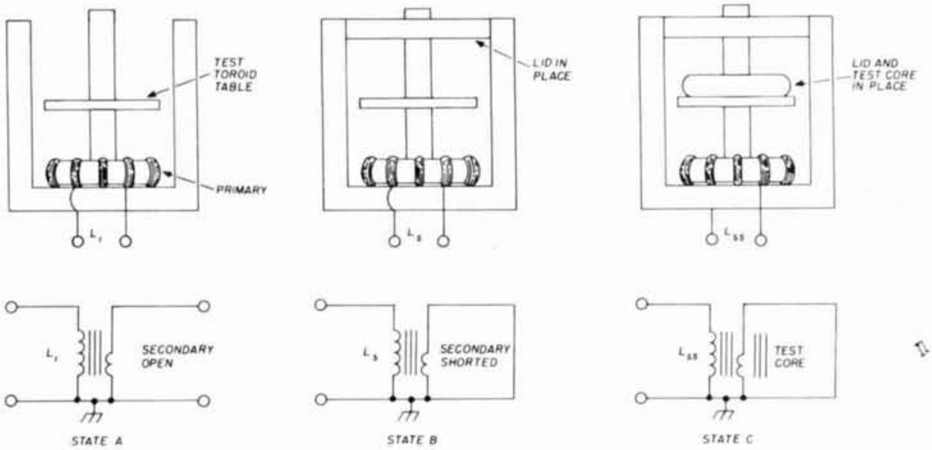
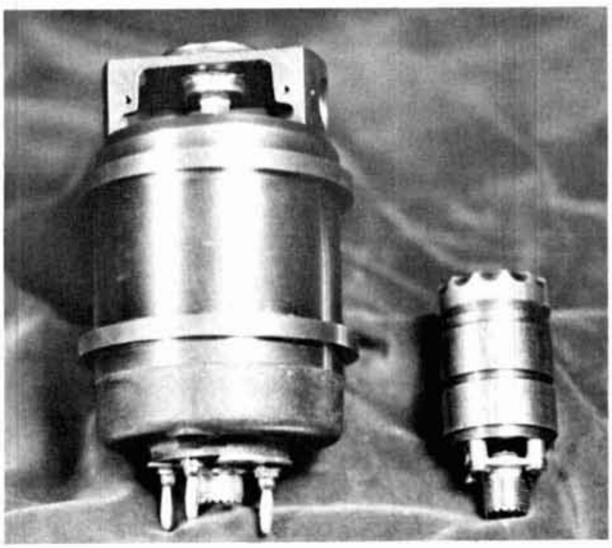


fig. 1. Toroid permeability-meter states. The permeability meter is a step-down transformer consisting of a copper or brass cup and lid secondary, with a reference-toroid-core primary lying on the inside bottom of the cup.



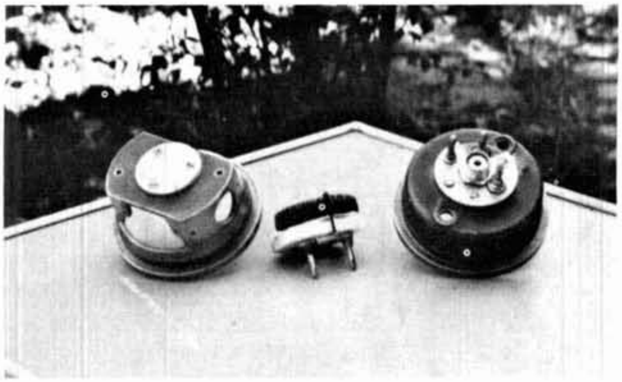
Selsyn-case cup, left, and minicup permeability meters. On the selsyn-case cup note the large end of the upper half of the pop-up stem center conductor soldered to its shaft as well as to the top end bell. Banana plugs and coax-connector primary toroid-coil terminals protrude from bottom end bell.

design

The rf toroid permeability meter is essentially a step-down transformer of novel design. The primary winding is on a toroidal core, in this case, about 2 inches (50mm) in diameter with n turns (the value of n is discussed later). The primary winding is lying on, or plugged into, the inside bottom of an approximately 3-inch (77mm) diameter copper or brass cup. The cup contains a center conductor that passes through the toroid core primary and fastens to the cup lid, as in **figs. 1** and **2**. The cup and center conductor, with its lid in place, form a husky, three-dimensional, one-turn, shorted secondary winding. The lid shorts the secondary by connecting the center conductor to the cup wall. With the lid off, the secondary is open since that connection is broken.

The permeability meter provides an input inductance to be measured across the primary winding for each of three cup states, **A**, **B**, and **C**. In state **A**, **fig. 1A**, the cup lid is off (open secondary). In state **B**, **fig. 1B**, the cup lid is on (shorted secondary). In state **C**, **fig. 1C**, the cup lid is also on, but with the core whose permeability is sought in place on its table in the cup. The table is made of low-loss material, such as Teflon, and is positioned in the upper third of the cup. (This position is not critical.)

Fig. 1 shows the three states of the permeability meter when checking the permeability of an unknown core. L_l, the input inductance measured across the reference toroid coil primary with the secondary open, is the largest of the three inductances. L_s, the input inductance with the secondary shorted, is the smallest of the three inductances because it represents the primary with reduced inductance because of the shorted secondary negative reactance reflected back to the primary. L_{ss}, the input inductance with the secondary shorted, but with an increased secondary inductance because of the addition of the unknown core inductance in the second-



Exterior view of top and bottom cup end bells. At left is the top end bell showing circular brass plate covering the top central hole. In the center is the primary toroid coil on a Teflon base with Teflon sheet pad and banana-plug terminals. At right is the bottom end bell showing three banana plugs (one is a dummy) and coax-connector primary-coil outlet terminals. Notch in bottom brass plate is a quick fix for a too-large hole. The normal plate has an intact circumference.

ary magnetic circuit, is always intermediate in value between L_l and L_s; *i.e.*, $L_l > L_{ss} > L_s$.

testing a core

For those who have the instrumentation, the rest is easy. After the instrument is built, and with its lid on and a test core in place, (state **C**), L_{ss} is measured on an inductance bridge. Then, with the lid off and no test core (state **A**), L_l is similarly measured. With the lid on and no test core (state **B**), L_s is also similarly measured. From the analysis in the **appendix** (**eq. 1-11**) the permeability of the unknown core is given by:

$$\mu - 1 = \left[0.004046\ h \log_{10} \frac{OD}{ID}\right]^{-1} \cdot \left[\frac{L_l^2}{n^2 (L_l - L_s)}\right] \cdot \left[\frac{L_{ss} - L_s}{L_l - L_{ss}}\right] \qquad (2)$$

OD = outside diameter of unknown core
ID = inside diameter of unknown core
L_l = input inductance in state **A** (μH)
L_s = input inductance in state **B** (μH)
L_{ss} = input inductance in state **C** (μH)

The other quantities are defined following **eq. 1**. Note that the middle factor of the expression above is a constant of the individual instrument. This will facilitate its calibration, as discussed later.

Using an electronic calculator with a log capability, permeability can be computed by plugging into **eq. 2** the measured value of input inductance, number of turns of the primary inductance, and the unknown core dimensions.

Suppose you have no sophisticated instrumentation or calculator? The remainder of this article is devoted to the problem of determining the permeability of unknown cores using a permeability meter built from surplus and other readily available low-priced materials. The permeability meter operates in conjunction with a grid-dip oscillator and a communications receiver with an accurate dial readout (3-figure accuracy below 10 MHz). The unknown permeability is obtained from a nomogram, together with the grid-dip oscillator frequency-dip data and the unknown core dimensions.

construction

As the accompanying figures show, the essential requirement is a copper or brass cup with a removable threaded or beveled lid that also holds the cup center conductor. For example, a center conductor made from 1/4-inch (6.5mm) threaded brass rod could be used with the cup. This, of course, would limit core sizes to IDs greater than 1/4 inch (6.5mm) so that the cores can slip over the center conductor when placed into the cup. Cup diameter should be large enough to accommodate the largest cores anticipated by the user. A cup diameter of 3½ inches (89.3mm) is used here. Cup height-to-diameter ratio isn't critical but should be close to unity (height ≈ diameter).

Two banana plugs and jacks provide plug-in capability inside the cup for the primary toroid coil and connections to the outside. The plugs and jacks are soldered back-to-back after removing the threaded portion of the banana plug. One plug is insulated from the cup bottom with Teflon shoulder washers, and one is in contact with the cup (see **fig. 2**). A third dummy banana plug is also provided to extend plug-in versatility, as explained later. This completes the instrument except for the primary coil (described below). I added an SO-239 coax connector to connect other test equipment, but this connector isn't necessary. The primary coil is mounted on a base of low-loss material with banana plugs, as shown in **fig. 2** and the photos.

Interior view of top and bottom end bells. At left is the interior of the top end bell integral (soldered) with the upper half of the pop-up stem center conductor. The center is the Teflon block table leaning on the wall of the central brass shell of the instrument. Right shows the interior of the bottom end bell with the jack ends of the banana-plug and jack primary-coil connectors. Note window of large end of pop-up stem to accommodate jumper wire (with sleeving) from coax center conductor to insulated banana-plug/jack. The lower plug/jack is grounded to the cup bottom.

Making the case. Many inexpensive surplus selsyn motors are available that are about 3 inches (77mm) in diameter by about 5 inches (128mm) high. These are typified by a 110 V, 60 Hz, 5 ampere, type M unit. They have a brass case and heavy flanges bearing manufacturing dates between 1943-1956.* All we need is the brass case. Remove the electrical terminals with a screwdriver, then remove the field coils using a claw hammer. Use pliers to remove the thin field-coil laminations, two or three at a time. Easy does it. The case will now come apart in three pieces: top and bottom end bells and center shell.

At each end of the bare case is a circular brass plate covering a 7/8-inch (22mm) hole. It turns out that a brass lawn sprinkler pop-up stem† found in hardware stores has a large end that just fits into the selsyn case end holes. Two of these pop-up stems screwed back-to-back also just fit into the selsyn case to make the permeability meter center conductor. The pop-up stems have a partially threaded bushing, one of which should be threaded all the way through so that the center conductor can be screwed and unscrewed as the instrument lid is installed and removed during use.

Soldering the large end of each pop-up stem onto its shaft, as well as into the end bells of the selsyn case, can be done easily with a small Butane torch. First, however, the bottom end bell of the selsyn case should be drilled with three holes through the bottom brass plate for the banana plugs and jacks for the primary toroid coil outlet, as mentioned previously. The three holes form a right triangle on the circular brass plate, with the insulated hole at the right-angle corner and the two non-insulated holes at the other two corners. The distances from the insulated hole to the noninsulated holes, along the short sides of the triangle, are 1 inch and 3/4 inches (25.5 and 19mm) respectively. These dimensions correspond to most inductance-bridge and Q-meter-terminal spacings.** These, and other dimensions, are shown in **fig. 2**.

The bottom end bell central hole will take a nut-type coax connector. This type of connector mounts with two large mounting nuts in a 5/8 inch (16mm) hole, instead of the usual square SO-239 mounting. The connector is then mounted in the circular brass end plate, with a 5/8 inch (16mm) hole punched out. This assembly is screwed back into the bottom end of the cup, as shown in **fig. 2**. Also, before soldering the pop-up stem into the bottom (done with the brass plate in, but with its accessories removed), a small window must be notched out of the large end of the stem to provide access to the SO-239 connector center conductor. The latter is jumpered with a short piece of sleeve-covered wire to the insulated banana jack inside the cup. This is the only wiring in the instrument. This connection is shown in

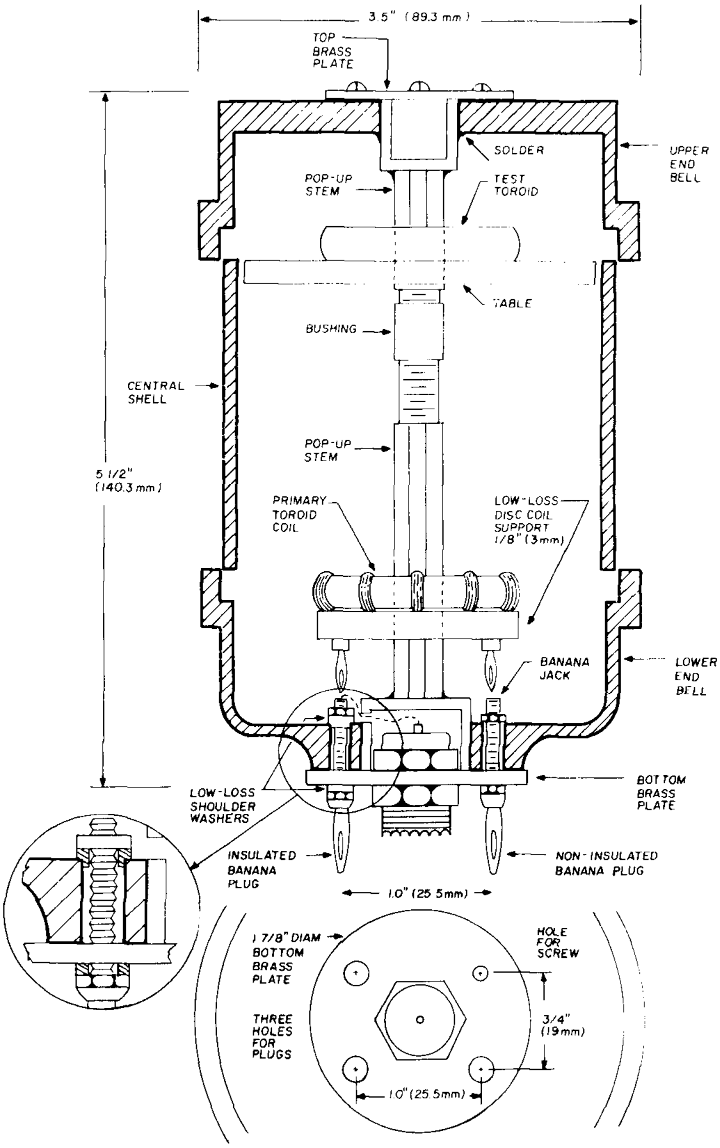


fig. 2. Toroid permeability meter built from a surplus selsyn-motor case and a lawn-sprinkler pop-up stem.

fig. 2. It is made after soldering the pop-up stem in place.

Primary coil. The primary coil consists of 31.6 turns (so that n^2 will equal 1000), no. 24 AWG (0.5mm) wire, which is equally spaced around an Arnold Engineering Company toroid core.* Any other core of equivalent size, quality, and permeability may be used providing it meets the constraints discussed later on using the grid-dip oscillator/amateur receiver combination to determine unknown core permeability.

From **fig. 2** and the photographs it is seen that the cup lid; *i.e.*, the upper end bell and the upper half of the

*Manufactured by Henschel Corporation, Amesbury, Massachusetts.

†Manufactured by Champion Brass Company, 1460 Naud Street, Los Angeles, California 90012. Available from Sears.

** Either of the two noninsulated plugs is a dummy, since the primary coil needs but two jacks to be seated inside the cup bottom. The dummy plug is included so that all three plugs, two at a time, accommodate most bridges. Of course, the insulated plug is one of the two used.

*Arnold Engineering Company, P.O. Box G, Marengo, Illinois 60152. Dimensions are 1.875 inch (48mm) OD; 1.375 inch (35mm) ID; and 0.375 inch (9.6mm) in height. Permeability, μ, is 125. Part no. is D18002/AM-12.

center conductor, are now integral and comprise the upper portion of the instrument. The middle portion consists of the thin brass shell, while the bottom end bell contains the lower half of the center conductor and the banana plugs and jacks for the primary coil. The central shell is a press-fit into the bottom end bell of the cup. It is also a press-fit into the top end bell as the center conductor is screwed together during test of a core. With a low-loss (Teflon) table or support strip drilled and tapped for the center conductor, where the table is positioned to occupy the upper third of the cup, the assembly is complete.

Minicup permeability meter. This instrument, which is also shown in the photos, is about salt shaker size and is used for measuring the permeability of tiny cores. It is made from scrap brass and is 1.5 inch (38mm) OD, 1.25 inch (32mm) ID, and 1.75 inch (45mm) long. It is mounted on an SO-239 chassis-mount connector. The primary coil is 31.5 turns of no. 28 AWG (0.3mm) wire spaced on an Amidon T-94-2 (red) core. The primary leads feed through a cup bottom hole; one is soldered to the SO-239 center conductor and the other is grounded to the cup. The center conductor is a 6-32 (M3-5) machine screw. The lid is made from a piece of cylindrical brass scrap.

In the appendix, **eqs. 1-13** and **1-14** give another version of the formula for permeability shown in **eq. 2**. This formula is:

$$\mu - 1 = \frac{k}{h\log_{10}(OD/ID)} \frac{(f^2_{HI} - f^2_{cor})}{(f^2_{cor} - f^2_{LO})}$$

$$f_{HI} > f_{cor} > f_{LO} \qquad \textbf{(3)}$$

k = a constant of the instrument obtained by calibration as discussed below

f_{HI} = the upper grid-dip oscillator frequency measured with no core inside the cup, and with the lid on. This corresponds to the L_s state **B** of the instrument

f_{LO} = the lower grid-dip oscillator frequency measured with no core inside the cup, and with the lid off. This corresponds to the L_l state **A** of the instrument

f_{cor} = the intermediate grid-dip oscillator frequency measured with the unknown core inside the cup, positioned on its table, and with the lid on. This corresponds to the L_{ss} state **C**

The other quantities in **eq. 3** are defined following **eqs. 1** and 2. Now, the analysis in the appendix (**eq. 1-15**) modified as above for the constant, k, gives for this constant and the secondary inductance, L_{coax},

$$k = \frac{L_{coax} - L_l/n^2}{0.004046} \qquad \textbf{(4)}$$

and

$$L_{coax} = 0.004046\ H \log_{10}(D_{cup}/D_{cen})$$

L_{coax} = secondary inductance (μH)
D_{cup} = cup inner diameter (cm)
D_{cen} = center conductor outer diameter (cm)
H = cup height (cm)

A slight digression is appropriate at this point concerning the structure of the toroid inductance formulas. First, **eq. 1** for the inductance of a toroid contains three main factors:

1. The permeability, μ, which is a measure of the inductance enhancement of the coil over its free (air-wound) space version.

2. The square of the number of turns, n^2.

3. The factor $0.004046\ h\ log_{10}\ (OD/ID)$, which is the inductance in μH of the free space that the toroidal core volume occupies.

Now, the shorted cup secondary surrounds a toroidal volume of free space and thus has an equivalent inductance, L_{coax}, as in **eq. 4**. Note that this formula has the same structure as the free-space inductance of the unknown toroid, **eq. 1**, as it should. L_{coax} also means that it is, at the same time, the inductance of a short, fat section of air-dielectric coax of length H and diameters D_{cen} and D_{cup}, which also happens to surround a toroid volume of free space.

Second, the formulas above for permeability hold ideally *only* when the primary leakage inductance is zero or negligibly small, as discussed in reference 3. However, with the number of bolt holes and cast cutouts in the surplus selsyn case, the primary leakage inductance can be appreciable since some of the primary flux that threads these holes is sure not to link the secondary, which contributes to the leakage flux and thus to the leakage inductance. Hence, the above constant, k, is not given accurately by **eq. 4**. Instead, k will be found by calibration with a number of cores of known permeabil-

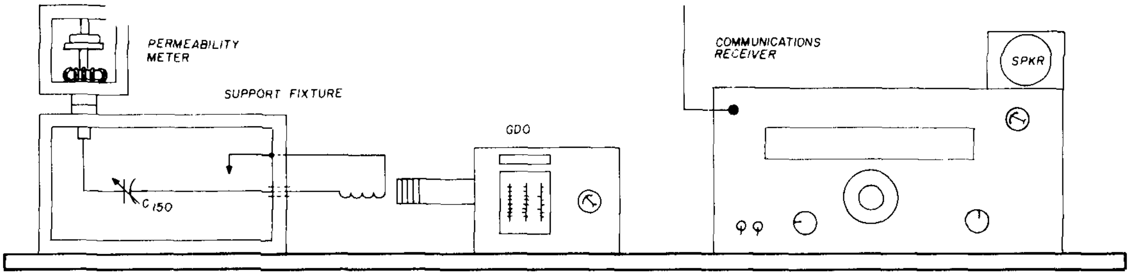


fig. 3. Instrumentation for obtaining input to the nomograms.

table 1. Toroid permeability meter calibration data obtained with primary toroid coil in the 7-8 MHz range.

serial number	core type (1)	known perm, μ	freq (2) (MHz)	OD inches	OD (mm)	ID inches	ID (mm)	height inches	height (mm)	f_{LO}(3) (MHz)	f_{HI}(3) (MHz)	f_{corr}(3) (MHz)	k(4)	calculated permeability
1	T94-10 (blk)	6	60-150	0.942	(24)	0.560	(14)	0.312	(8)	7.090	7.984	7.903	8.47	6.94
2	T94-6 (yel)	8	10-90	0.942	(24)	0.560	(14)	0.312	(8)	7.087	7.993	7.884	8.64	9.10
3	T94-2 (red)	10	1-30	0.942	(24)	0.560	(14)	0.312	(8)	7.086	7.982	7.862	9.78	10.2
4	T94-3 (grey)	35	0.05-0.5	0.942	(24)	0.560	(14)	0.312	(8)	7.089	7.989	7.696	11.93	29.5
5	FT-114-63	40	1.5-25	1.142	(29)	0.748	(19)	0.295	(7.5)	7.087	7.980	7.613	7.26	54.7
6	T94-41 (grn)	75	0.001-0.10	0.942	(24)	0.560	(14)	0.312	(8)	7.077	8.004	7.894	0.81	9.07
7	FT-114-61	125	0.2-10	1.142	(29)	0.748	(19)	0.295	(7.5)	7.087	7.984	7.545	16.75	75
8	FT-114-43	950	60-200	1.142	(29)	0.748	(19)	0.295	(7.5)	7.085	7.986	7.153	9.95	955
9	FT-114-72	2000	0.001-1	1.142	(29)	0.748	(19)	0.295	(7.5)	7.065	7.995	7.091	13.35	1498
10	Arn. Eng AM-12	125	0.2-20	1.875	(48)	1.375	(35)	0.375	(9.6)	7.089	7.992	7.646	24.08	111

notes:
(1) Amidon Associates nomenclature except core 10. T prefix indicates powdered iron; FT prefix indicates ferrite.
(2) Manufacturer's recommended frequencies.
(3) Meaured using a 1947 Millen grid-dip oscillator and surplus ARR-41 receiver.
(4) Calculated using **eq. 3** with frequency measurements described in text and known permeabilities. $k_{AV} \approx 10$.

ity. Again, k could ideally be calculated from **eq. 4**, and no calibration would be necessary if, a) the leakage inductance were negligible, and b) the correct cylindrically equivalent physical dimensions; *i.e.*, H, D_{cup}, D_{cen}; of the selsyn cup, were known.

For calibration as well as for normal use, the permeability meter is connected in series with a variable capacitor and a 2- or 3-turn link and coupled to a grid-dip oscillator, as in **fig. 3**. A communications receiver with a reasonably accurate dial readout is used to pick up the grid-dip oscillator signal. A small antenna about a foot (30cm) long on the receiver is sufficient. This combination of a common grid-dip oscillator and a medium-quality communications receiver should provide a frequency readout to three figures. Amateur-band-only receivers can usually provide better accuracy. However, the permeability meter must be matched in physical construction so that the frequency spread, $\Delta f = f_{HI} - f_{LO}$, doesn't span more than 500 kHz, which is the usual vfo range on most ham-band-only receivers. This constraint is discussed in detail in a later part of the article.

calibration data

For ten cores of known permeability, claimed to be correct within 5-10 per cent, **table 1** gives the grid-dip oscillator/receiver data I used to calibrate the permeability meter shown in the photos. Using an average k of about 10 from the computations described above, unknown toroid permeabilities should be about ±10-20 per cent of their actual values. This can be seen by examining **table 1** data and by computing the deviation of k from the average k for each core (except core no. 6). Other statistics, such as the mean-square deviation of the permeabilities for a series of readings on each core, are left as an exercise for the interested reader.

Core no. 6 gave a computed permeability way out of the ballpark compared with its claimed value. This is probably because it was measured in a frequency range of 7-8 MHz, which is remote from its intended use range (0.001-0.1 MHz). Data from this core was therefore not used in computing the average k.

using the nomogram

As mentioned earlier, for those who don't have calculators with a logarithm capability, or inductance bridges or Q Meters to obtain the unknown toroid permeability, **figs. 4** and **5** comprise a nomogram to use for this purpose. The unknown toroid dimensions and grid-dip oscillator frequency-dip readings are required as input to the nomogram. **Eq. 3** for the permeability, μ, slightly modified as in **eq. 5**, is the basic equation from which the nomogram is constructed. If your permeability meter is about the same size as mine, the constant, k, will be about 10, as seen from **table 1**. For other sizes of instruments, and thus other values of k, corresponding adjustments can be made in the nomogram as explained below.

A second parameter of the instrument is the ratio f_{LO}/f_{HI}, called a, of the instrument. It appears when **eq. 3** for the permeability, μ, is rewritten as

$$\mu - 1 = \left[\frac{k}{h\log_{10}(OD/ID)}\right] \bullet \left[(1 - f^2)/(f^2 - a^2)\right] \quad \textbf{(5)}$$

where $f = f_{cor}/f_{HI}$, and the other quantities are as defined earlier. The nomogram in **figs. 4** and **5** will yield permeabilities of unknown toroid cores for corresponding sets of h, OD, ID, a, and f values.

Directions for use. The following directions for using the nomogram are augmented later by a step-by-step example. For now, noting the sketches at the bottom of **fig. 5**, the directions are as follows.

1. Beginning on **fig. 4** nomogram, with a transparent straightedge, find the value of the unknown toroid core OD on line 1 and its ID on line 2. Span these values and lightly mark where the straightedge crosses line 3.

2. Span the straightedge from the line 3 mark to the value of height on the toroid height line 4, then lightly mark where it crosses diagonal line 5. Note the value of the latter mark as it will be needed later.

3. Referring to the **fig. 5** nomogram, from right to left, and knowing the values of a and f from the grid-dip oscillator frequency measurements, span from the a value on line 6 to the f value on line 7. Lightly mark where the straightedge crosses line 8.*

4. Spanning from the mark on line 8 to f on the f line (line 9), lightly mark where the diagonal line 10 is crossed.

5. Noting the value on line 10, find the same value on the diagonal line 10 on the **fig. 4** nomogram.

6. Span from the **fig. 4** line 10 value to the diagonal line 5 value on the same figure. Mark where the straightedge crosses the permeability line 11.

What you have really found on line 11 of **fig. 4** is $\mu - 1$ but for all except the smallest values of permeability, the above line 11 mark is essentially the permeability (within the accuracy of the instrument and the nomogram procedure). For small permeability values near one, merely add one to the value found on line 11.

For those whose permeability meter dimensions are radically different from this selsyn-case instrument, the line 11 scale should be interpreted as $10(\mu - 1)/k$, where k is the constant for your instrument. With a number of cores of known permeability, you will be able to determine a k for your instrument such that the permeability line 11 will fit the known toroid permeabilities. Then you can relabel the scale on line 11 accordingly. This is the reason for the numbers less than one that are on the line 11 scale. For the minicup permeability meter, k is 3.5, while k for the selsyn-case cup is 10.

Examples. Here are examples for obtaining the permeability of three unknown toroids. The first example involves a toroid from my batch of surplus cores. The core is medium gray in color with the black numerals 262-55582-A2 around the edge. Its dimensions are 1.375 inch (35mm) OD, 0.94 inch (24mm) ID, and height is 0.35 inch (9mm). Grid-dip oscillator/receiver frequency-dip measurements were: f_{LO} = *7.07 MHz*, f_{HI} = *7.98 MHz*, and f_{cor} = *7.46 MHz*. Immediately, a is *0.885* and f is *0.934*. Using a triangle for a straightedge, the procedure is:

1. On **fig. 4** span from an OD of 1.375 inch (35mm) on line 1 to an ID of 0.94 inch (24mm) on line 2. This yields 0.15 on line 3.

2. Span from 0.15 on line 3 to the core height of 0.35 inch (9mm) on line 4. Note that the straightedge crosses line 5 at 0.14. Note this number as it will be used later.

3. Go to the nomogram in **fig. 5**. Span the straightedge

*On lines 6 through 9 of **fig. 5**, two quantities represent each line. In each case, either one can be used, whichever is convenient, as the lines are scaled accordingly.

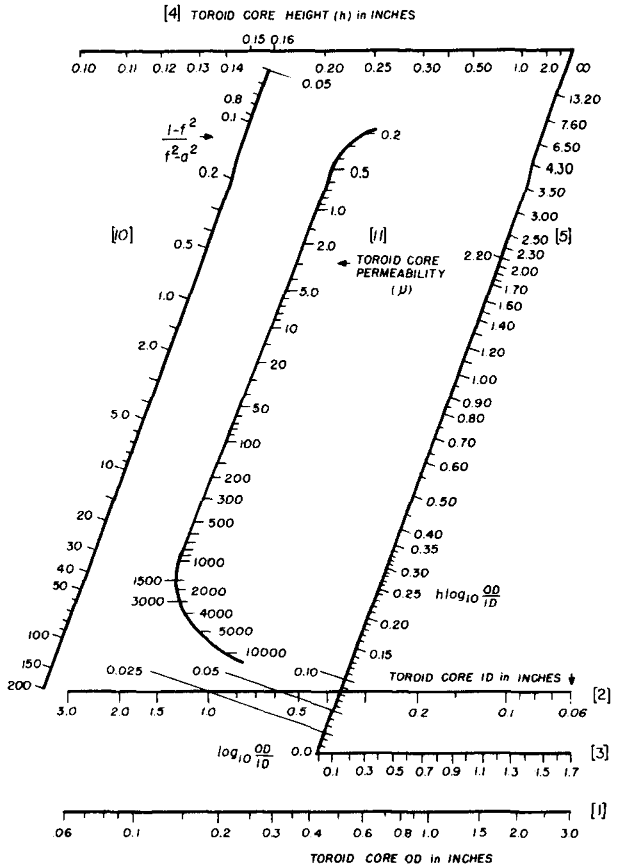


fig. 4. Nomogram for toroid dimensions and core permeability.

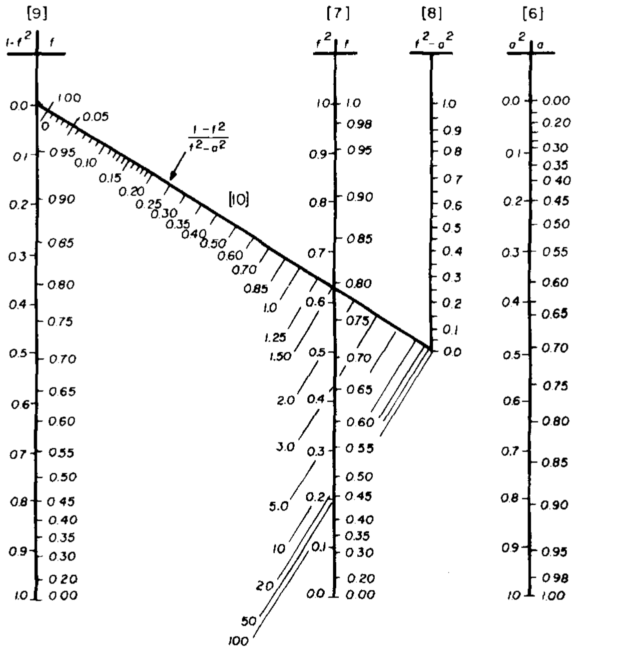
from a = *0.885* on line 6 to f = *0.934* on line 7. The straightedge crosses line 8 at about 0.10.

4. Span from 0.10 on line 8 to f = *0.934* on line 9. The straightedge crosses diagonal line 10 at 1.5.

5. Go back to the **fig. 4** nomogram. Find 1.5 on diagonal line 10. With 1.5 on diagonal line 10 and 0.14 on diagonal line 5, which you obtained from **step 2**, note that the straightedge span between these two numbers crosses the diagonal line 11 at about 90, which is the permeability of this core.

So the permeability of this toroid core is 90 ±10-20 per cent. When I obtained the grid-dip oscillator frequency dip data with the core inside the cup and lid on to get f, the dip was rather broad compared with dips without the core, and the corresponding grid-dip oscillator dial feel was sluggish. This implies that the toroid Q is not very high, at least in the range between 7-8 MHz, as compared with the usual toroid rf core. I would estimate the Q in this case, from the reciprocal of the fractional frequency spread, at about 10.

The conclusions are that this toroid core has a moderately high permeability of about 100 and probably was used in an af or i-f application. The surface, after some paint was scraped away, had the dull gray color and texture associated with powdered iron. This is meant in the sense that powdered iron is supposed to be dull gray in color and somewhat frangible, while ferrite is supposed to be darker, shinier,



THE SKETCHES BELOW ARE A PICTORIAL DESCRIPTION OF STEP-BY-STEP USE OF THE NOMOGRAMS IN FIGS. 4 AND 5.

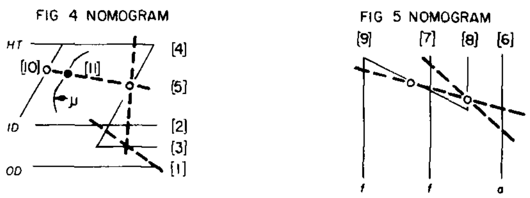


fig. 5. Nomogram for grid-dip oscillator/communications receiver frequency data.

and harder. Lending some credence to this rationale is that the core was roughly textured, which smacks of hurried manufacture.

For a second example, two toroid cores from my surplus batch are tan colored and both are imprinted with the red-colored edge number 2090257, surrounded by two red squares and a tiny five-pointed star with what looks like "AL" in its center. On the opposite edges are dark brown and dark red dabs of paint. Both have ODs of 0.8 inch (20.5mm), IDs of 0.5 inch (12.5mm), and heights of 0.25 inch (6.5mm).

The grid-dip oscillator/receiver measurements for the dark brown and dark red cores gave $a = 0.863;\ 0.863$ and $f = 0.943;\ 0.945$ respectively. Running them through the nomograms in the manner described above yielded permeabilities of 60 for the dark brown and 55 for the dark red core. (Their permeabilities are probably identical.) Scraping the paint on each revealed a shiny solder-like surface coating, and beneath that, the dull gray powdered-iron appearance. They seem to have been carefully made, as all edges had the appearance of having been tumbled before painting. The grid-dip oscillator dips seemed sharp, so their *Qs* are reasonably high at rf – probably a high i-f, low rf application.

A third example is an all-black, unpainted core with no markings of any kind. Its OD is 0.875 inch (22mm), ID is 0.56 inch (14.5mm), and height is 0.25 inch (6.5mm). Grid-dip oscillator/receiver data was $a = 0.863$ and $f = 0.876$. The nomograms yielded a permeability of about 850. The grid-dip oscillator dip was sharp, so the Q is high at rf. The core is well made of what seems to be hard ferrite material.

grid-dip oscillator/ham-band-only receiver combination

As mentioned earlier, the grid-dip oscillator frequency spread, $\Delta f = f_{HI} - f_{LO}$, for this selsyn-case instrument is about 1 MHz. This spread is about right for my surplus ARR-41 receiver, which tunes 1 MHz with each band change. For amateur-band-only receivers, the usual vfo range is 500 kHz per band, with a frequency readout-accuracy of three or four places below 10 MHz (to 1 kHz) being commonplace.

To exploit this readout accuracy, which means to make the Δf of the permeability meter 500 kHz or less to fit the vfo range, implies the existence of a design criterion of some sort. From the appendix (**eq. 1-18**) such a relationship is obtained between the primary toroid inductance, L_l, the secondary cup inductance, L_{coax}, and the fractional frequency spread, $\Delta f/f_{LO}$. It is

$$L_l/n^2 L_{coax} = 1 - \frac{1}{(1 + \Delta f/f_{LO})^2} \quad \textbf{(6)}$$

To illustrate this criterion, consider two examples.

1. Suppose you'd like to use, for grid-dip-oscillator measuring purposes, the 500 kHz of the 20-meter position of a ham-band-only receiver. Then the corresponding fractional frequency spread is

$$\Delta f/f_{LO} = 0.5\ MHz/14\ MHz = 1/28 \quad \textbf{(7)}$$

Note that, from **eq. 1**, $L_l/n^2 = \mu_l L_{la}$ as well, where μ_l is the permeability of the primary core, and the inductance of the volume occupied by this is $L_{la} = 0.004046\ h_l \log_{10} (OD)_l/(ID)_l$ in μH. Subscripts l refer to the parameters of coil L_l.

Now from **eq. 6**, with $\Delta f/f_{LO} = 1/28$, yields

$$L_l/n^2\ L_{coax} = \mu_l\ L_{la}/L_{coax} = 1/14 \quad \textbf{(8)}$$

This means that the ratio of L_l/n^2 or $\mu_l\ L_{la}$ to L_{coax} must be 1/14 so that the instrument will accommodate the grid-dip oscillator frequency spread of 500 kHz.

If your instrument is about the same size as mine, then L_{coax} will be about 56 nH.* Then **eq. 8** says that $L_l/n^2 = 4\ nH$ is required for the 500 kHz of the 20-meter band on the ham-band-only receiver. Or, saying the same thing, $\mu_l\ L_{la} = 4\ nH$ must also hold by virtue of the preceding discussion. Then the core dimensions, number of turns, and permeability are chosen to satisfy both of these two expressions – plus the fact that

*The values of 13 μH for L_l and 10 μH for L_s for this instrument were measured on a Boonton 260 Q meter. From the appendix (**eq. 1-12**), L_{coax} is 56 nH using these values and n^2 is 1000.

inductance must be compatible with the capacitor, C, in the grid-dip oscillator loop to resonate in the 20-meter band. The latter point is discussed below.

These relationships are not required to hold exactly. This is meant in the sense that the capacitor, C, can be used to partially compress or expand the frequency spread, Δf, to compensate for discrepancies between theory and practice. However, this is limited by the maximum and minimum values of capacitor C.

2. Another criterion must be satisfied; namely, the capacitor, C, must tune the primary inductance, L_l, in the frequency range of interest. The following example illustrates this point as well.

Selsyn-case cup permeability meter in place on a support fixture, which also supports the grid-dip oscillator link. Knob on the fixture varies the loop capacitor, C. The grid-dip oscillator is next to and slightly to the rear of the support fixture. The toroid primary coil is in the foreground. It is laced to Teflon-sheet padding on a Teflon base. Banana plugs allow coil to be plugged into bottom of cup. The coil consists of 100 turns of no. 24 AWG (0.5mm) wire on an Arnold Engineering type AM-12 core (core no. 10 in table 1).

Assume I want to convert my selsyn-case instrument to work with the grid-dip oscillator in the first 200 kHz of 20 meters with a Collins 75S-series receiver. The corresponding fractional frequency spread is $\Delta f/f_{LO} = 0.2\ MHz/14\ MHz = 1/70$. From **eq. 6**, my new $L_l/n^2L_{coax} = 1/35$. My present value is 3/13, as calculated from **eq. 6** with $\Delta f/f_{LO} = 1\ MHz/7\ MHz = 1/7$. This means I must reduce 3/13 by a factor of about 8 to obtain 1/35. I certainly am not going to change the cup dimensions; *i.e.*, L_{coax}, so I must alter the primary toroid coil to achieve this change. This means I must simultaneously satisfy both the fractional-frequency-spread relationship, $L_l/n^2L_{coax} = \mu_l L_{la}/L_{coax} = 1/35$, which characterizes the primary coil physical dimensions and the resonant-frequency relationship, $L_lC = 1/4\pi^2 \cdot (14\ MHz)^2$, which fixes the number of turns, n. When all of this is amalgamated, using the L_{coax} value of 56 nH, the implication is that the two relationships, $\mu_l\ h_l\ log_{10}\ (OD)_l/(ID)_l = 0.4$, and $n^2 = 80,000/C$ (capacitance in pF) hold at the same time.

Choosing $C = 50\ pF$ yields $n = 40$ for the number of turns required on the primary inductance. For the former relationship, if I were to use the same physical dimensions (for h, OD, and ID) for this new primary core as for the old core (core 10 in **table 1**), then its permeability works out to about 3. I then start looking through the catalogs for such a core.

In neither of the examples above would the nomogram require modification within the accuracies discussed earlier. This is true because (from **eq. 1-16** in the appendix) the instrument constant, k, as expressed as $k = (L_{coax} - L_l/n^2)/4$ (inductances now in nH) will still remain the same.

To illustrate, note that the present L_{coax} is 56 nH. With $L_l/n^2 = 13\ nH$, the corresponding k is about 11 from the above expression. Now compare **example 1**, where $L_l/n^2 = 4\ nH$, which then gives about 13 for the above k. The same holds for **example 2**, where $L_l/n^2 = 1.6\ nH$, yielding only a slightly different k of 13.5.

If the fractional frequency spread, $\Delta f/f_{LO}$, is always somewhat less than unity, then k won't change very much. This is also shown in the appendix. However, the other instrument constant, a, will change; but once computed, a is easily dealt with when going through the nomogram procedure.

future work

A number of areas of improvement in construction of the permeability meter suggest themselves. The selsyn case for the cup and the pop-up stems, whose dimensions were just right, were admittedly a coincidence. Many other copper or brass cup-like objects of this size are available. Hence, with a little extra scrutiny in this direction, you should be readily rewarded by a copper or brass cylinder of the right dimensions, even with top and bottom covers. An obvious approach, if facilities are available, is to build a cup from brass, with a bolted bottom and screw-on top and silver flashed circular fingers to receive the center conductor when the lid is on. This was done at the Bureau of Standards, as described in references 3 and 4.

A second area is that of achieving a more accurate frequency readout than can be obtained in the average communications receiver, by using an amateur-band receiver. Most receivers of this type have 3- or 4-place accuracy or better, but they span 500 kHz per band so that you cannot switch to the adjacent band segment to get the remainder of the frequency spread if the latter is greater than 500 kHz. To make the grid-dip oscillator measured frequency span 500 kHz or less, attention must be given to the physical dimensions of the instrument design.

Discussion is continued here to point out the matching of the cup dimensions to the grid-dip oscillator/receiver combination for optimum operation. For example, optimum operation criteria could include cup design specifications for increasing the frequency spread, Δf, of a continuous-coverage quality receiver to obtain maximum accuracy of grid-dip oscillator frequency data.

As a case in point, it is shown in **appendix 1**, last paragraph, using a criterion of uniform measurement accuracy from low to high permeabilities, that the optimum a^2 in the formula for the unknown inductance **(appendix 1 eq. 1-19)**, namely

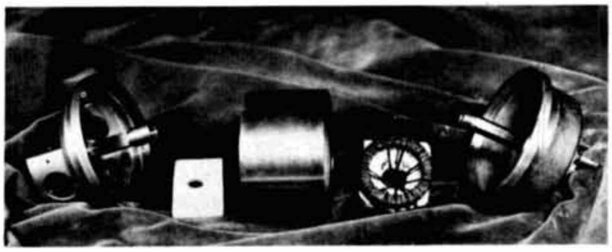
$$L_\mu = \left[L_{coax}\right]\left[\frac{a^2(1 - f^2)}{(f^2 - a^2)}\right] \qquad \textbf{(8)}$$

should equal 1/2. That is, $a^2 = (f_{LO}/f_{HI})^2 = ½$. This, then, implies that, for optimality in this sense, the following should hold:

1. The fractional-frequency spread, $\Delta f/f_{LO} = \sqrt{2} - 1 = 0.414$, or $f_{HI} = \sqrt{2}\, f_{LO}$, so that the optimal upper grid-dip oscillator frequency should be 41.4 per cent greater than the lower grid-dip oscillator frequency. In my case, if I insist on $f_{LO} = 7\ MHz$, then $f_{HI} = 10\ MHz$.

For an amateur-band receiver (or transceiver), with a vfo frequency spread of $\Delta f = 500\ kHz$, the corresponding $f_{LO} = \Delta f/0.414 = 1.2\ MHz$, which is smack in the middle of the broadcast band. So the next best thing is to use 80 meters (3.5-4.0 MHz), or 160 meters, in the off chance that 500 kHz is available on your receiver.

2. From the above, the instrument design relationship is now such that $L_l/n^2\ L_{coax} = 1 - a^2 = ½$, as well. This also means that $\mu_l L_{la} = ½ L_{coax}$; *i.e.*, the primary toroid geometrical or free-space inductance should equal half that of the secondary or cup geometrical inductance.



Exploded view of the selsyn-case cup premeability meter. Top end bell is at left with bushing screwed onto upper half of pop-up stem center conductor. Next is the Teflon-block table to support unknown toroid core. In the center is a thin brass shell, which press fits into the top and bottom end bells. Next is a primary coil of 31.5 turns of no. 24 AWG (0.5mm) wire on an AM-12 core (core no. 10 in table 1). At right is the bottom end bell integral with the lower half of the center conductor. Banana plugs and coax connector protrude to the right of cup bottom end bell.

A third area of improvement is the construction of a more efficient nomograph, perhaps one with fewer scales. Or you could resort to graphical plotting procedures, which would amount to the same complexity.

A fourth area of improvement might be a direct-reading permeability meter. A tertiary winding could be introduced within the cup beneath the table. This winding would surround the center conductor to sample the cup magnetic-field strength. The tertiary winding terminals could easily be brought out through the cup wall. The internal cup field would vary in strength with cores of varying permeability. Diode rectification of the corresponding probe current could be used to deflect a permeability-calibrated dc meter. However, for sufficient meter current, the primary would probably need to be excited directly by a signal generator instead of the grid-dip oscillator. The introduction of a signal generator puts this improvement at least into the semiprofessional class. This certainly would be the case if a digital permeability readout were incorporated. Professionalism as such is not frowned upon; things just get more expensive.

A final note is that the permeability of an unknown core can be found even if a coil of relatively few turns is wound on it. Of course, the coil leads are not connected when it is placed in the instrument.

Most of the above effort is concentrated in the hf range; for example, the selsyn-case instrument/grid-dip oscillator dips occurred in the 7-8 MHz range. To investigate the detailed behavior of cores from, say, 30 MHz to vhf many permeability meters would probably have to be built. They would range from the cup size shown here to thimble sizes for vhf and uhf applications.

acknowledgements

I wish to express my appreciation for the calibration cores of known permeability that were graciously supplied to me by Bill Amidon of Amidon Associates, in North Hollywood, California. Further, without the encouragement and substantial help of Mssrs. Eulalio Vela and H.A. Dickerson, this effort would have been impossible.

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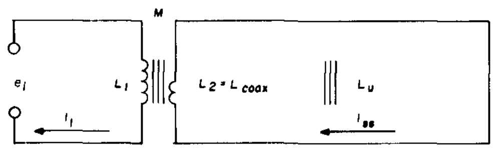
appendix

permeability-meter circuit analysis

The following analysis is straightforward ac circuit theory and derives the basic formulas contained in the text. It centers around the three states discussed previously. Subscripts l denote the primary coil parameters in all three states. Subscripts s and ss designate the secondary winding parameters in states **B** and **C** respectively. Primary- and secondary-loop equations are written for each state as it is discussed. It is assumed that all resistance in the loops is negligible compared with the corresponding inductive reactances so that no resistances will appear in the loop equations. This assumption, besides the simplification in the loop equations it affords, is valid for the inherent high-Q circuitry of the permeability meter itself. When consideration of the unknown toroid dissipation factor is of interest, then these resistances must be included in the analysis. This is done in reference 3.

Consider state **C** (lid on; *i.e.* shorted secondary, and unknown toroid inside) as sketched below. The primary and secondary loop equations are:



STATE C EQUIVALENT CIRCUIT

$e_l = i\omega L_l I_l - i\omega M I_{ss}$ for the primary loop **(1-1)**

$0 = i\omega(L_2 + L_u)I_{ss} - i\omega M I_l$ for the secondary loop **(1-2)**

e_l = applied voltage of the grid-dip oscillator input

I_l and I_{ss} = corresponding loop currents

M = mutual inductance between primary and secondary windings

L_u = unknown toroid inductance;

$L_u = (\mu - 1)\ L_{ua}$, where

$L_{ua} = 0.004046\ h\ log_{10}\ (OD/ID)\ in\ \mu H$

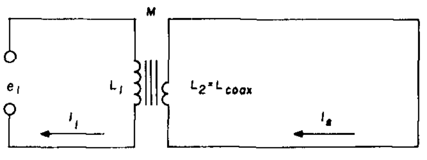
$\omega = 2\pi\ f$, where f is the grid-dip oscillator input frequency to the primary toroid winding.

Estimating I_{ss} between **eqs. 1-1** and **1-2** gives for state **C** input impedance, Z_{ss}, and input inductance, L_{ss}:

$Z_{ss} = e_l/I_l = i\omega[L_l - M^2/(L_2 + L_u)]$; i.e.,

$$L_{ss} = L_l - [M^2/(L_2 + L_u)] \quad \textbf{(1-3)}$$

For state **B** (lid on, secondary shorted, with no toroidal inside), the loop equations are



STATE B EQUIVALENT CIRCUIT

$e_l = i\omega L_l I_l - i\omega M I_s$ for the primary loop **(1-4)**

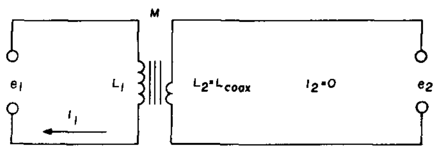
$0 = i\omega L_2 I_s - i\omega M I_l$ for the secondary loop **(1-5)**

where the symbols follow from the above explanation. Elimination of I_s between **eqs. 1-4** and **1-5** gives for state **B** input impedance, Z_s, and input inductance, L_s:

$Z_s = e_l/I_l = i\omega(L_l - M^2/L_2)$; *i.e.*,

$$L_s = L_l - M^2/L_2 \quad \textbf{(1-6)}$$

For state **A** (lid off, secondary open, no toroid inside), where e_2 is the open-circuit secondary voltage when grid-dip oscillator voltage e_l is applied, the loop equations are



STATE A EQUIVALENT CIRCUIT

$e_l = i\omega L_l/I_l$ for the primary loop **(1-7)**

$e_2 = -i\omega M I_l$ for the (open) secondary, and

$Z_l = e_l/I_l = i\omega L_l$ **(1-8)**

If the leakage inductance is negligible, then dividing **eq. 1-7** by **eq. 1-8** yields

$e_l/e_2 = n = -L_l/M$, where n is the turns ratio, or the number of primary turns for our one turn secondary. **(1-9)**

Substituting into the right-hand portions of **eqs. 1-3** and **1-6**, from **eq. 1-9** to eliminate M and L_2 results in (a) an equation for the unknown inductance, L_u, namely,

$$L_u = \left[\frac{L_l^2}{n^2(L_l - L_s)}\right]\left[\frac{(L_{ss} - L_s)}{(L_l - L_{ss})}\right] \quad \textbf{(1-10)}$$

or from above, with the unknown inductance usually written in terms of its free-space inductance as $L_u = (u - 1)L_{ua}$, then equivalently,

$$\mu - 1 = \left[\frac{1}{0.004046 h log_{10}(OD/ID)}\right]\left[\frac{L_l^2}{n^2(L_l - L_s)}\right]\left[\frac{(L_{ss} - L_s)}{(L_l - L_{ss})}\right] \quad \textbf{(1-11)}$$

and (b), an equation for the secondary inductance,

$L_2 = L_{coax}$; $L_{coax} = L_l^2/n^2(L_l - L_s)$ **(1-12)**

Notice from **eqs. 1-10** and **1-12** that now $L_u = L_{coax}\ (L_{ss} - L_s)/(L_l - L_{ss})$.

Generally, the grid-dip oscillator resonant dip frequency, f_i^2, or $\omega_i^2 = 1/L_iC$, where the inductance, L_i can correspond to $i = 1, 2, s, ss$, and C is the capacitor in the grid-dip oscillator loop. Again, with **eq. 1-10**, using **eq. 1-12**, yields for the unknown inductance, L_u (recalling that $f_s = f_{HI}$, $f_l = f_{LO}$ and $f_{ss} = f_{cor}$ from the text):

$L_u = (L_{coax} - L_l/n^2)(f_{HI}^2 - f_{cor}^2)/(f_{cor}^2 - f_{LO}^2)$ **(1-13)**

With $L_u = (\mu - 1)L_{ua}$, from **eq. 1-13** where $f = f_{cor}/f_{HI}$, and $a = f_{LO}/f_{HI}$, gives

$$\mu - 1 = \frac{k}{h log_{10} OD/ID}\ (1 - f^2)/(f^2 - a^2) \quad \textbf{(1-14)}$$

where the instrument constant, k, is written as

$k = (L_{coax} - L_l/n^2)/0.004046$ (inductances in μH) **(1-15)**

$k = (L_{coax} - L_l/n^2)/4.046$, or $k = (L_{coax} - L_l/n^2)/4$ **(1-16)**

where these inductances are now in nH.

To obtain **eq. 6** in the text, rewrite **eq. 1-12** as

$$L_{coax} = L_l/n^2 \cdot (1 - L_s/L_l) \quad \textbf{(1-17)}$$

Now $L_s/L_l = f_l^2/f_s^2 = f_{LO}^2/f_{HI}^2$, when using the resonating capacitor, C, in the loop. If these frequency ratios are inserted into **eq. 1-17**, where the fractional frequency spread is $\Delta f/f_{LO} = (f_{HI} - f_{LO})f_{LO}$, then

$$L_l/n^2 = (1 - \frac{1}{(1 + \Delta f/f_{LO})^2})\ L_{coax} \quad \textbf{(1-18)}$$

which is **eq. 7** in the text.

To show that the instrument constant, k, will change little, as long as the fractional frequency spread, $\Delta f/f_{LO}$, is small compared to one, eliminate L_l/n^2 between **eq. 6** and **eq. 1-16** above. This gives $k = L_{coax}/4(1 + \Delta f/f_{LO})^2$. It is therefore seen that k will remain essentially constant for small changes in $\Delta f/f_{LO}$ as long as the latter is small compared to one. Again, this then means that the permeability scale on line 11 of the nomogram of **fig. 4** does not have to be adjusted for small changes in the fractional frequency spread from instrument to instrument.

Eq. 1-13 for the unknown inductance can be rewritten, using **eq. 1-12** and noting that $L_sL_l = a^2$, as

$L_u = [L_{coax}]\ [a^2\ (1 - f^2)/(f^2 - a^2)]$ **(1-19)**

or,

$L_u + a^2L_{coax} = L_{coax}a^2(1 - a^2)/(f^2 - a^2)$ **(1-20)**

which is a hyperbolic relationship between the unknown inductance, L_u, and f^2, with asymptotes at $f = a$ and $L_u = -a^2L_{coax}$. Since $0<a<f<1$, simply using as a criterion uniform measurement accuracy for the lowest to the highest permeabilities, it is seen that $[L_{coax}]\ [a^2(1 - a^2)]$ should be maximized to make the above hyperbola uniformly as "fat" as possible. Since L_{coax} is fixed by the physical cup dimensions, then for any given L_{coax} it is apparent that the maximum of $[L_{coax}]\ [a^2(1 - a^2)]$ will occur at $a^2 = ½$.

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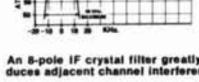
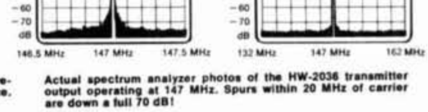


An 8-pole IF crystal filter greatly reduces adjacent channel interference.



Actual spectrum analyzer photos of the HW-2036 transmitter output operating at 147 MHz. Spurs within 20 MHz of carrier are down a full 70 dB!

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how many signals does a receiver see?

While listening on the various ham bands, I am always amazed at signal reports where stations supposedly receive reports of 50 or 60 dB over S9. Since I have always felt that this was an incorrect statement, I went through the difficulty of actual band scanning with a calibrated instrument – in this case a Hewlett-Packard Spectrum Analyzer. A special Rohde & Schwarz omnidirectional active receiving antenna was used. Its design is based upon principles developed by Professor Meinke (Technical University of Munich). With very small mechanical dimensions, this antenna provides the equivalent of 6 dB omnidirectional gain, up to 30 MHz. During a recent 24-hour monitoring period, pictures were taken of the analyzer display approximately every three hours.

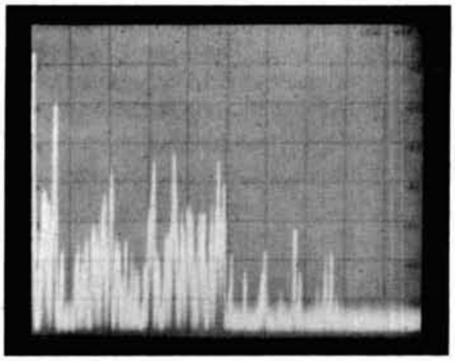
The HP spectrum analyzer was calibrated so that 0 dB on the y-axis was equal to 0 dBm (224 mV) and the scan width was 2 MHz per division, with 20 MHz placed on the extreme right edge of the graticule. The following pictures show the results. The strongest signal (−20 dBm, 22.4 mV into 50 ohms) was the local a-m broadcasting station around 1.6 MHz. This station was located about 10 miles from the receiving site. The most interesting result of this test was that the 15-MHz broadcasting stations were almost as loud as the 80-meter stations, though they were never stronger than −30 dBm. Using the definition that S9 = 50 μV, these signals were approximately 40 dB over S9. However, no signals of 50 or 60 above S9 could be seen.

These pictures are interesting from another point of view, *i.e.*, they clearly show the change in the maximum usable frequency (MUF). It was most interesting to see that the 15-MHz region was open until very early in the morning and then suddenly the signals disappeared. In addition, the dead zone for the low frequencies can be seen. About midday the propagation on the lower frequencies was extremely poor and then improved later in the day.

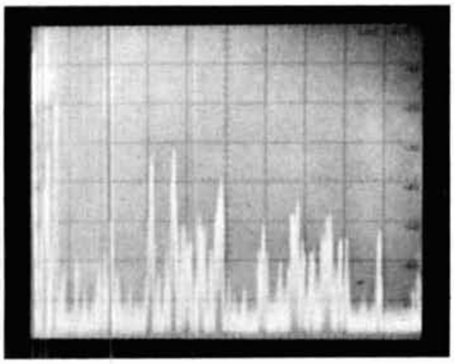
ham radio

By Ulrich L. Rohde, DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

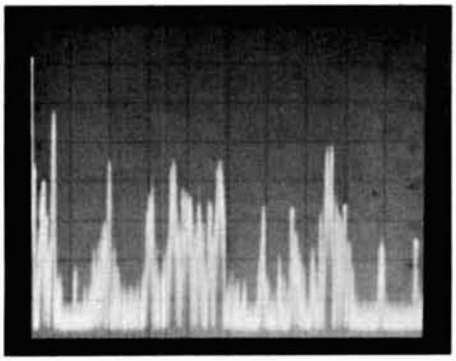
The following pictures were taken from the display of an H-P-141T Spectrum Analyzer system, 8553B rf section and 8552B mainframe. The antenna was a special omnidirectional model that provided 6 dB gain.



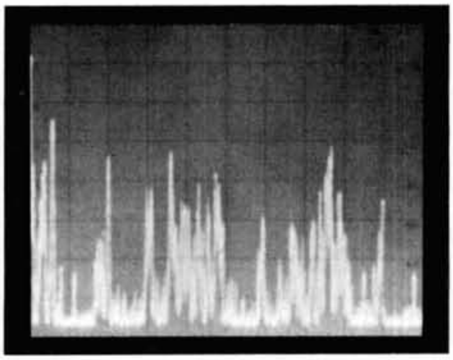
1. 8:00 AM



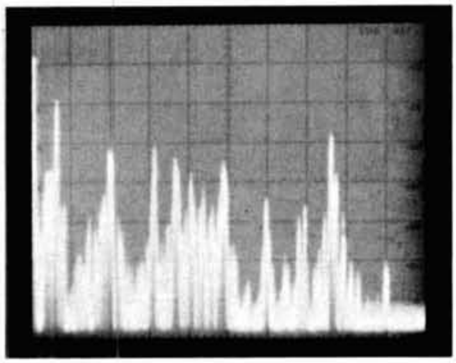
2. 11:30 AM



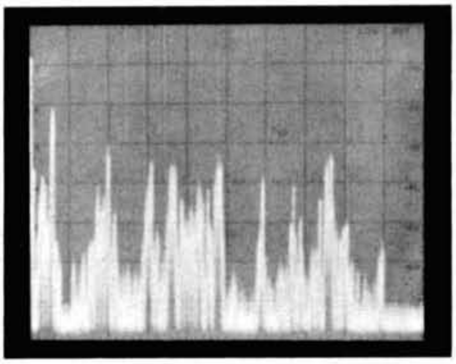
3. 3:00 PM



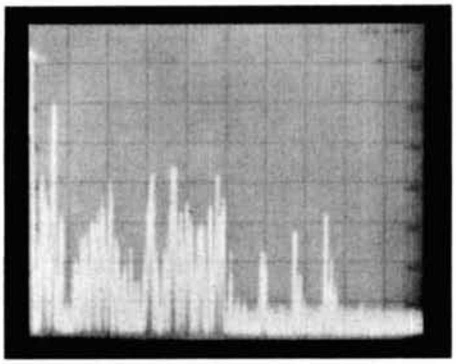
4. 6:00 PM



5. 10:00 PM



6. 1:00 AM



7. 4:00 AM

short vertical antenna for 7 MHz

With a good ground system, this antenna will give a good account of itself for those with limited space

Polar radiation patterns in the vertical plane for a low horizontal antenna and a short vertical antenna are shown in **fig. 1**. These patterns are typical for horizontal antennas less than about 0.3 wavelength above the ground and for vertical antennas less than about 0.6 wavelength high. For working beyond a range of a few hundred miles on 7 MHz, only the energy radiated at angles less than about 45 degrees is of much value. **Fig. 1** shows that a low horizontal antenna radiates most of its energy at very high angles, with the maximum being straight up. A low horizontal antenna on 7 MHz is one that is less than about 40 feet (12m) high. Now look at the short vertical pattern. Its radiation straight up is zero and its maximum radiation is at a very low angle, approaching zero. In practice, ground losses change the pattern so that little energy is radiated parallel to the ground. A short vertical antenna for 7 MHz is one that is less than about 80 feet (24m) high.

I erected my antenna in the center of my roof at the peak and ran the radials for the ground plane from that point to the roof edge. Thus, the radials are of different lengths and are not at right angles to the antenna. Instead they slope down. This was the easiest and most logical way to install the radials, and the antenna works well.

All this activity occurred in the fall of the year and i wanted to try the idea quickly, so I made use of the materials I had on hand. For the vertical radiator I used a 10-foot-long (3m) piece of thin-wall electrical conduit with an 8-foot (2.4m) whip attached to the top. I mounted the vertical portion of the antenna on a 5-foot (1.5m) tripod of the type used for television antennas. I made two insulators of plastic to mount the electrical conduit to the tripod. The bottom end of the conduit is about a foot (0.3m) above the roof, and a brass plate with the radials attached is on the roof under the conduit. The two coils of the matching network (discussed later) are mounted to the brass plate with their axes at right angles.

Two things are essential to the success of an antenna of this type. One is an impedance-matching network to allow the efficient transfer of energy from the transmission line to the radiator. The other is a proper ground system.

antenna impedance

An impedance-matching network is necessary because the impedance of a short vertical antenna is not a good match for most transmission lines. Typically a short vertical antenna will have an impedance consisting of a low resistance and a high capacitive reactance. A conjugate match can be obtained by designing a network to match the transmission line to the resistive part of the antenna impedance and adding an impedance between

By H. H. Hunter, W8TYX, 1106 Carolyn Avenue, Columbus, Ohio 43224

the network and the antenna equal to the antenna reactance but of the opposite sign. Thus for my short vertical, the matching network had to match the transmission line to the few ohms impedance of the antenna. I also included a series inductance having a reactance equal to that of the antenna capacitive reactance. This series inductance compensates for the effect of antenna capacitance.

It would have been desireable to erect the antenna and then measure its impedance at the operating frequency. However, I did not have an impedance bridge. La Port's *Radio Antenna Engineering* contains a good discussion of vertical antennas. A portion of the book is directed to medium-frequency, vertical broadcast antennas, but the data presented seem to work well at the higher frequencies. Curves are presented in the book that show the impedance (resistance and reactance) of vertical antennas as a function of antenna height and antenna length-to-diameter ratio. The diameter of my antenna is not constant since the 8-foot (2.4m) whip is tapered to about 0.125 inch (3mm) diameter at the top. I assumed an average diameter for the entire antenna of 0.5 inch (12.5mm). Thus, the ratio of length to diameter is 18 x 12/0.5 or 432. The antenna height in degrees, where 360 degrees is one wavelength, is 46.3, which is just a little over 1/8 wavelength. Then, from the figures in reference 1, the antenna base resistance will be about 5 ohms and the reactance will be about −j320 ohms. The impedance-matching network thus had to match the 50 ohms of the RG-8/U coax cable I used to 5 ohms. Also, the network had to provide a reactance of +j320 ohms to compensate for the antenna capacitive reactance.

Table 1 lists the approximate resistance and reactance for several lengths of vertical antennas. This data was obtained from reference 1. In all data shown here, an operating frequency of 7.037 MHz is assumed, but the data should be useful over the entire 40-meter band.

matching network

I used an L network for the impedance-matching system. These networks are easy to design and work well. I used the equations in reference 2. **Fig. 2** shows an L network. The network can work either way. That is, it can transform from a high to a low impedance (the way

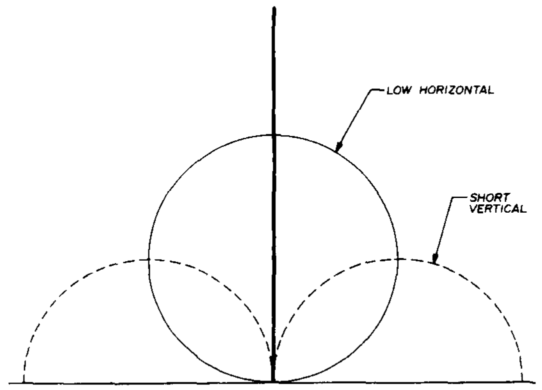


fig. 1. Typical polar radiation patterns in the vertical plane for a low horizontal and short vertical antenna.

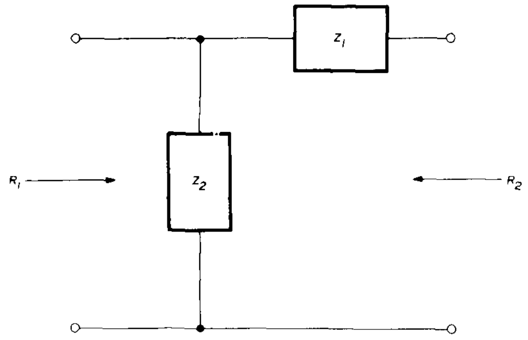


fig. 2. L network.

I wanted to go) or the opposite. The network consists of two reactances with opposite signs. That is, one must be capacitive and the other inductive. The values of these reactances are given by:

$$Z1 = \pm j\sqrt{R2(R1 - R2)} \qquad (1)$$

$$Z2 = \pm jR1\sqrt{\frac{R2}{R1 - R2}} \qquad (2)$$

where $Z1$ = reactance of the series arm (ohms)
$Z2$ = reactance of the shunt arm (ohms)
$R1$ = resistance (impedance) of the transmission line; 50 ohms for RG-8
$R2$ = radiation resistance of the antenna; 5 ohms in the case of my 18-foot (5.5m) antenna.

In the equations above, $R1$ must always be larger than $R2$. In designing the network, you can use either the two top signs or the two bottom signs of **eqs. 1** and **2**. That is, the series arm can be an inductance and the shunt arm a capacitance, or vice versa. I chose to use an inductance for the shunt arm and a capacitance for the series arm. The reason for doing this is described later.

To match a 50-ohm transmission line to the 5-ohm radiation resistance of the antenna, substitution in **eqs. 1** and **2** gives

$$Z1 = -j\sqrt{R2(R1 - R2)}$$
$$= -j\sqrt{5(50 - 5)} = -j15 \qquad (3)$$

$$Z2 = +j\,R1\sqrt{\frac{R2}{R1 - R2}}$$
$$= +j50\sqrt{\frac{5}{50 - 5}} = +j16.7 \qquad (4)$$

To cancel the effect of the antenna capacitive reactance I added an inductive reactance having the same magnitude, 320 ohms, between the L network and the antenna base.

Fig. 3 shows the complete matching network. The capacitive reactance of $Z1$ is in series with the added inductive reactance of $+j320$ ohms. Thus the effective series arm is $-j15$ and $+j320$, or $+j305$ ohms. By using the network as shown, the vertical radiator is connected electrically to the ground plane. Thus, if the ground plane is connected to an earth ground the entire antenna system can be grounded.

The value of the two inductances in the complete matching network can be calculated from

$$X_L = 2\pi f L \qquad (5)$$

where

X_L = reactance of the inductance (ohms)
f = frequency of operation (Hz)
L = inductance (H)

The known values of X_L and f are substituted into **eq. 5** and the equation then is solved for L. For the shunt arm, having a value of 16.7 ohms and a frequency of 7.037 MHz or 7.037×10^6 Hz, **eq. 5** becomes

$$16.7 = 2\pi \cdot 7.037 \cdot 10^6 \cdot L1 \qquad (6)$$

and

$$L1 = 0.378 \cdot 10^{-6}\ H = 0.378\ \mu H \qquad (7)$$

Similarly for the series arm, having a value of 305 ohms,

$$305 = 2\pi \cdot 7.037 \cdot 10^6 \cdot L2 \qquad (8)$$

and

$$L2 = 6.90 \cdot 10^{-6}\ H = 6.90\ \mu H \qquad (9)$$

The number of turns required for the two inductances can be calculated if coil length and diameter are assumed. A formula given in the *ARRL Handbook* is:

$$L = \frac{a^2 n^2}{9a + 10b} \qquad (10)$$

where
L = inductance (μH)
a = coil radius (in.)*
b = coil length (in.)*
n = number of turns in coil

Rearranging **eq. 10**,

$$n = \left[\frac{L(9a + 10b)}{a}\right]^{1/2}$$

The shunt coil is 0.378 μH and requires 2.7 turns if the diameter and length are 3 inches (76.5mm). Similarly, for the series coil (6.9 μH), 18 turns are required if the diameter is 3 inches (76.5mm) and the length is 9 inches (230mm).

Even with modest power the current in these coils can be fairly large, so the coils should be constructed accordingly. I used 1/4-inch (6.5mm) copper tubing for the shunt coil and 1/8-inch (3.0mm) copper tubing for the series coil. The coils should be mounted at right angles to each other to minimize inductive coupling.

ground system

The ground system used with a short vertical antenna is a big factor in the performance of the antenna. The ground system consists of a number of radials, which are wires that extend out from the base of the antenna and in a direction more or less normal to the antenna. Imagine a wheel laid flat on the ground or on a roof with the axle sticking straight up and the wheel spokes extending outward from the axle. The axle represents the radiating part of the antenna and the spokes represent the radials.

* If a and b are in mm, the factors 9 and 10 in the denominator of **eq. 10** are 229.5 and 255 respectively. **Editor**

table 1. Approximate resistance and reactance for several lengths of vertical antenna at 7 MHz.

length feet	(m)	diameter inches	(mm)	length/diameter	length degrees	resistance (ohms)	reactance (ohms)
15	(4.6)	0.5	(12.5)	360	38.6	4.5	−j450
15	(4.6)	0.75	(19)	240	38.6	3.5	−j400
15	(4.6)	1.0	(25.5)	180	38.6	2.9	−j380
18	(5.5)	0.5	(12.5)	432	46.3	5	−j320
18	(5.5)	0.75	(19)	288	46.3	4	−j300
18	(5.5)	1.0	(25.5)	217	46.3	3.5	−j280
20	(6)	0.5	(12.5)	480	51.3	8	−j270
20	(6)	0.75	(19)	320	51.5	5	−j250
20	(6)	1.0	(25.5)	240	51.5	4.5	−j230
20	(6)	1.25	(32)	192	51.5	4	−j200
25	(7.6)	0.5	(12.5)	600	64.3	14	−j160
25	(7.6)	0.75	(19)	400	64.3	12	−j150
25	(7.6)	1.0	(25.5)	300	64.3	11	−j140
25	(7.6)	1.25	(32)	240	64.3	10	−j130
30	(9)	0.5	(12.5)	720	77.2	22	−j55
30	(9)	0.75	(19)	480	77.2	22	−j50
30	(9)	1.0	(25.5)	360	77.2	22	−j45
30	(9)	1.25	(32)	288	77.2	22	−j41
30	(9)	1.5	(38)	240	77.2	22	−j40

Most a-m broadcast stations use vertical antennas and a ground system consisting of 120 radial wires spaced every 3 degrees from the base of the antenna. Radial length seldom exceeds a half wavelength and often is less.

Reference 1 shows the field strength produced by a vertical radiator as a function of the height of the antenna and the number and length of the radials in the ground system. Two conclusions are apparent from the information presented: (1) use as many radials as you possibly can, and (2) make the radials as long as possible. With regard to the number of radials, broadcast stations use 120, which gives an efficiency close to the theoretical maximum. There probably are not too many amateurs who will use 120 radials, but use as many as you possibly can.

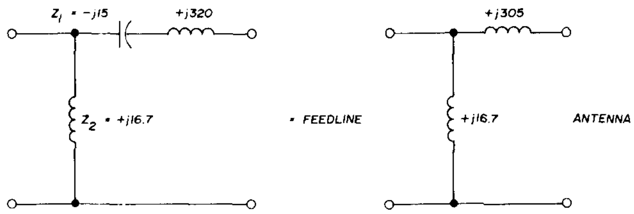


fig. 3. Complete matching network for the short vertical antenna.

With regard to the length of the radials, there is little to be gained by making the length greater than a half wavelength. For many it's not possible to even approach a half wavelength for the radials; there simply is not that much room on a city lot or on the average rooftop. If you can't make the radials that long, you can compensate somewhat for their shortness by using more of them. Thus again, the rule is to use as many radials as you possibly can and make them as long as possible. The shorter the antenna, the more important this is.

The ground system for a broadcast antenna is usually buried a few inches to get it out of the way, and the wires are at right angles to the vertical radiator. With my roof-mounted vertical antenna, the radials are on the roof and slope away from the vertical radiator instead of being at right angles. However, this antenna performs well, so the sloping radials apparently do not effect performance adversely.

In my system I mounted an 8 x 10 inch (20x25cm) brass plate under the lower end of the vertical radiator. I drilled a series of holes around the edge of this plate and tapped them for 6-32 (M3.5) machine screws. When I installed the radial wires I attached each one under the head of a screw mounted in one of the tapped holes. After I had attached all of the radials to the plate I soldered the wires and the screws to the base plate.

My house is single story about 29 x 41 feet (9x12m) with a hip roof. I mounted the antenna in the center of the roof at the peak. I used 20 radials of no. 24 (0.5mm) bare copper wire and spaced them equally. My longest radial is about 25 feet (7.6m); the shortest is about 15 feet (4.6m).

adjustment

Adjustment of the matching-network series coil is necessary. As explained earlier, the purpose of this coil is to compensate for the antenna capacitive reactance. The information about the capacitive reactance of the antenna is approximate rather than exact. Thus it's necessary to either measure the reactance of both the antenna and series coil and make them equal or adjust the inductance after it is in place and make it "tune out" the antenna capacitance. I did the latter.

With the antenna, ground system and feed line installed, I used a vswr meter at the transmitter and adjusted the series-coil inductance for the lowest swr at the low end of the 7-MHz band. I obtained a swr of 1.5:1. I adjusted the coil by spreading or compressing the length of the coil. This adjustment is critical. It should be possible to perform this adjustment with a noise bridge or with a rf-current indicator connected at the junction of the series coil and the vertical-antenna base. For the noise bridge, adjust the series coil so the antenna is resonant at the desired frequency. For the rf-current meter, adjust the coil for maximum current.

results

This short vertical antenna has given a good account of itself. If your antenna farm is smaller than you'd like, or if you can't erect a 70 footer (20m) consider a vertical, even a short one. Consider also installing the antenna on top of the house. Installing the antenna on the roof reduces clutter in the back yard and allows the antenna to "see" the horizon much better than at ground level.

references

1. E. A. La Port, *Radio Antenna Engineering*, 1952 edition, McGraw-Hill Book Company, New York.
2. F. E. Terman, *Electronic and Radio Engineering*, 4th edition, McGraw-Hill Book Company, New York, page 115.

ham radio

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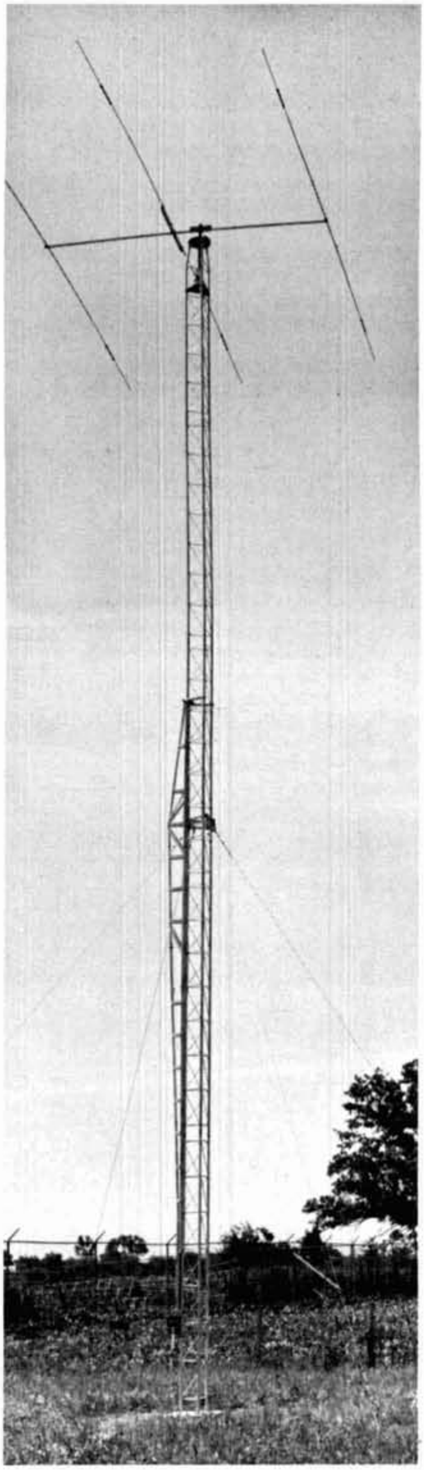
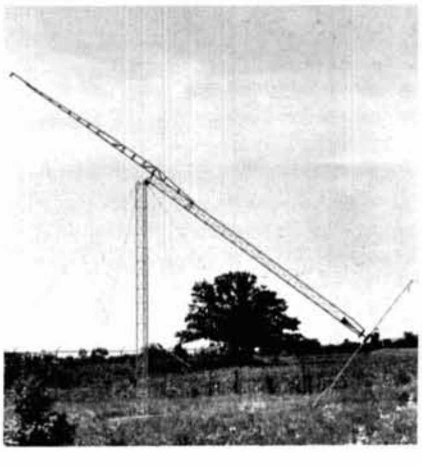




More Details? CHECK—OFF Page 126

video modulated four-tube amplifier for 1270-MHz television

Construction details of a four-tube ATV amplifier which provides continuous-power output of 120 watts at 1270 MHz

The need to develop adequate power on fast-scan television, at 1270 MHz, came about in 1974 when the Chicago area ATV group decided to look to bands other than 420-450 MHz for A5 operation. Results with low power and small antennas proved to be very useful and it was felt that higher power and better antennas would help to eliminate some problems with snow and improve the range.

After consulting with George, W9WCD, and reading the *ham radio* article by WB6IOM, I decided to try a cavity amplifier using four 2C39A tubes in a similar ring cavity. The only requirement not covered by the WB6IOM amplifier was the plate-circuit tuning range. I decided that it would be desirable to tune the entire 1215 to 1300 MHz range because of the great deal of radar interference. This tuning range would also allow us to stay away from any interference resulting from harmonics of 439 and 427 MHz ATV transmissions. It was found that the radar and other undesired transmissions were at a minimum at 1266 MHz and full ATV duplex was very easily accomplished.

1270 MHz amplifier

The complete four-tube amplifier is mounted on a 17x13x3-inch (43.2x33.0x7.6cm) chassis, complete with modulator, modulator power supply, air cooling system, video detector, and filament transformer. The amplifier is operated with the grids at dc and rf ground; bias is supplied from the direct-coupled modulator. For normal video-modulated operation, drive requirements are about 15 watts, at a plate current of 400 mA and a plate voltage of 800 volts. Normal power output from the amplifier is about 120 watts average, as measured on a Bird model 43 wattmeter. Since the amplifier is used for continuous operation, it is suggested that adequate cooling and moderate plate voltage and current be used. It is also important that the power supplies are rated for continuous service.

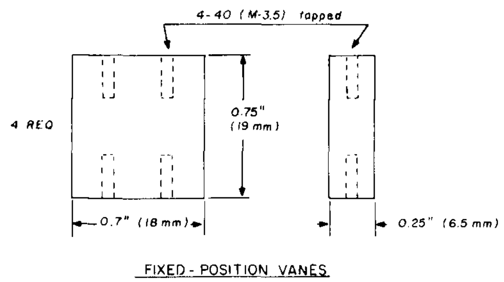


fig. 1. Anode cavity fixed-position vanes.

By Ronald Stefanskie, W9ZIH, 8950 South Maple Lane, Hickory Hills, Illinois 60457

The amplifier was built with the aid of a lathe and drill press. It is built entirely of brass and bronze that was obtained mainly from the local junk yards. The circular brass flat material was initially rough cut from 0.08-inch (20mm) sheet stock using a power sabre saw and then cut to size on the lathe with the use of a face

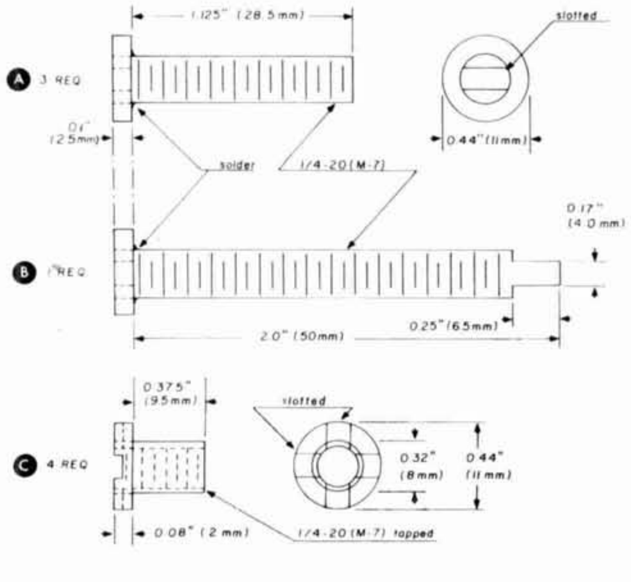


fig. 2. Plate-tuning disks and chucks. The three short tuning disks are slotted to permit adjustment during initial tune-up of the amplifier.

plate. The large clearance hole on the circular disks was also cut on the lathe.

anode cavity

The four 2C39As are placed around a 2.57 inch (6.5mm) diameter circle within a cavity of 4.741 inch (12cm) OD and a height of 0.74 inch (1.9cm). The natural resonant frequency of the cavity is below the 1270-MHz band but is raised by the use of four 0.69x0.74x0.25-inch (17.5x19.0x6.5mm) fixed-position

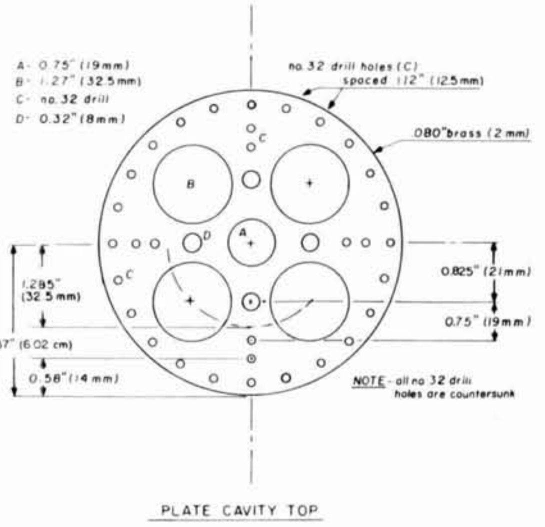
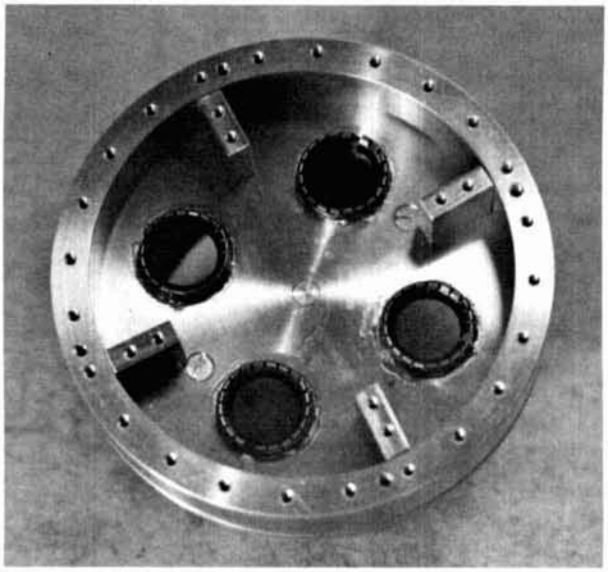


fig. 3. Top plate for the anode cavity.



Looking into the anode cavity from the top. The four fixed-position vanes are installed.

brass vanes (fig. 1). The vanes are place inside and bolted to the top and bottom walls through countersunk, tapped 4-40 (M3.5) holes.

Four plate-tuning disks, placed in a 1.65-inch (4.2cm) diameter circle, and centered between each tube, adjust the plate-circuit resonant frequency. Only one tuning disk is required for external plate-tuning adjustment. The others are fixed and placed in approximately the same position within the cavity. The four plate-tuning disks (see fig. 2) are machined from brass stock and ¼-20 M7 threaded brass rod. The 0.44-inch (1.1cm) disk is drilled in the center, large enough to force fit a small filed-down portion of the threaded rod. The rod is then soft soldered to the rear of the disk.

A 0.357x0.44-inch (9.5x11.0mm) chuck is machined from brass stock and then drilled and tapped in the center for the ¼-20 (M7) shank of the plate tuning disks (fig. 2C). A fine-bladed saw is used to slot the walls of the chuck so they can be compressed and adjusted for a snug fit. Four holes (0.32 inch, [8mm]) are drilled in the anode cavity top plate (fig. 3) and each chuck is inserted and soft soldered in place with the aid of a propane torch.

The output-coupling probe (fig. 4) is built by fashioning machined brass tubing, rod, flat stock, and Teflon rod into a length of concentric line (coax). The shield is made by inserting and soldering the ½-inch (12.5mm) OD brass tubing into the cable end of a UG-23B/U type-N connector; the end of the connector must be enlarged to accept the tubing. The center conductor of the probe is made from a small piece of 0.123-inch (3mm) diameter brass rod. One end is attached to the center pin of the connector while the other is cut for 6-32 (M3.5) threads. The Teflon rod is machined to fit inside the brass tubing. A hole is drilled through the center of the Teflon rod to accept the brass rod. The probe's disk (0.48 diameter, x 0.1 inch thick [12x2.5mm]) is drilled and tapped for 6-32 (M3.5) threads. The disc is then screwed onto the probe assembly and held by a stop nut. A 0.975x0.75-inch (25x19mm) chuck (**fig. 5**) for the output-coupling probe was machined from brass stock and slotted for a snug fit to the output-coupling probe. The chuck is also soldered to the underside of the center hole in the top plate of the anode cavity.

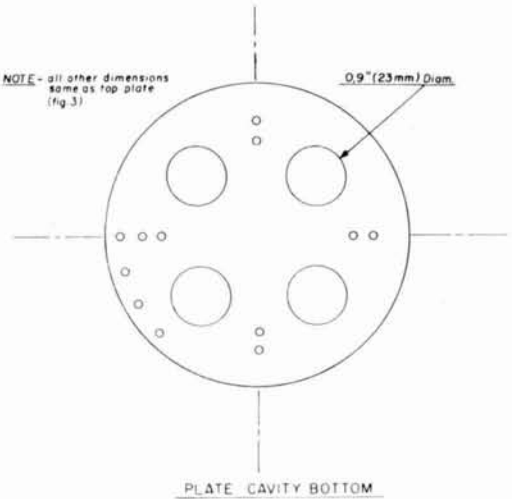
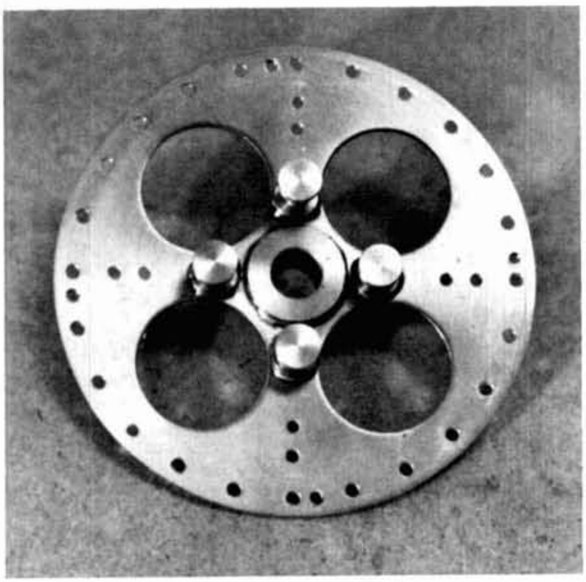


fig. 7. Bottom plate for the anode cavity.

The plate connector and bypass disk (**fig. 6**) is 0.08-inch (20mm) thick by 4.742-inch (12cm) OD with appropriate clearance holes for the tubes, plate-tuning chuck, and output-coupling chuck. The finger stock for the anode connector (Instrument Specialty Company 97-70) is soldered in the four tube holes on the disk. A piece of 0.01-inch (0.5mm) Teflon sheet stock is used as the insulation for the plate-bypass capacitor. Clearance holes are cut out for the four tubes and the chucks.

Homemade Teflon shoulder insulators were used to



Anode cavity top plate showing the four tuning disks. The holes have been countersunk on the other side to permit the Teflon insulator to lie flush against the plate.

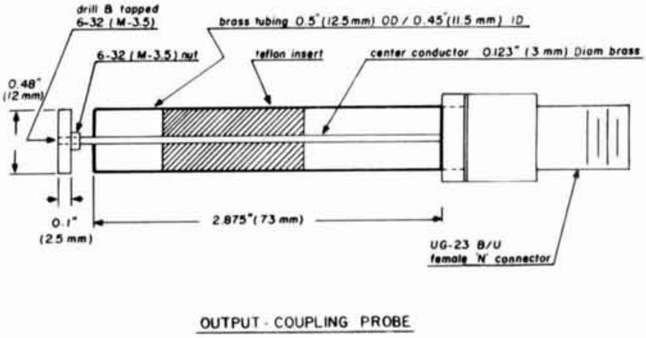


fig. 4. Output-coupling probe.

mount the plate connector and bypass disk to the top of the plate cavity. The anode cavity top and bottom plates (**fig. 7**) are bolted to the side wall portion by 4-40 (M3) screws. The countersunk holes are spaced every ½ inch (12.5mm) on the top plate and every 5/8 inch (16mm)

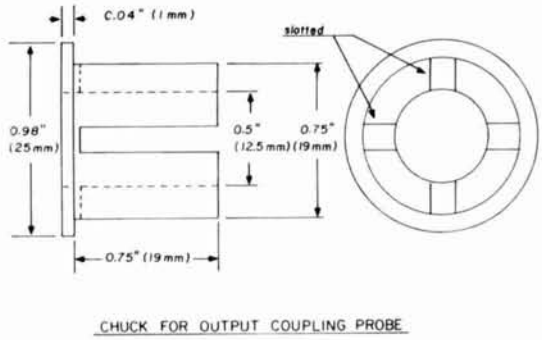


fig. 5. Chuck for the output-coupling probe.

on the bottom plate. The bottom plate contains the finger stock (Instrument Specialty Company 97-74) for grounding the grid connections.

The cathode compartment shown in **fig. 8**, measures 3.98 inch (10.1cm) OD by 2.05 inch (5.2cm) deep and

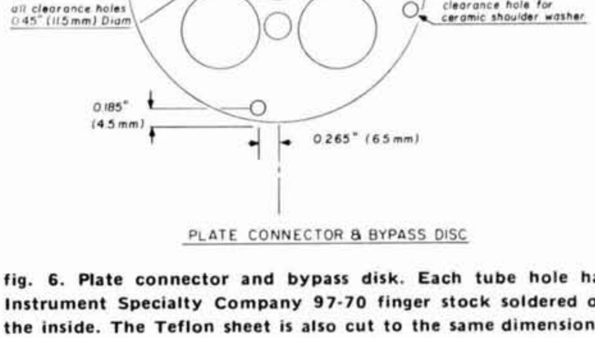


fig. 6. Plate connector and bypass disk. Each tube hole has Instrument Specialty Company 97-70 finger stock soldered on the inside. The Teflon sheet is also cut to the same dimensions.

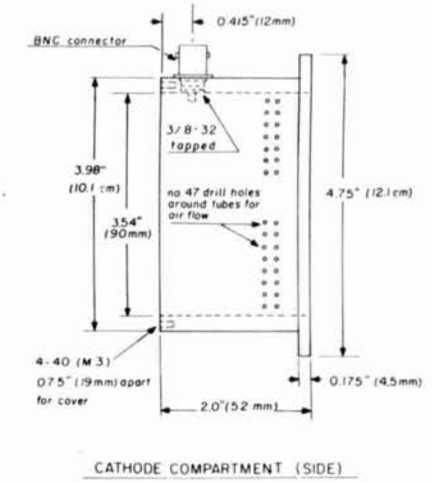


fig. 8. Side view of the cathode compartment. The ring that bolts to the anode cavity is shown in its normal position.

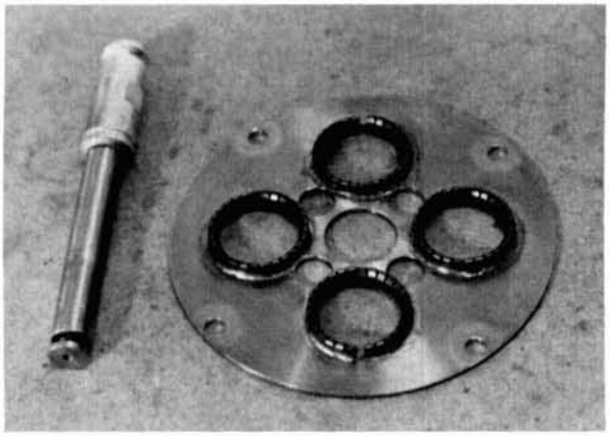


Plate bypass and connector disk shown with the output-coupling probe.

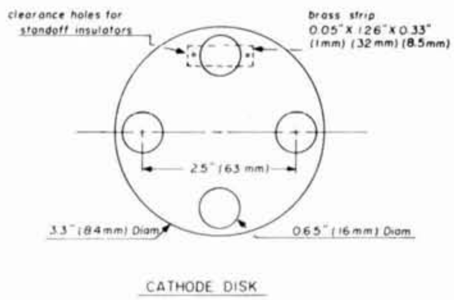


fig. 9. Cathode disk.

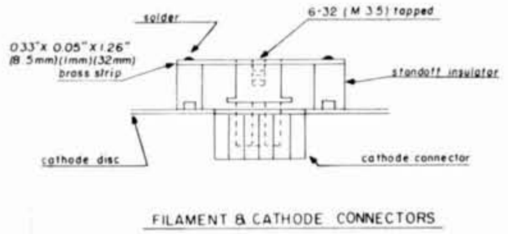


fig. 10. Filament and cathode connectors.

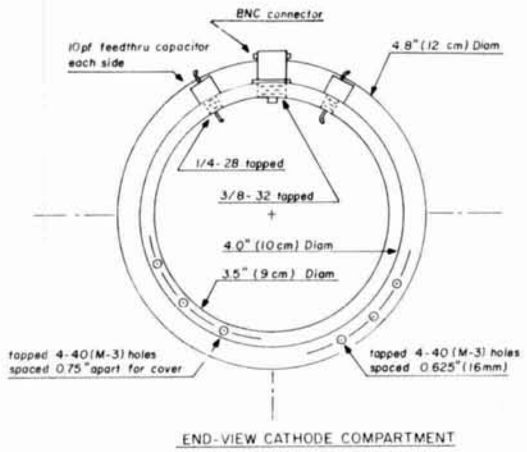


fig. 11. End view of the cathode compartment showing the placement of the video input connector and the filament feed through capacitors.

contains the cathode and filament connections for the 2C39 tubes. Its side wall is soldered to a 4.755-inch (12.1cm) OD by 3.98-inch (10.1cm) by 0.175-inch (4.5mm) brass ring. This allows the cathode compartment to be bolted to the anode cavity assembly by 4-40 (M3) screws spaced 5/8 inches (16mm) apart.

The cathode disk is a piece of 0.08-inch (20mm) thick brass cut into a 3.3 inch (8.4cm) diameter circle (**fig. 9**) with four 0.65-inch (16.5mm) holes in which Instrument Specialty Company 97-76 finger stock is fitted and soldered. The four filament connectors, shown in **fig. 10**, are made with two 3/8-inch (9.5mm) high insulated standoff terminals supporting a 1.26-inch (3.2cm) by 0.33-inch (8.5mm) by 0.05-inch (1mm) brass strip. The filament connectors, Instrument Specialty Company 97-280, are centered in the middle of the strips.

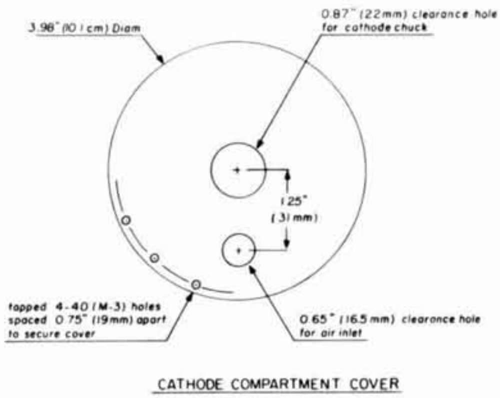
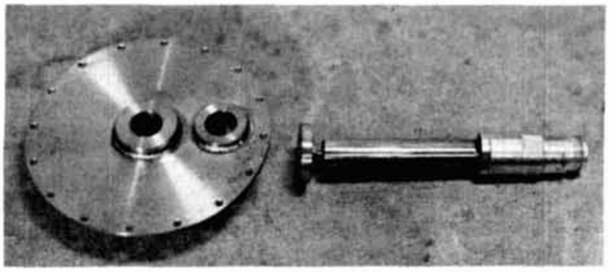


fig. 12. Cathode compartment cover.

A UG-625B/U BNC connector is screwed into the tapped hole into the cathode compartment outside wall. The center pin is connected to the cathode disk through a 10-turn rf choke. As shown in (**fig. 11**) two 10 pF feed through capacitors are also screwed into the tapped holes. They are used to furnish filament voltage to the tubes. The cathode circuit used in this amplifier is similar to the circuit used in the WB6IOM amplifier.



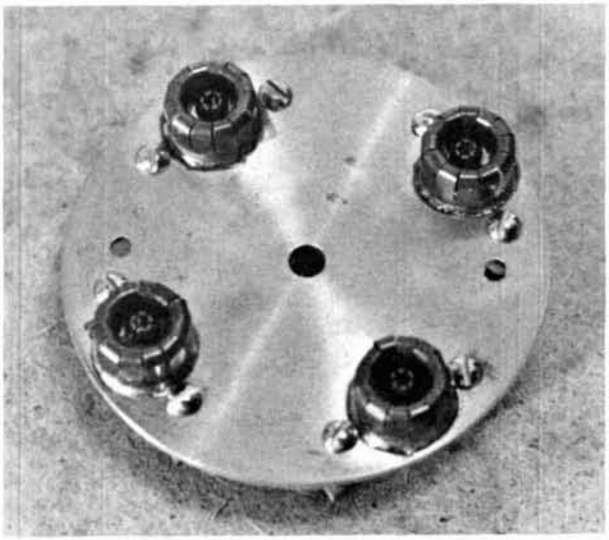
Cathode compartment cover disk with the input probe.

A 0.08-inch (20mm) by 3.98-inch (10.1cm) diameter disk with a 0.87-inch (22mm) center hole is used as the cathode compartment cover (**fig. 12**). A clearance hole (0.65 inches [16.5mm]) is also provided in the cover for an 1/8-inch (3mm) long air-inlet hose connector (**fig. 13**) which is soldered in place.

The cathode probe is 2.25 inches (5.7cm) long with another type-N connector on one end and a 0.978-inch (26.5mm) disk on the other. The assembly shown in **fig. 14** is assembled in the same manner as the anode probe. The chuck for the input probe is machined to an 0.87-inch (22mm) OD, 1.125 inches (28.5mm) long (**fig. 15**) and is soldered in place after being inserted into the cathode compartment cover center hole. The chuck is slotted and adjusted for a snug fit to the input probe.

cooling

Cooling for the amplifier is provided by a Dayton (model 2C781) 100-cfm, squirrel-cage, blower mounted



Completed cathode disk showing the filament and cathode connections.

on the rear of the main chassis. The air is forced into the enclosed chassis and then directed at the anode cooling fins by a plexiglass manifold and a cutout in the chassis. Air for the filament and grid seals is provided by a 5/8 inch (16mm) ID hose inserted into a chassis connector, *similar to the one mounted on the amplifier, and the* hose connector on the cathode compartment cover.

tuning up

It is suggested that before power is applied to the amplifier a few accessories be available. A 50-ohm

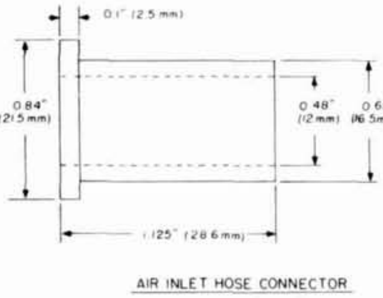


fig. 13. Air inlet hose connector.

dummy load, or resonant antenna, and a power indicator is necessary for proper amplifier tune-up. To prevent destroying the modular output transistors, the output cable is removed and an adjustable wirebound, 100-ohm 25-watt potentiometer is used to replace the modulator.

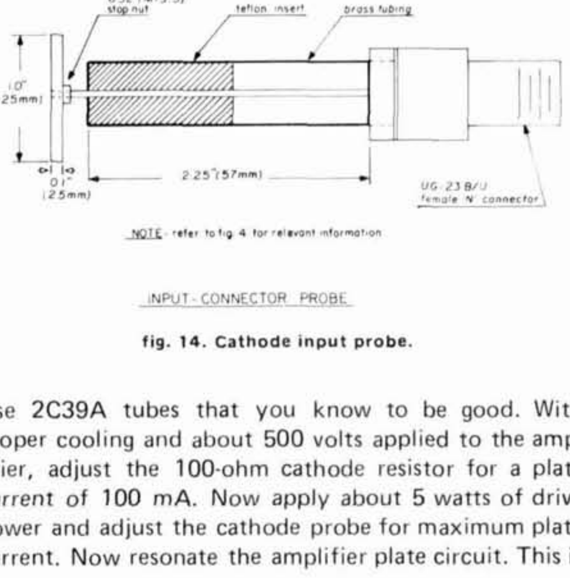


fig. 14. Cathode input probe.

Use 2C39A tubes that you know to be good. With proper cooling and about 500 volts applied to the amplifier, adjust the 100-ohm cathode resistor for a plate current of 100 mA. Now apply about 5 watts of drive power and adjust the cathode probe for maximum plate current. Now resonate the amplifier plate circuit. This is

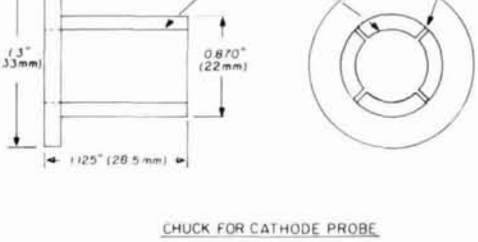


fig. 15. Cathode probe chuck.

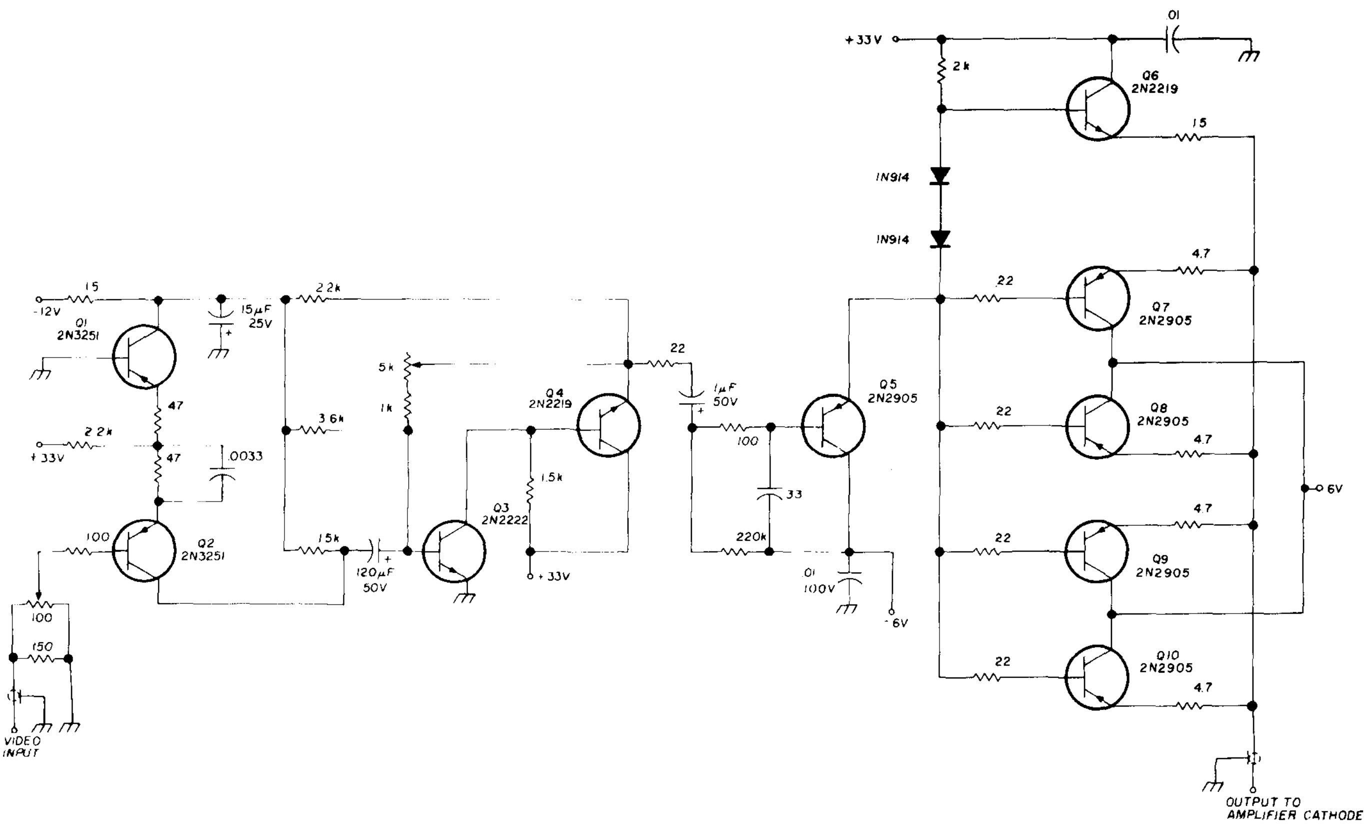


fig. 16. Schematic diagram of the cathode-video modulator. All resistors are ¼-watt. The supply voltages go to the modulator through 0.001 µF feedthrough capacitors.

done by starting with the four plate-tuning disks at about midrange. While closely observing the power-output indicator, vary the four tuning adjustments several turns in or out until there is power output. Since the output-probe adjustment and plate-tuning adjustments interact it's necessary to repeat all adjustments until maximum output power is obtained. All four tuning adjustments should be positioned equally and the output-coupling probe position optimized before the plate voltage is increased. After the plate voltage has been raised to 800 volts, the tuning must be readjusted for maximum output.

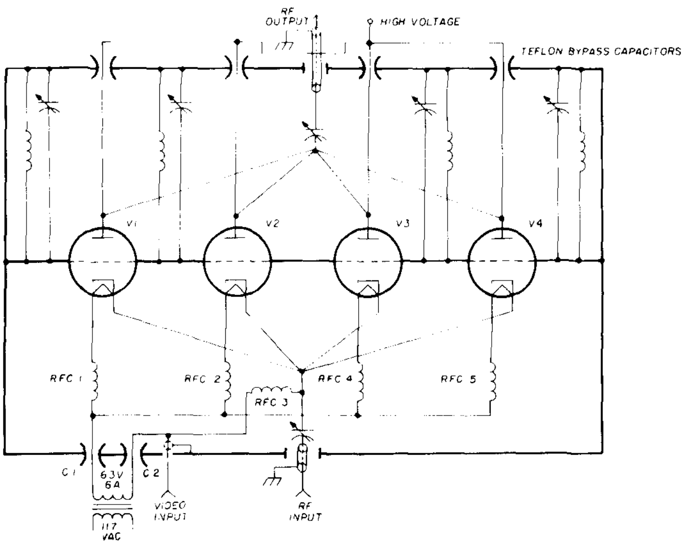


fig. 17. Schematic representation of the four-tube 1270-MHz amplifier. C1 and C2 are 10 pF, RFC1 through RFC5 are each 10 turns no. 22 AWG (0.6mm) airwound on 1/8-inch (3mm) diameter forms.

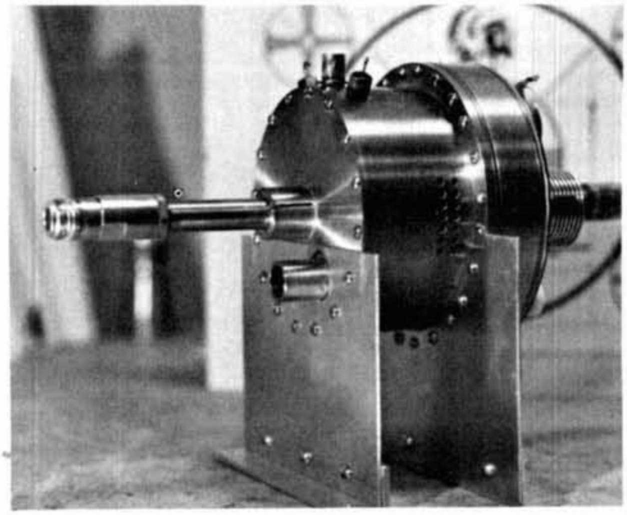
The video modulator may now be reconnected to the amplifier, and with the camera connected to the modulator it should be possible to obtain excellent pictures from an inline detector on the output transmission line.

cathode video modulator

The modulator circuit used for the amplifier is very similar to one in use by a uhf television station in the Chicago area. However, higher power-supply voltages were used to increase the video output swing, and emitter peaking was used on Q2. I also found it was desirable to include a bias adjustment of Q4 and to include two more 2N2905 transistors in the output stage to handle the additional cathode current. Originally this circuit was used to modulate a single 2C39A which was operated at lower plate current than normal. Normal video output level from closed-circuit TV cameras is 2 volts p-p ±1.5 volts, which is adequate for the modulator.

The power-supply requirements are −12 volts at 35 mA, +33 volts at 70 mA, and +6 volts at 600 mA. All voltages are regulated by 10-watt Zener diodes. Transistor Q5 is cooled with a clip-on heatsink.

The modulator is built on a piece of copper-clad

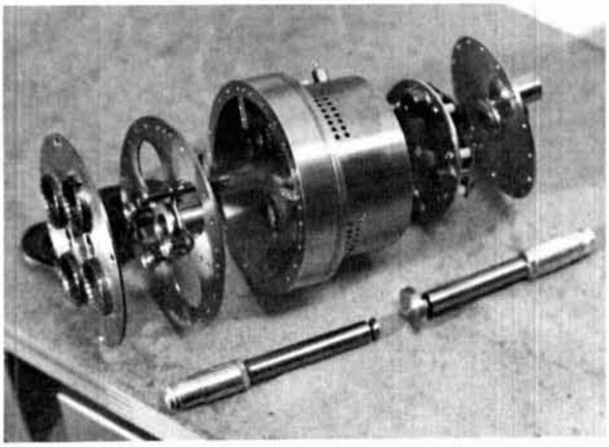


Rear view of the completed amplifier. Note the sandwich-type construction of the individual plates. Starting from the left: cathode comparment cover plate, cathode compartment, anode cavity bottom plate, anode cavity, anode cavity top plate, Teflon insulator, and anode bypass and connector plate.

printed-circuit board, mounted in a 2x3½x6¼-inch (5x8.9x15.9cm) Minibox. Stages Q7 through Q10 are mounted on an insulated 1-3/4x4-3/4 inch (4.5x12cm) heatsink on the top of the Minibox. The modulator output is directly coupled to the four-tube cavity through a short length of RG-59/U coaxial cable.

acknowledgements

I would like to thank the people in the Chicago area who worked together to prove that excellent results on ATV may be accomplished on the 1270-MHz band.



Disassembled amplifier showing the individual components.

Stations that participated in the venture include WA9CGZ, W9LK, W9DUT, WA9EUN, K9HDE, W9NAU, and W9YTM. It may be note worthy that with the many megahertz available in this band, it seems wasteful to tie up a large portion of the 440-MHz band with ATV repeaters.

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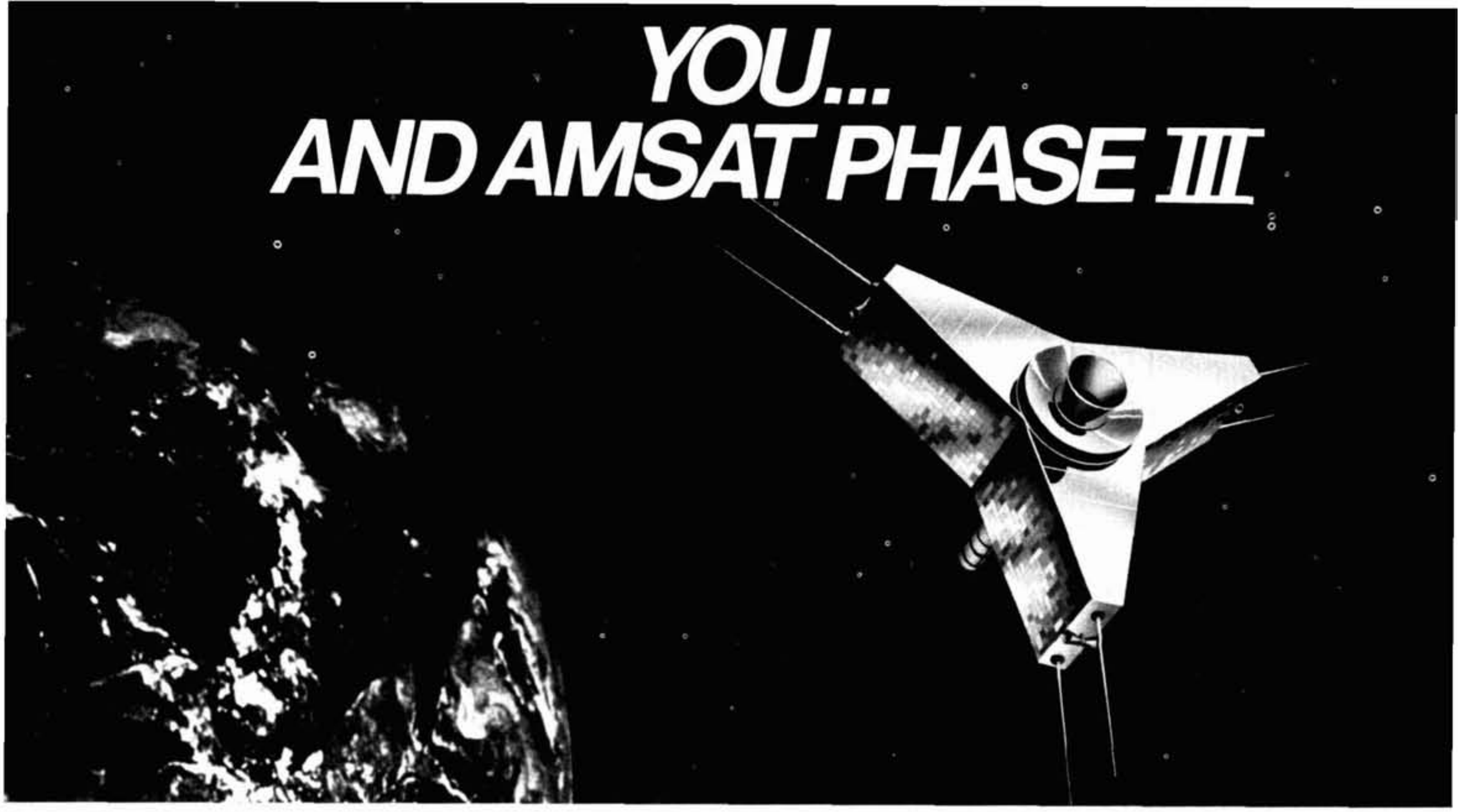

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MICROPROCESSORS

microprocessor interfacing: register pair instruction

In our last installment we discussed the single-byte data transfer instructions, MOV D, S and MVI <B2>, in the 8080 microprocessor instruction set. **Fig. 1** summarizes the significant points by indicating that for MVI r <B2> instruction, the data byte transferred to a register comes from the instruction itself, whereas for the MOV D, S instruction, the data byte is copied into the destination register D from the source register S.

Since the 8080 chip has a 16-bit address bus, there exists within the chip both a 16-bit *program counter* and a 16-bit *stack pointer*.[1] Due to this architecture, it would be very convenient to be able to manipulate full 16-bit address words within the chip. This is achieved by using the *register pair* operations. The six general purpose registers are treated as three sets of register pairs, B and C, D and E, and H and L. Each register pair is then designated by a unique 2-bit register pair code,

register pair	HI byte	LO byte	2-bit register pair code
B	B	C	00
D	D	E	01
H	H	L	10

The final 2-bit register pair code, 11, is reserved for either the stack pointer (SP) or the program status word (PSW), which contains the contents of the accumulator and the flag bits. *In register pair operations, the HI byte is always the most significant eight bits in the 16-bit memory address.* The LO byte is the least significant eight bits. Registers B, D, and H function as HI address bytes, and C, E, and L as LO address bytes.

As one example of a register pair operation, consider the three-byte *load register pair immediate* instructions:

LXI rp	
<B2>	[LO byte]
<B3>	[HI byte]

This permits you to move the data bytes contained within the second and third bytes of the instruction to the register pair, rp. The general form of the instruction is

0 0	R P O	0 0 1
Instruction class	2-bit code for register pair	

Some examples include:

data transfer operation	mnemonic	octal instruction code
<B2> → C	LXI B	001
	<B2>	<B2>
<B3> → B	<B3>	<B3>
<B2> → E	LXI D	021
	<B2>	<B2>
<B3> → D	<B3>	<B3>
<B2> → L	LXI H	041
	<B2>	<B2>
<B3> → H	<B3>	<B3>

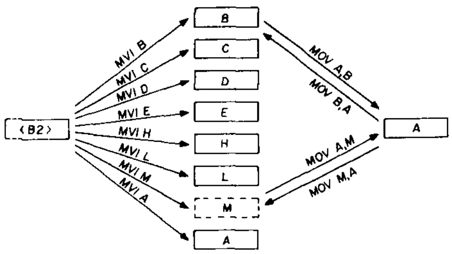

fig. 1. Schematic representation of the MVI r and several MOV D,S instructions.

It should be noted that the accumulator and "memory register" are not used as a register pair. To transfer data from memory location HI = 030 and LO = 123, you execute the following simplified program:

octal instruction code	mnemonic	comments
041	LXI H	Load register pair H with the following LO and HI address bytes
123	123	L register byte
030	030	H register byte
126	MOV D, M	Move data from the memory location addressed by register pair H to register D

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

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

Fig. 2 summarizes the four LXI rp instructions. Keep in mind that you cannot arbitrarily pair registers. For example, if you wish to load registers C and D with data for an operation, you will not be able to use an LXI rp instruction; use MVI C and MVI D instead. If you could substitute register E for register C, then you can use an LXI D instruction to load both register bytes into the indicated registers. Other useful register pair operations shown in **fig. 2** are:

register pair operation	octal instruction code	comments
XCHG	353	Exchange the contents of register pair H with the contents of register D. The HI bytes, H and D, exchange with each other and the LO bytes, L and E, exchange with each other.
SPHL	371	Load the contents of register pair H into the stack pointer.
PCHL	351	Load the contents of register pair H into the program counter.
LXI SP <B2> <B3>	061 <B2> <B3>	Move instruction bytes <B2> and <B3> into the stack pointer. <B2> is the LO byte and <B3> is the HI byte.

Since the program counter always contains the address of the next instruction to be executed, the register pair instruction PCHL is actually a jump instruction.

Two other useful instructions for manipulating 16-bit memory address words are the *increment register pair*, INX rp, and *decrement register pair*, DCX rp, instructions.

0 0	R P 0	0 1 1
instruction class	2-bit code for register pair	increment operation
0 0	R P 1	0 1 1
instruction class	2-bit code for register pair	decrement operation

With these instructions, the 16-bit contents of the register pair are either incremented or decremented as a single 16-bit word. However, no condition flags are affected by the INX rp or DCX rp instructions. You do not get a carry out of the most significant bit, the zero flag is not set when the register pair contents are zero, etc. It should be clear that the INX rp and DCX rp instructions are designed for 16-bit address operations, not for multiprecision arithmetic.

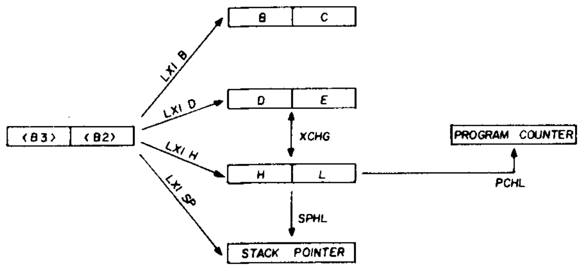

fig. 2. Representation of several register pair instructions, including LXI rp, XCHG, SPHL, and PCHL.

Other register pair operations include:

register pair operation	octal instruction code	comments
XTHL	343	Exchange the top of the stack with the contents of register pair H
DAD rp	011, 031, 051, or 071	Add contents of register pair rp to register pair H. Only the carry flag is affected.
PUSH rp	305, 325, 345, or 365	Push register pair rp on stack
POP rp	301, 321, 341, or 361	Pop register pair rp off stack
SHLD <B2> <B3>	042 <B2> <B3>	Move contents of register L to memory location specified in instruction bytes <B2> and <B3>. Move contents of the H register to the succeeding memory location.
LHLD <B2> <B3>	052 <B2> <B3>	Load register L with contents of memory location specified in instruction bytes <B2> and <B3>. Load register H with the contents of the succeeding memory location

The SHLD and LHLD instructions can be used as 16-bit I/O instructions if the input-output devices are addressed as memory locations.[2,3]
These instructions will be discussed in subsequent columns.

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3. B. Gladstone and P. D. Page, "Programming Hints Ease Use of Familiar Microprocessor," *Computer Design 15*, August, 1976, page 77.

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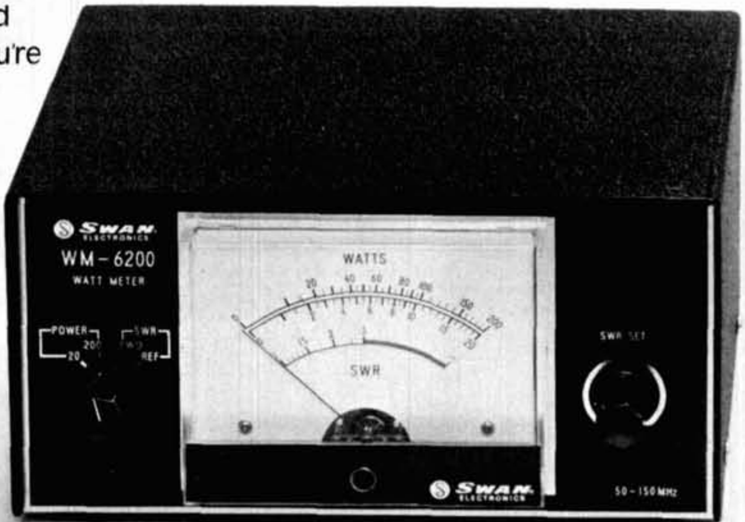

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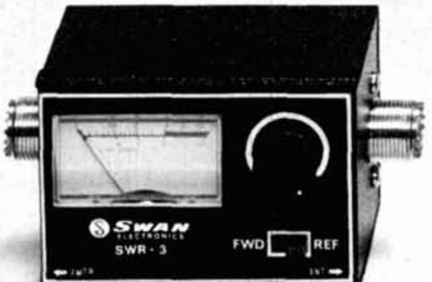
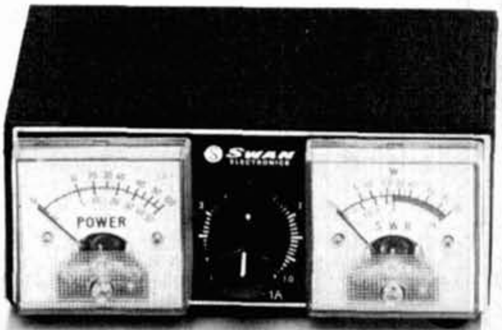
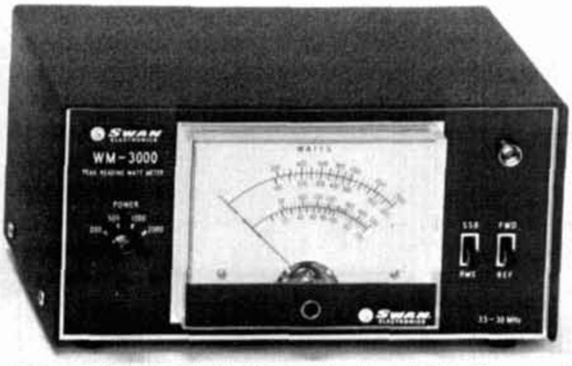


Easy-on-the-pocket pocket SWR. Mighty mite SWR meter with high accuracy. SWR-3 indicates 1:1 to 3:1 SWR at 50 ohms on frequencies from 1.7 to 55 MHz. Precision PC board directional coupler makes it a solid value at a rock-bottom price. **SWR-3 Pocket SWR Meter** ... **$12.95**



SWR bridge bridges the price barrier. This little jewel gives you relative forward power and SWR on two 100 microampere meters at a remarkably low price. Rear mounted coax connectors for easy, neat installation. Capable of handling 1000-watt signals on frequencies from 3.5 to 150 MHz. With low insertion loss, it's great for mobile operations, too. **SWR-1A Relative Power Meter and SWR Bridge** **$25.95**



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WM2000 In-Line Wattmeter. . $59.95

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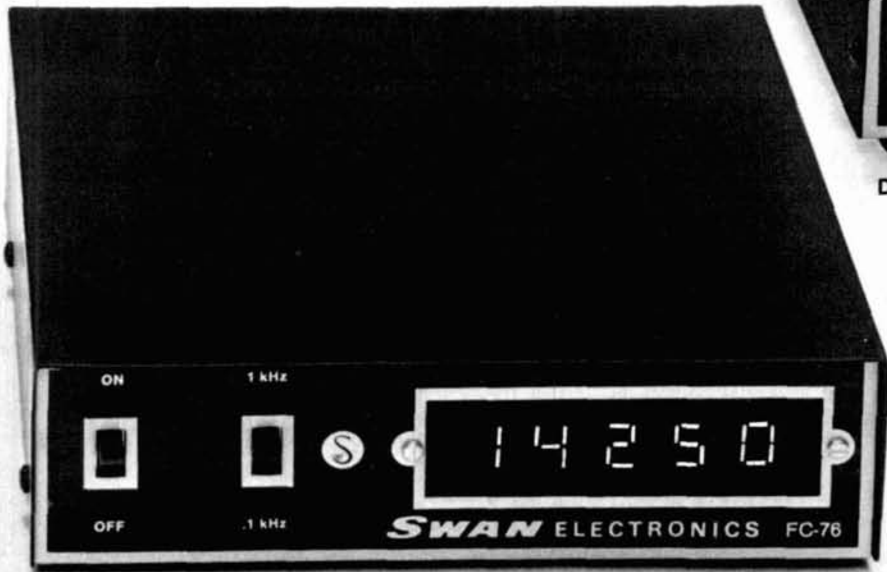

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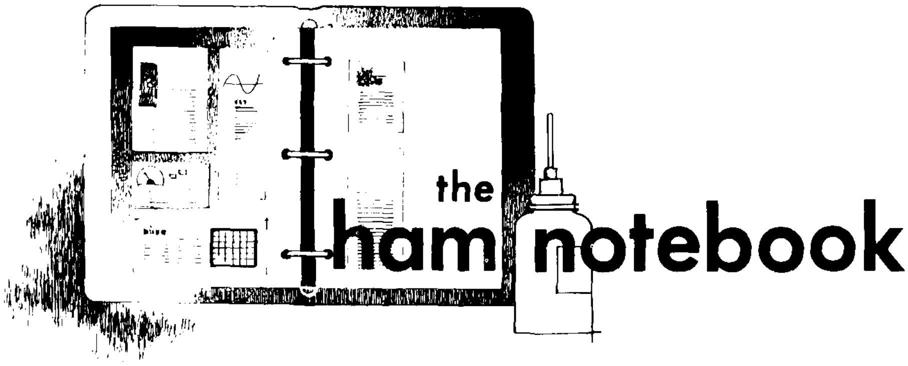

simple filter reduces frequency synthesizer sidebands

In the past few years frequency synthesizers have become popular for vhf and hf use. Most follow the basic scheme of **fig. 1** and use the Motorola 4044 (or Fairchild 11C44) phase detector IC. While good results are fairly easy to obtain, many designs miss a trick by simply copying the filter circuit in the data sheets. As a result, too much reference frequency feeds through and undesirable fm sidebands appear in the output.

As shown in **fig. 1**, the vco tunes from 7 to 8 MHz as its control voltage goes from 8 volts to 4 volts. The control

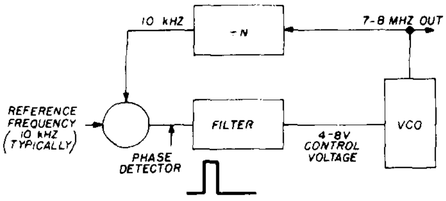

fig. 1. Frequency synthesizer loop diagram showing relationship of basic components.

voltage is only an amplified and filtered version of the phase detector output (ideally it is pure dc). A change in frequency over this range requires that the phase detector provide more or less dc accordingly. The trouble is that the phase detector outputs a pulse train and you are interested only in its *average* dc value (it is rather like a switching regulated power supply). If you filter very heavily, you will get pure dc but the control voltage will respond too slowly to be useful. Notice that if the vco is right on frequency, the phase detector

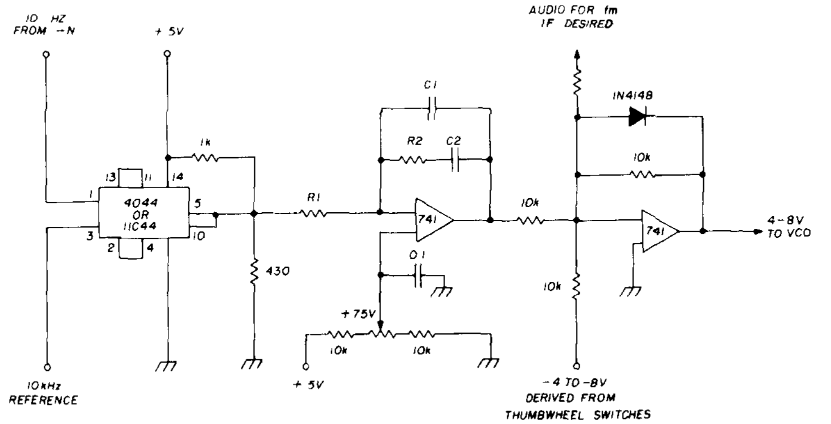

fig. 2. Schematic diagram of improved frequency synthesizer with filter circuit using readily available components, and featuring greater than 60 dB suppression of sideband frequencies.

must only provide a very small correction voltage. Clearly, the answer is to make the phase detector do as little work as possible. Since it operates by putting out pulses of constant amplitude and varying duty cycle, you want to reduce the pulses to nearly zero width.

The first trick is to keep the phase detector from doing all the work itself. While digital thumbwheel switches are used to set the frequency divider, the same digital information can be used (via a simple D-to-A converter) to provide some of the vco control voltage that will be needed. The other trick is to find a way to get more dc gain without heavier ac filtering. Discard the cheap and easy circuitry provided on the phase detector chip in favor of the slightly more complex arrangement shown in **fig. 2**. Note that at dc, the first 741 op amp runs at full throttle (at gain of 100,000), with the same filtering as the former circuit. Components can be chosen by the formulas provided for good performance with reference frequencies greater than 5 kHz, and LC oscillators for the vco.

$$R1 = \frac{1}{4.7 \times 10^6} \cdot \frac{K_o}{nC1} \ (ohms)$$

$$R2 = \frac{1}{1350 \times C1} \ (ohms)$$

$$C2 = \frac{C1}{14} \ (farads)$$

where K_o is the vco characteristic (in Hz/V) and should be approximately 200 kHz per volt, n is the middle-of-the-range divider reading, and C1 (farads) is chosen to provide a value of R1 of approximately 10k ohms.

It is really worth the cut-and-try to get a reasonable linear frequency output vs control voltage characteristic, as this results in sideband reduction. A very abrupt varicap can help here. I have had good luck with the KV5001 ($.70 from Solid States Sales, Somerville, Massachusetts).

To derive a tuning voltage from the thumbwheel switches you can buy a good $12 D-to-A converter from Hybrid Systems, Burlington, Massachusetts or make your own from cmos gates and some resistors (as shown in the RCA COSMOS circuits manual).

Tuneup is simple. Set the synthesizer to the middle of its range and observe the phase detector output on a good fast oscilloscope. You should see a dc level of about 1.5 volt with a string of needle-narrow pulses shooting either up to 2.25 volts or down to 0.75 volt, adjust the pot until they nearly vanish or are up half the time and down the other. Sidebands should now be more than 60 dB below the carrier.

Eric Blomberg, K1PCT

improved reliability for Collins KWM-2 transceivers

During the past fifteen years Collins has sold over 27000 KWM-2 transceivers, an impressive testimonial to the reliability and conservative design of this piece of equipment. However, minor changes to the KWM-2 over the years have been made to bring the design to its present state; and one of these changes is important to owners of older-model KWM-2s.

Connoisseurs of Collins equipment are quick to point out that KWM-2s with the new "meat ball" emblem on the panel are better than the older models which sport the "wing" emblem. The "wing" models are alleged to have "relay" problems, solved in the later models by incorporation of encapsulated plug-in relays.

In my opinion this subtle distinction is not valid, inasmuch as the alleged relay problems don't seem to have anything to do with the *type* of relays installed, and contact burn (normally avoided by encapsulated relays) is not a problem. Instead, the difficulty appears to be that the vox relay (K2) coils in earlier model KWM-2s sometimes have a tendency to burn out, whereas later models do not have this problem. And a vexing problem it is, too, as anyone who has attempted to rewire the vox relay in a KWM-2 can tell you! It is an onerous, time-consuming, exacting, and humbling job.

A much easier approach is to modify the KWM-2 relay coil circuit to prevent coil burn-out, a modification recommended to owners of older KWM-2s that may not have already incorporated this simple change.

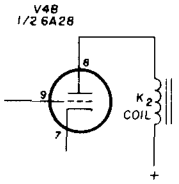

fig. 3. Original vox relay circuit used in the Collins KWM-2 transceiver, without protective resistor.

The vox relay circuit for older KWM-2 models is shown in **fig. 3**. The vox relay actuator tube (V4B) operates under vox control, push-to-talk, and when the transceiver is actuated for CW. When the vox circuit is actuated, the relay tube conducts heavily and plate current flows through the coil of vox relay K2. In this design, the plate current of V4B is an order of magnitude greater than the maximum specified relay coil current. Overheated relay coils have a tendency to burn out, and the relay in the KWM-2 is no exception.

In some models of the KWM-2 a *protective bias resistor of 330 ohms is* connected between the cathode of the vox relay amplifier tube V4B (pin 7) and the tip contact of the microphone jack. The presence of this resistor can be determined with an ohmmeter. If present, the following description for adding a 12k, 2-watt resistor in the plate circuit may be disregarded.

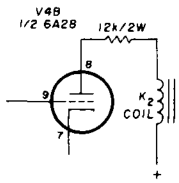

fig. 4. Simple modification of the relay circuit to provide protective 12k resistor between plate of tube V4B (pin 8) and coil of the vox relay.

The circuit modification is simple (**fig. 4**). A 12k, 2-watt resistor is added in series with the relay coil to reduce the coil current to a safe value. This addition in no way affects relay operation.

You can check your KWM-2 to find out if this modification has already been included in the circuit. In most models, the resistor is connected from pin 8 of tube socket XV4 (6AZ8) to an adjacent terminal point. If this resistor or the cathode resistor is missing, the following modifications are suggested:

1. Disconnect the red-white wire from pin 8 of socket XV4 and connect it to a terminal tie point. It is possible to mount a small standoff terminal on a bolt of the socket, or else to epoxy a miniature terminal strip adjacent to the socket.

2. Install a 12k, 2-watt resistor (R202 in the latest KWM-2 circuitry) from the terminal point to pin 8 of socket XV4. This completes the modification. Apply power and test the vox circuit for activity.

William I. Orr, W6SAI

Heath intrusion alarm for mobile equipment

The Heathkit GD-39 ultrasonic intrusion alarm should be ideal for mobile operators who want to protect their equipment. While I haven't made a battery-operated vehicle installation, the facts speak for themselves: 12V operation, 17.5V out of the rectifiers into the regulator, 13 mA current drain while waiting, and room on the back panel to mount some connectors (**fig. 5**). Some modifications will have to be made: a relay is required for spst operation (in-

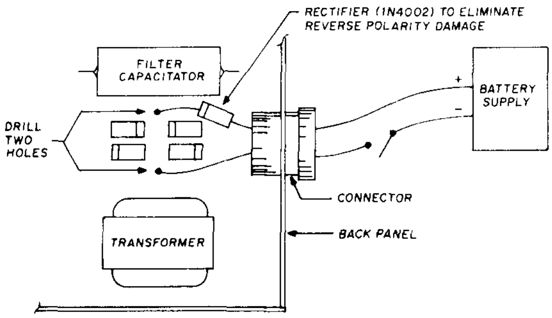

fig. 5. Adaptation of Heathkit GD-39 intrusion alarm for mobile equipment.

stead of switching the line), and two holes must be drilled into the board to bring in the battery supply.

William B. Rossman

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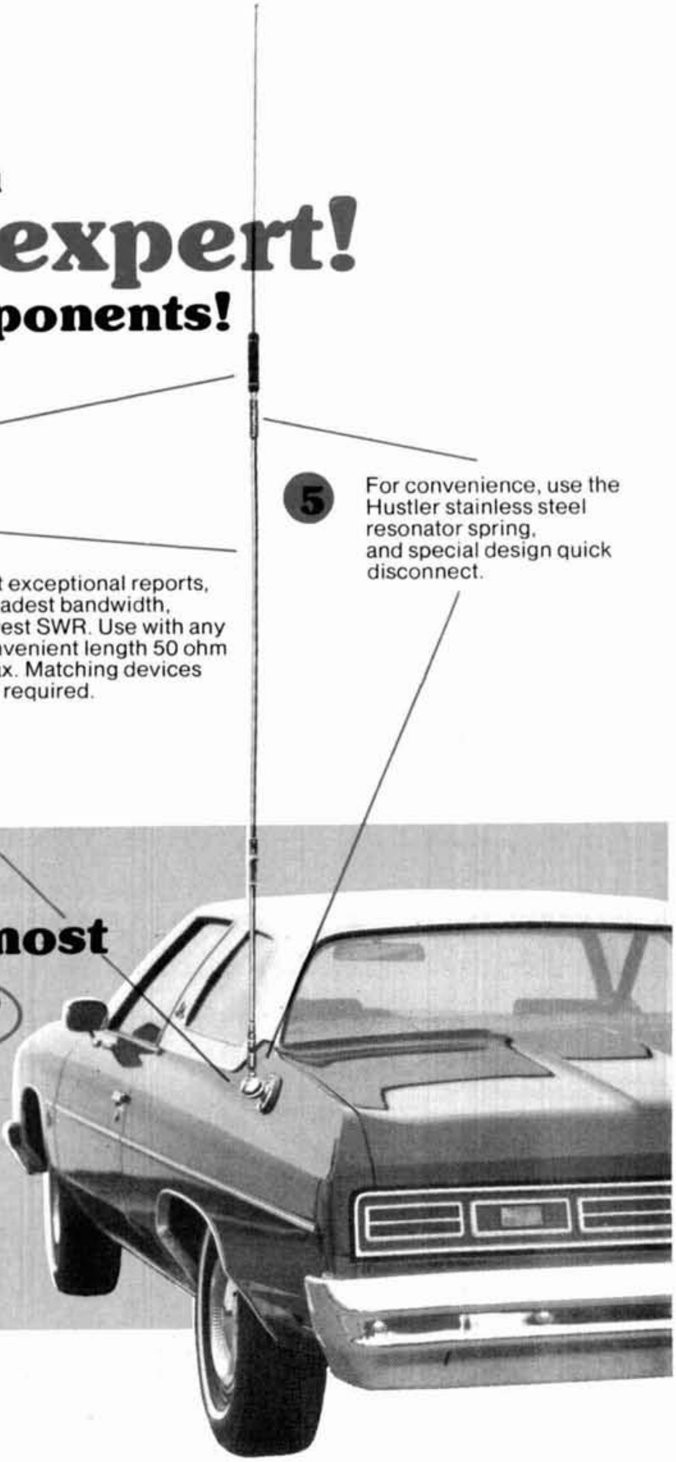

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NEW products

Millen 90673 antenna bridge

The Millen 90673 Antenna Bridge is a radio frequency impedance measuring device which is intended for use in the frequency range from 1.8 to 30 MHz. An "extra" provided by this bridge is the measurement of rf resistance outside the normal high-frequency limit. Using additional coils, specially made for the vhf range, it is possible to obtain approximate resistance readings up to 60 MHz, but without any accompanying reactance calibration.

A null detector, with diode rectifier and dc amplifier, is built in and indicates on a 1-mA meter. The bridge requires an external oscillator, such as a grid-dip oscillator.

Resistance measurements are made by varying a differential air dielectric variable capacitor. The resistance range covered is approximately 5 to 150 ohms. At frequencies below 15 MHz, values up to 200 ohms are measurable. The accuracy of resistance measurement is better than ± 10 per cent over most of the range.

Reactance is measured by tuning a coil-and-capacitor circuit, first to an initial null with a resistive reference and then to a new null with the unknown impedance. This requires simultaneous balancing of resistance and reactance controls. The reactance range is calibrated in ohms of reactance at 1MHz and indicated inductive and capacitive values. The reactance value at the operating frequency is obtained by dividing the indicated reading by the frequency in use.

The Millen 90673 Antenna Bridge is based on the "Macromatcher" described in *QST*, (January, 1972). The circuit has been modified and the mechanical layout arranged to produce the commercial product. Since the original Millen 90672 Antenna Bridge was the starting point for the Macromatcher, the new 90673 Antenna Bridge is a logical and useful further development. The bridge is approximately 8 x 8 x 5 inches (20x20x13cm), not including plug-in coil storage box.

The 90673 Antenna Bridge normally measures coaxial-line connected impedances (unbalanced). A set of balun coils, with series and parallel reactances tuned out for amateur bands, is available at extra cost to measure balanced-line systems.

power amplifier and preamp for 2-meter hand-helds

Hamtronics recently announced a new PA/Preamp Unit, the T130, for use with two-meter hand-held units and other low-power transceivers in either mobile or base-station applications. The *T130 PA/Preamp Unit* provides up to 25-watts output for inputs between 200 milliwatts and 6 watts (specify output power of your transceiver).

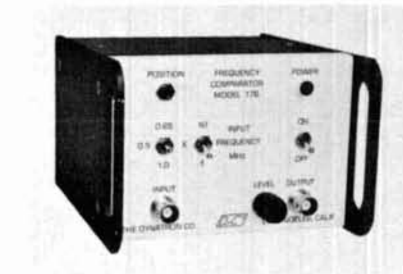
The power amplifier section is solid state, requires no tuning, and the preamplifier portion provides 20-dB gain, using a low-noise fet cascode preamp circuit. Transmit-receive switching is automatic, and a front-panel switch allows the unit to be turned off, and the transceiver bypassed to the antenna, when fringe-area coverage is not required.

The PA/Preamp unit is attractively packaged in a high-impact *Cycolac* case with brushed-aluminum panels. Your hand-held connects to a front-panel BNC connector, while the antenna is connected through a rear-panel uhf connector. *On/off* status is indicated by an LED. In operation, the current drain is 4 amperes.

If desired, the unit is also available with a front-panel loudspeaker and audio amplifier to provide improved receiver audio for your transceiver. Another option diverts the transceiver's audio to your car's a-m broadcast receiver, for the same purpose. The 3-1/2 by 7-1/2 by 7-inch (9x19x18mm) cabinet provides ample room for your own additions; *e.g.*, microphone connector and hanger, a clip to hold your hand-held unit, and a 12-volt lead for its operation.

The basic T130 PA/Preamp Unit is $129.95. A version without the preamplifier is available for $109.95. For a complete catalog of vhf and uhf products, send a self-addressed, stamped envelope to: Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

frequency comparator



The Dynatron Company has announced a new frequency comparator, the *DyCo Model 176*, intended for use in calibrating crystal oscillators and time bases in digital frequency counters, communications monitors, and frequency meters against television network atomic frequency standards.

The calibration scheme, devised by The National Bureau of Standards, can now achieve in minutes the same results that previously took hours by receiving WWVB. Traceability to NBS is assured by the monthly publication of the network offset frequencies in the *NBS Services Bulletin*.

A color TV receiver (any set will do) receives the network color subcarrier and displays the result of the comparison as a vertical rainbow bar. Frequency calibrations to accuracies within parts in 10^{10} take just a few minutes.

The *DyCo Model 176* accepts inputs of 0.25, 0.5, 1.0, 2.5, 5.0, or 10 MHz. Price of the *Model 176* is $179.95. For additional information, write The Dynatron Company, Box 48822, Los Angeles, California 90048.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74S00	.35	555	.50	309K	.99	NPN 2N3904 type	10/$1.00
74S112	.75	556	.75	309H	.99	PNP 2N3906 type	10/$1.00
7447	.79	567	1.75	340K-12	1.25	NPN Power Tab 40W	3/$1.00
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7475	.50	LED DRIVER		7812	.99	FET MPF-102 type	3/$2.00
7490A	.55	75491	.50	7815	.99	UJT 2N2646 type	3/$2.00
74143	3.50	75492	.50	7818	.99	2N3055 NPN Power	.75

SPECIAL

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

LED DISPLAYS

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FND 510 ...... .5" C.A. 1.25
DL 707 ...... .33" C.A. 1.25
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741 OP-AMP SPECIAL

Factory prime mini dip with both Xerox and 741 part numbers
10 for $2.00

SOCKETS

14 PIN 5/$1.00
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Assortment of 12 most used IC sockets. Good to *have around the* shop. $1.95

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Factory Prime — includes FREE socket!

21L02-1 $1.95 Fast 450 ns Low Power

60 Hz XTAL TIME BASE

- Runs on 5-15 VDC
- Low current (2.5 ma)
- Operate clocks in car, boat, plane
- 1 minute/month accuracy

Kit, TB-7 ... $5.50
Assembled & Calibrated ... $9.95

CHEAP CLOCK KIT $9.95

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

Does not include board or transformer

PC Board $2.95
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FM Wireless Mike Kit $2.95

Transmit up to 300' to any FM broadcast radio. Sensitive mike input requires crystal ceramic or dynamic mike. Runs on 3 to 9 V.

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

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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.

- Powered by 6-16vdc, unregulated
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$3.00 each

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Swan beam antennas are precision engineered to give you a full 2000-watt P.E.P. rating. They're designed for a VSWR of 1.5:1 or better at resonance. They'll give you optimum gain and they're built tough and rugged to stand up to some of the meanest environments.

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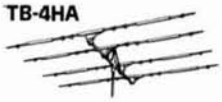
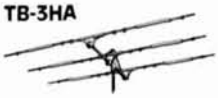
TB-4HA



Heavy-duty, four-working-element antenna for 10, 15 and 20 meters. **$259.95**

TB-3HA



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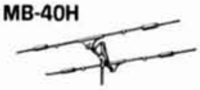
TB-2A

Light-weight, two-working-element antenna for 10, 15 and 20 meters. **$129.95**

MB-40H



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TB-3HA	16' x 1.5"	28'-2"	16'	100 mph	110 lbs.	4 sq.ft.	44 lbs.
TB-2A	6.5' x 1.5"	27'-8"	14'-3"	80 mph	60 lbs.	1.8 sq.ft.	18 lbs.
MB-40H	15.75' x 1.5"	30'-4"	17'-6"	100 mph	80 lbs.	2.5 sq.ft.	40 lbs.

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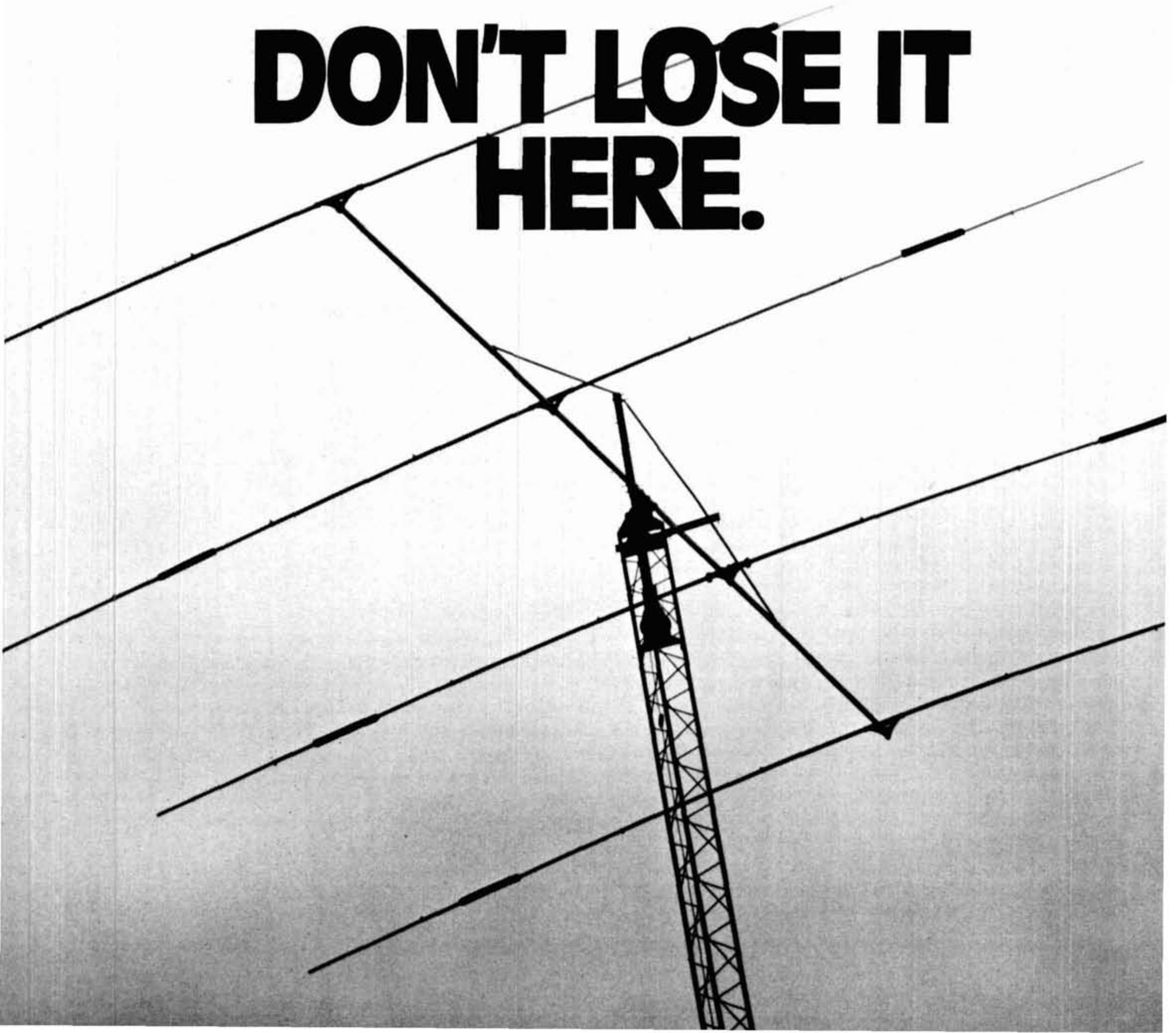

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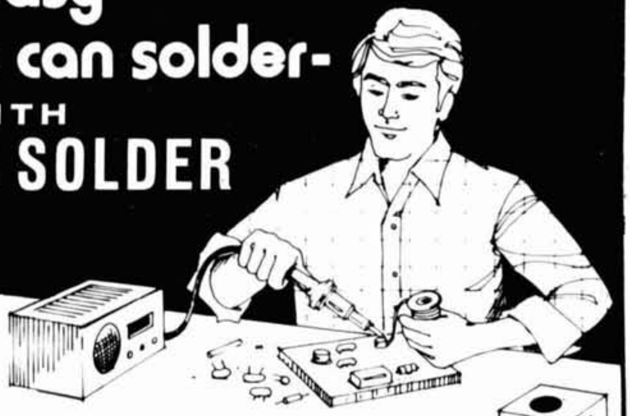

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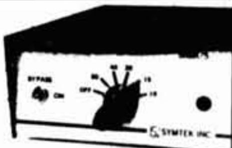
SABA-5



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10		OTHER (Specify Below) Such as Antenna Matcher, Speech compresser, etc.	Request Information (Tell us about your eqpt.		

TOTAL FOR MDSE	
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FLA. RES. ADD 4% S.T.	
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NEW!
FM144-10SXRII

KDK

All Solid State-PLL digital synthesized — No Crystals to buy! 5KHz steps — 144–149 MHz-LED digital readout PLUS MARS-CAP.*

• 5MHz Band Coverage — 1000 Channels (instead of the usual 2MHz to 4MHz — 400 to 800 Channels) • Priority Channel • Audio Output 4 Watts • 15 Watts Output • Unequaled Receiver Sensitivity and Selectivity — 15 POLE FILTER, MONOLITHIC CRYSTAL FILTER AND AUTOMATIC TUNED RECEIVER FRONT END — COMPARE!! • Superb Engineering and Superior Commercial Avionics Grade Quality and Construction Second to None at ANY PRICE.

- **FREQUENCY RANGE:** Receive: 144.00 to 148.995 MHz, 5KHz steps (1000 channels). Transmit 144.00 to 148.995 MHz, 5KHz steps (1000 channels) + MARS-CAP.*
- **FULL DIGITAL READOUT:** Six easy to read LED digits provide direct frequency readout assuring accurate and simple selection of operating frequency.
- **AIRCRAFT TYPE FREQUENCY SELECTOR:** Large and small coaxially mounted knobs select 100KHz and 10KHz steps respectively. Switches click-stopped with a home position facilitate frequency changing without need to view LED'S while driving and provides the sightless amateur with full Braille dial as standard equipment.
- **FULL AUTOMATIC TUNING OF RECEIVER FRONT END:** DC output of PLL fed to varactor diodes in all front end R-F tuned circuits provides full sensitivity and optimum intermodulation rejection over the entire band. **No other amateur unit at any price** has this feature which is found in only the most sophisticated and expensive aircraft and commercial transceivers.
- **TRUE FM:** Not phase modulation — for superb emphasized hi-fi audio quality second to none.
- **FULLY REGULATED INTEGRAL POWER SUPPLIES:** Operating voltage for all circuits, i.e., 12v, 9v and 5v have independently regulated supplies. 12v regulator effective in keeping engine alternator noises out and protects final transistor from overload.
- **MONITOR LAMPS:** 2 LED'S on front panel indicate (1) incoming signal-channel busy, and (2) un-lock condition of phase locked loop.
- **DUPLEX FREQUENCY OFFSET:** 600KHz **plus** or **minus**, 5KHz steps. Plus simplex, any frequency.
- **MODULAR COMMERCIAL GRADE CONSTRUCTION:** 6 unitized modules eliminate stray coupling and facilitate ease of maintenance.
- **ACCESSORY SOCKET:** Fully wired for touch-tone, phone patch, and other accessories.
- **RECEIVE:** .25 uv sensitivity. **15 pole filter** as well as monolithic crystal filter and **automatic tuned** LC circuits provide superior skirt selectivity.
- **AUDIO OUTPUT: 4 WATTS.** Built in speaker.
- **HIGH/LOW POWER OUTPUT:** 15 watts and 1 watt, switch selected. Low power may be adjusted anywhere between 1 watt and 15 watts, fully protected — short or open SWR.
- **PRIORITY CHANNEL:** Instant selection by front panel switch. Diode matrix may be owner re-programmed to any frequency (146.52 provided).
- **DUAL METER:** Provides "S" reading on receive and power out on transmit.
- **OTHER FEATURES:** Dynamic microphone, mobile mount, external speaker jack, and much, much, more. Size: 2⅜ x 6½ x 7½. All cords, plugs, fuses, mobile mount, microphone hanger, etc., included. Weight: 5 lbs.

NEW! 6 METER FM50-10SXRII

Same specifications as above except transmit/receive: 51.00-53.995 MHz. 600 channels
Introductory Price $389.00

NEW!
TONE ENCODER/ DECODER

SC-12A ... 12 CHANNELS ... DUAL TONE
Introductory Price $119.00

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$389.00
VALUE $599.00

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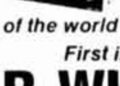

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MODEL FMTP-1
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WANTED Heath HW-18 SSB XCVR. State Price and condition. Dennis Vernacchia WB6LGI, 8993 Covina St., San Diego, Ca. 92126.

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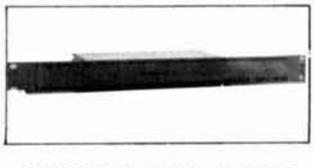
SAVE! BEARCAT SCANNERS. $259.95. New. E.R. Brooks Guns, 6612 Eastern Ave., Bell Gardens, CA 90201 (213) 773-5193.

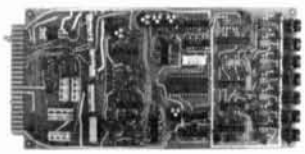
SELL Drake 2C, 2CQ, 2NB, Calibrator, Mint condition, w/manuals, $250. All new parts for WA4DSY HT220 Synthesizer, $75. Price includes shipping. Wanted 4BTV. WA9WDB/5 John Teles, 10511 Tenneco, Houston, Tx. 77099. Phone 713-495-2340.



REPEATER AUTO PATCH

It's complete — a single digit access/disconnect Auto Patch facility. All you need is a repeater and the phone line. Complete with automatic disconnect, dial-in capability, two way audio monitor plus remote control. When used with a rotary dial exchange, Data Signal's DPC-221 dial converter is also required. P.C. board or Rack Mount available.

RAP-2 PC **$99.50** Rack **$149.50**



DATA TONE TO DIAL PULSE CONVERTER

Convert standard 0-9 Data Tone digits to Bell System compatible dial pulse code. Completely solid state. Includes state-of-the-art Phased Locked Loop anti-falsing Data Tone decoder, large capacity 64-digit memory and solid state pulsing. Starts dialing on first incoming digit. Memory will not become congested due to rapid succession of incoming digits. Cancel and redial function. * and # digits are decoded and provided for remote control purposes. Available as p.c. board or rack mounting.

DPC-221 P.C. Board **$219.50**
DPC-221R Rack Mount **$299.50**

DELUXE C-MOS P.C. KEYER

Complete C-MOS keyer in kit form, or wired, just right for your custom enclosure. Versatile controls allow wide character weight variations, speeds from 5 to 50 w.p.m. plus volume and tone control. Solid-state output switching saves power, eliminates all those annoying relay problems and is compatible with both grid block and solid-state circuitry. With its side-tone monitor and 90 day warranty the Data Signal PC Keyer is the one for you.

C-MOS Keyer Wired **$24.95**; Kit **$19.95**

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flea market

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.

GLOBE SCOUT Mod. 65A transmitter (has 160 meters) need copy of schematic. George Lazarus WA9YEN, 639 Stephen Drive, Palatine, IL 60067.

MOTOROLA HT220, HT200, and Pageboy service and modifications performed at reasonable rates. WA4FRV (804) 320-4439, evenings.

TECH MANUALS for Govt. surplus gear — $6.50 each: SP-600 JX, URM-25D, OS-8A/U, TS-173/UR. Thousands more available. Send 50¢ (coin) for 22-page list. W3IHD, 7218 Roanne Drive, Washington, DC 20021.

POWERFUL, ADJUSTABLE, REGULATED, three output power supply and 900 easily removable parts in complete CARTRIVISION television recorder electronic assembly with documentation. Perfect for MICROPROCESSOR, IC, transistor, television, CB radio applications. $21.45. Free brochure. MADISON ELECTRONICS, INCORPORATED, 369, D77, Madison, Alabama 35758. SATISFACTION GUARANTEED.

MOBILE BONDING STRAPS under 50¢ each. Literature. Estes Engineering, 930 Marine Drive, Port Angeles, Wash. 98362.

NEED INFORMATION on BC-639-A. Identity of VT202 and VT203 please. W5PAE, 3308 Bahama NE, Albuquerque, New Mexico 87111.

FERRITE BEADS: w/specification and application sheet - 10/$1.00. Assorted PC pots - 5/$1.00. Miniature mica trimmers, 3-40 pf. - 5/$1.00. Postpaid. Includes latest catalog. Stamp for catalog alone. CPO Surplus, Box 189, Braintree, MA 02184.

THE "CADILLAC" of QSL's! New! Samples: $1.00 (Refundable) — MAC'S SHACK; Box #1171-D; Garland, Texas 75040.

VERY in-ter-est-ing! Next 4 issues $1. "Ham Trader" Yellow Sheets, Sycamore, IL 69178.

WANTED — Diagrams and or manuals for Collins R648/ARR/41 Rcvr. VE3FEZ, 421 Lodor Street, Ancaster, Ontario.

MODERN CODE PRACTICE. 0-22wpm on four 60 min. cassettes, $10. Royal, P. O. Box 2174, Sandusky, Ohio 44870.

SALE 6 pole 250 Hz crystal filter for Yaesu FT101 $48. 5-171, Jojo-cho, Kasugai-city, Aichi, 486, Japan JA2UEO. Koichi, Kawai.

EXCLUSIVELY HAM TELETYPE 23rd year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 35¢. $3.50 per year. Box 837, Royal Oak, Michigan 40868.

FERRIC CHLORIDE, Liquid 40% Board etching grade. Pints, quarts, gallons. Send SASE Bob's Electronics, Box 393-H, Bay City, MI 48707.

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada. MIS 3B4.

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

POWER SUPPLIES — Special low prices on High Quality power supplies. For specifications write: ASTRON Corporation, 1971 South Ritchey, Santa Ana, California 92705.

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WANTED CRYSTAL IMPEDANCE METERS TS-330, TS-683 or TSM-15 Glenn R. Kurzenknabe, K3SWZ, 403 Centerview Av., New Cumberland, Pa 17070.

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.

SWEEP GENERATOR, Telonic Model SPX-5, 250 to 500 MHz, adjustable sweep width, marker generators. Up to 5 watts output. $295, plus freight collect. Vanguard Labs, 196-23 Jamaica Ave., Hollis, NY 11423.

STAINESS AND GALVANIZED STEEL antenna guy wire our specialty. Wilcox Electronics, Box 1331, S. L. C., Utah 84110.

TEST EQUIPMENT

All equipment listed is operational and unconditionally guaranteed. Money back if not satisfied — equipment being returned must be shipped prepaid. Include check or money order with order. Prices include UPS or motor freight charges.

BOONTON 190A Q-mtr 30-200mHz	$425
BOONTON 202E AM-FM sig gen 54-216mHz	425
DEI TDU-4 30mHz video display	55
GR1001A LF sig gen 5kHz-50mHz	385
HP100ER Freq stand .05 parts/mil Outputs 10,100Hz; 1,100,100kHz	155
HP160B(USM105) 15mHz scope with reg horiz, dual trace vert plugs	375
HP166B(Mil) Delay sweep for above	130
HP170A(USM140) 30mHz scope with reg horiz, dual trace vert plugs	475
HP175A 50mHz scope with reg horiz, dual trace vert plugs	565
HP185A Sampling scope to 1 gHz 186B xstr rise time plug	585
HP202B LF Osc .5Hz-50kHz 10v out	75
HP205AG Lab audio gen .02-20kHz	195
HP212A Pulse gen .06-5kHzPRR	65
HP524D Freq counter-basic range 10Hz-10mHz extends w/plug-ins	195
HP540B Trans osc to 12.4gHz for use w/HP524-tyoe counters	145
HP616 Sig gen 1.8-4gHz FM-CW	365
HP686 Sweep gen 8.2-12.4gHz sweep range 4.4mHz-4.4gHz	495
HP803A VHF Ant bridge 50-500mHz	135
SOLITRON 200A SCR tester — all functions w/scope output	135
Tek181 Time-mark scope calib.	55
Tek190 Sig gen(const ampl) 50mHz	125
Tek536 11mHz X-Y scope — uses two letter-series plug-ins	295
Tek565 Dual beam 10mHz scope less plug-ins (3 series)	625
TS505 Std VTVM (rf-500mHz)	65

For complete list of all test equipment send stamped, self-addressed envelope

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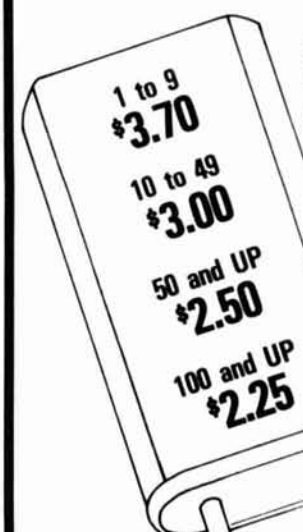

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flea market

U.S. and FOREIGN CALLBOOK, Radio Handbook, and other ARRL books, 1950-to 1975 edition, Ham Radio, QST, CQ, 73, Popular Electronics magazines, second-hand, I want, cash-paid. Also your unused tubes, fet-mosfet, transistors, IC, Xtal, filters, parts. Send the price-list to: Vittorio Bruni I-05030 COLLESTATTE (Terni) Italy.

HELP FOR YOUR NOVICE, general, advanced ticket. Recorded audio-visual theory instruction. Free information. Amateur License Instruction, P. O. Box 6015, Norfolk, Va. 23508.

FREE CATALOG. Solar Cells, Nicads, Kits, Calculators, Digital Watch Modules, Ultrasonics, Strobes, LEDS, Transistors, IC's, Unique Components. Chaneys, Box 27038, Denver, Colo. 80227.

20A, 13.8VDC POWER SUPPLY: 115VAC input, 50% duty cycle; excellent electronic regulation; electronic over-current and short circuit protection; crowbar over voltage protection, perfect for 200W class transistorized transceiver; $99 postpaid. WAØNSY, Box 728, Huron, S.D. 57350.

WIRE DIPOLE CUTTING CHART & notes on installation. Chart covers 2 meters through 160 meters. Helpful information. $1.00 ppd. W.E. Schroedter, 6108 N. Sherman Blvd., Milwaukee, Wisc. 53209.

SELL: HQ170A with manual. WA7DKZ; 508 Clark, Laramie, WY 82070.

WANTED, PANADAPTER. Send info. & price Bob Somers, W2QYH, 411 Hamilton Rd., Glassboro, NJ 08028, 609-881-2826.

Coming Events

THE MANKATO AREA RADIO CLUB will hold its picnic, auction, and swapfest on July 31, between 10:00 A.M. and 4:00 P.M. at Spring Lake Park in North Mankato, MN. Talk-in on 3.93, 146.94, and 146.25/85 MHz. Bring your own dinner. For further info write to Mankato ARC, Box 1961, North Mankato, MN 56001, or call Allen Windhorn at (507) 931-1349.

COME TO CANADA this summer for Ontario Hamfest 77. July 8-10 1977, sponsored by Burlington Amateur Radio Club. Weekend camping, fleamarket, auction, many displays. Write Box 836 Burlington Ont. L7R3Y7 for descriptive brochure.

2ND ANNUAL ERIE HAM JAM Sunday, September 25, 1977. Rainbow Gardens — Waldameer Park. Door prizes — Flea Market — Forums — Large indoor facilities. Write for more info, Radio Assoc. of Erie, P.O. Box 844, Erie, PA 16512.

MEMPHIS IS BEAUTIFUL IN OCTOBER! The Memphis ARRL-sponsored Hamfest, bigger and better than the 4,500 who attended last year, will be held at State Technical Institute, Interstate 40 at Macon Road, on Saturday and Sunday October 1 and 2. Demonstrations, displays, MARS meetings, flea market, ladies flea market, too! Hospitality room, informal dinners, XYL entertainment, many outstanding prizes. Dealers and Distributors welcome. Contact Harry Simpson W4SCF, PO Box 27015, Memphis, TN 38127 for further information.

DELTA QSO PARTY All amateurs are invited to participate in the eighth annual Delta QSO Party which is sponsored by the Delta Division of the American Radio Relay League. Contacts must take place from 1800Z Sept. 24 to 2400Z Sept. 25, 1977. No time or power restrictions. SASE for details from W5RUB.

NEW YORK CITY. The 4th annual Hall of Science A.R.C. Flea Market and Hamvention. 111th Street and 48th Ave. Corona-Queens. Refreshments, Zoo Museum. Dealers booths, test bench. Family fun Sunday June 12 (rain date June 19) 9 AM-3. Admission $2.00 door prizes. Talk in .34/94. Information (212) 699-9400.

BETTER THAN EVER—1977 EDITION Golden Spread Hamfest and Flea Market- Holiday Inn West Amarillo, Texas Aug. 12, 13 & 14. Six big tech sessions. Commercial exhibits. Family recreation. Two Hospitality Hours. Big pre-registration prize and super Grand Prize, others. $3.00 advance, $4.00 at door. For info. pre-registration packet, P.O. Box 10221, Amarillo, Texas 79106.

NORTHEAST WISCONSIN SWAP FEST — Neenah, Wisconsin, Sunday, July 24, 1977, 0900 — 1500. Talk-in on .94/.94 — Admission $1.50 — Tables $2.00. Send advance reservations to 3-F's, Box 1032, Neenah, Wisconsin 54956.

More Details? CHECK — OFF Page 126

7400N TTL

SN7400N	.16	SN7459A	.25		
SN7401N	.16	SN7460N	.22	SN74154N	1.00
SN7402N	.21	SN7470N	.45	SN74155N	.99
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SN7404N	.18	SN7473N	.37	SN74157N	.99
SN7405N	.24	SN7474N	.32	SN74160N	1.25
SN7406N	.20	SN7475N	.50	SN74161N	.99
SN7407N	.29	SN7476N	.32	SN74163N	.99
SN7408N	.25	SN7479N	5.00	SN74164N	1.10
SN7409N	.25	SN7480N	.50	SN74165N	1.10
SN7410N	.18	SN7482N	.98	SN74166N	1.25
SN7411N	.30	SN7483N	.70	SN74167N	5.50
SN7412N	.33	SN7485N	.89	SN74170N	2.10
SN7413N	.45	SN7486N	.39	SN74172N	8.95
SN7414N	.70	SN7488N	3.50	SN74173N	1.50
SN7416N	.35	SN7489N	2.49	SN74174N	1.25
SN7417N	.35	SN7490N	.45	SN74175N	.99
SN7420N	.21	SN7491N	.75	SN74176N	.90
SN7421N	.33	SN7492N	.49	SN74177N	.90
SN7422N	.49	SN7493N	.49	SN74180N	.99
SN7423N	.37	SN7494N	.79	SN74181N	2.49
SN7425N	.29	SN7495N	.79	SN74182N	.95
SN7426N	.29	SN7496N	.89	SN74184N	1.95
SN7427N	.37	SN7497N	4.00	SN74185N	2.20
SN7429N	.42	SN74100N	1.00	SN74186N	15.00
SN7430N	.26	SN74107N	.39	SN74187N	6.00
SN7432N	.31	SN74121N	.39	SN74188N	3.95
SN7437N	.27	SN74122N	.39	SN74190N	1.19
SN7438N	.27	SN74123N	.50	SN74191N	1.25
SN7439N	.25	SN74125N	.60	SN74192N	.89
SN7440N	.15	SN74126N	.60	SN74193N	.89
SN7441N	.89	SN74132N	1.09	SN74194N	1.25
SN7442N	.59	SN74136N	.95	SN74195N	.75
SN7443N	.75	SN74141N	1.15	SN74196N	1.25
SN7444N	.75	SN74142N	4.00	SN74197N	.75
SN7445N	.75	SN74143N	4.50	SN74198N	1.75
SN7446N	.81	SN74144N	4.50	SN74199N	1.75
SN7447N	.69	SN74145N	1.15	SN74200N	5.59
SN7448N	.79	SN74147N	2.35	SN74279N	.90
SN7450N	.26	SN74148N	2.00	SN74251N	1.79
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CD4016	.56	CD4051	2.95	74C151	2.90
CD4017	1.35	CD4053	2.95	74C154	4.00
CD4019	.55	CD4060	3.25	74C157	2.15
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CD4022	1.25	CD4069	.45	74C161	3.25
CD4023	.25	CD4071	.45	74C163	3.00
CD4024	1.50	CD4081	.45	74C164	3.25
CD4025	.25	CD4511	2.50	74C173	2.60
CD4027	.69	CD4518	2.50	74C193	2.75
CD4028	1.65	MC14566	3.00	74C195	2.75
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LM307CN	.35	LM380N	1.39	LM2901N	2.95
LM308H	1.00	LM380CN	1.05	LM3065N	.69
LM308CN	1.00	LM381N	1.79	LM3900N	.55
LM309H	1.10	LM382N	1.79	LM3905N	.60
LM309K	.99	NE501K	8.00	LM3909	1.25
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LM311H	.90	NE531H	3.00	MC5558V	1.00
LM311N	.90	NE536T	6.00	LM7525N	.90
LM318CN	1.50	NE540L	6.00	LM7535N	1.25
LM319N	1.30	NE550N	.79	8038B	4.95
LM320K-5	1.35	NE555V	.39	LM75450	.49
LM320K-5.2	1.35	NE560B	5.00	75451CN	.39
LM320K-12	1.35	NE561B	5.00	75452CN	.39
LM320K-15	1.35	NE562B	5.00	75453CN	.39
LM320T-5	1.75	NE565H	1.25	75454CN	.39
LM320T-5.2	1.75	NE565N	1.75	75491CN	.79
LM320T-8	1.75	NE566CN	1.25	75492CN	.89
LM320T-12	1.75	NE567H	1.95	75494CN	.89
LM320T-15	1.75	NE567V	1.50	RCA LINEAR	
LM320T-18	1.75	LM703CN	.45	CA3013	2.15
LM320T-24	1.75	LM709H	.29	CA3023	2.56
LM323K-5	9.95	LM709N	.29	CA3035	2.48
LM324N	1.80	LM710N	.79	CA3039	1.35
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74LS02	.29			74LS151	1.55
74LS03	.29	74LS74	.49	74LS153	1.89
74LS04	.35	74LS75	.69	74LS157	1.55
74LS05	.35	74LS76	.49	74LS162	2.25
74LS08	.29	74LS83	1.75	74LS163	2.25
74LS10	.29	74LS85	2.49	74LS164	1.95
74LS13	.69	74LS86	.49	74LS175	1.95
74LS14	1.75	74LS90	1.25	74LS181	3.69
74LS20	.29	74LS92	1.25	74LS190	2.85
74LS26	.39	74LS93	1.25	74LS191	2.85
74LS27	.39	74LS95	1.95	74LS192	2.85
74LS28	.39	74LS96	1.89	74LS193	2.85
74LS30	.79	74LS107	.59	74LS194	1.89
74LS32	.39	74LS109	.59	74LS195	1.89
74LS40	.39	74LS112	.59	74LS257	1.75
74LS51	.29	74LS132	1.25	74LS260	.55
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XC209	Green	4/$1	XC111	Green	4/$1
XC209	Orange	4/$1	XC111	Yellow	4/$1
XC209	Yellow	4/$1	XC111	Orange	4/$1

.200" dia.			.185" dia.			.200" dia.		
XC22	Red	10/$1	XC526	Red	5/$1	XC556	Red	10/$1
XC22	Green	4/$1	XC526	Green	4/$1	XC556	Green	4/$1
XC22	Yellow	4/$1	XC526	Yellow	4/$1	XC556	Yellow	4/$1
XC22	Orange	4/$1	XC526	Orange	4/$1	XC556	Orange	4/$1
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SPECIAL * — XC556 Red 100/$8.00 1000/$60.00 — SPECIAL *

DISPLAY LEDS

DL707 DL338

TYPE	POLARITY	HT		TYPE	POLARITY	HT	
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MAN 2	5 x 7 Dot Matrix	300	4.95	MAN 4710	Common Anode-Red	400	1.95
MAN 3	Common Cathode	125	3/1.00	DL701	Common Anode-red ±	300	.99
MAN 4	Common Cathode	187	1.95	DL704	Common Cathode	300	.99
MAN 7	Common Anode	300	1.25	DL707	Common Anode	300	.99
MAN 7G	Common Anode-green	300	1.95	MAN 4740	Common Anode-Red	400	.99
MAN 7Y	Common Anode-yellow	300	1.95	DL741	Common Anode	600	1.50
MAN 52	Common Anode-green	300	.99	DL 747	Common Anode	600	2.25
MAN 64	Common Anode-red	400	.99	DL 750	Common Cathode	600	2.49
MAN 74	Bommon Cathode	300	1.50	DL 338	Common Cathode	110	.50
MAN 82	Common Anode-yellow	300	.99	FND70	Common Cathode	250	.75
MAN 84	Common Cathode-yellow	300	.99	FND503	Common Cathode	500	1.00
MAN 3620	Common Anode -orange	300	1.75	FND507	Common Anode	500	1.00

FACTORY REJECTS ATARI GAME BOARDS

BOARD A — 8½" x 16"
Over 60 each reusable IC's
Misc. Transistors, Resistors, Diodes, Caps, Crystals, Switch, etc.
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HP 5082-7300 Multi-Digit Series

- ⅛" Ht. • Common Cathode • Dip Package
- 3 to 5 volts @ 5 mils per segment
- 7 segment Monolithic • Red Display

2 Digit	$ .79
3 Digit	.89
4 Digit	.99
5 Digit	1.19

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

Plastic Push Button Switch

- 18 AWG Solid Wire - 5" Long
- .50 (wide) X .60 (high) ⅜-27 Thread
- 8 AMP @ 14 Volt - 1 AMP @ +10 Volt

		1-9	10-Up
J-188-1	Push On-Push Off	59	49
J-188-2	Normally Open	59	49
J-188-3	Normally Closed	59	49

MINATURE TOGGLE SWITCH

JMT-221	DPDT	on/off/on	$1.95
JMT-223	DPDT	on/none/on	$1.75
JMT-121	SPDT	on/off/on	$1.50
JMT-123	SPDT	on/none/on	$1.25

CLIPLITE 8/$1.49

LED MOUNTING SYSTEM use with XC556 LEDS
Specify Colors — Red - Green - Amber - Yellow

DIP SWITCHES

SPST Slide Action
#206-4 (8 pin dip) 4 switch unit $1.75 ea.
#206-7 (14 pin dip) 7 switch unit $1.95 ea.
#206-8 (16 pin dip) 8 switch unit $2.25 ea.

50 PCS. RESISTOR ASSORTMENTS $1.75 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.

ASST. 8R Includes Resistor Assortments 1-7 (350 PCS.) **$10.95 ea.**

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HOBBY-WRAP TOOL-BW-630

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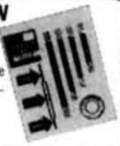
- Weighs ONLY 11 Ounces
- Wraps 30 AWG Wire onto Standard DIP Sockets (.025 inch)
- Complete with built-in bit and sleeve

$34.95
(batteries not included)

WIRE-WRAP KIT — WK-2-W
WRAP • STRIP • UNWRAP



- Tool for 30 AWG Wire
- Roll of 50 Ft. White or Blue 30 AWG Wire
- 50 pcs. each 1", 2", 3" & 4" lengths — pre-stripped wire

$11.95

WIRE WRAP TOOL WSU-30
WRAP • STRIP • UNWRAP - $5.95

WIRE WRAP WIRE — 30 AWG

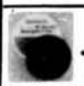
25 ft. min. $1.25 50 ft. $1.95 100 ft. $2.95 1000 ft. $15.00
SPECIFY COLOR — White - Yellow - Red - Green - Blue - Black

CUTTER CRIMPER TOOL (CS-8)

- Plier Nose (serrated-jaw)
- Scissors Action Cutting
- 6 Bolt Cutters (4-40, 5-40, 6-32, 8-32, 10-32, 10-24)
- Crimp Stations (7mm Auto — 22-20 to 12-X0 elect.)
- "Up-Front" Wire Cutting
- Scissors Action Stripping (No. 22-20 to No. 10)
- Crimp Stations — insulated (2w-20 to 12-10 elect.)

Actual Size - 8¼" length **$8.50**

Permacel Electrical Tape



- ¾" (wide) X 66 ft. (long) • All weather • Not import

$1.25 per roll — $9.95 per 10 roll package

ZENERS — DIODES — RECTIFIERS

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

SCR AND FW BRIDGE RECTIFIERS

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

TRANSISTORS

MPS A05	5/$1.00			PN4249	4/$1.00
MPS A06	5/$1.00			PN4250	4/$1.00
2N2219A	3/$1.00	PN3567	3/$1.00	2N4400	4/$1.00
2N2221	4/$1.00	PN3568	4/$1.00	2N4401	4/$1.00
2N2222A	5/$1.00	PN3569	4/$1.00	2N4402	4/$1.00
2N2369	5/$1.00	2N3704	5/$1.00	2N4403	4/$1.00
2N2369A	4/$1.00	2N3705	5/$1.00	2N4409	5/$1.00
2N2484	4/$1.00	2N3706	5/$1.00	2N5086	4/$1.00
2N2906A	4/$1.00	2N3707	5/$1.00	2N5087	4/$1.00
2N2907A	5/$1.00	2N3711	5/$1.00	2N5088	4/$1.00
2N2925	5/$1.00	2N3724	$ .65	2N5089	4/$1.00
2N3053	2/$1.00	2N3725	$1.00	2N5129	5/$1.00
2N3055	$ .89	2N3903	5/$1.00	2N5138	5/$1.00
MJE3055	$1.00	2N3904	4/$1.00	2N5139	5/$1.00
MJE2955	$1.25	2N3905	4/$1.00	2N5209	5/$1.00
2N3392	5/$1.00	2N3906	4/$1.00	2N5951	5/$1.00
2N3398	5/$1.00	2N4013	3/$1.00	C106B1SCR	2/$1.00
		2N4014	3/$1.00	2N5432	$2.00
		2N4123	10/$1.00		

CAPACITOR CORNER

50 VOLT CERAMIC DISC CAPACITORS

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

100 VOLT MYLAR FILM CAPACITORS

.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

1977 HAWKEYE HAM & COMPUTERFEST August 20 and 21 sponsored by Des Moines Amateur Radio Association. WBØURB for details.

LEXINGTON, KY Bluegrass Amateur Radio Club Central Kentucky Hamfest August 14, 1977 at the Lexington National Guard Armory located adjacent to the Bluegrass Field on Airport Road, Lexington, Kentucky. Grand Prizes, Hourly Door Prizes, Manufacturer's Exhibits, indoor/outdoor Fleamarket, Guest Speakers and Forums. P.O. Box 4411, Lexington, KY 40504 for details.

NEW ORLEANS HAMFEST/COMPUTERFEST at the Hilton Inn in Kenner, LA, September 24 & 25. Information on tickets, room reservations and etc., will be furnished upon request by contacting the New Orleans Hamfest/Computerfest; P. O. Box 10111, Jefferson, LA 70181.

THE NORTHERN BERKSHIRE Amateur Radio Club Hamfest, July 9th and 10th Cummington Fair Grounds, Cummington Mass. Free overnight camping. Tech talks, Demos, and dealers. Flea Market $1. Admission $3 with XYL $5 advanced $2 and $4, for information write Hildy Sheerin WA1ZNE, 79 Greylock Ter., Pittsfield, Ma. 01201.

MISSOURI: Second Annual Hamfest and Communications Show, Sunday, July 24, 1977 at the Slater Park with fly-in facilities. Registration $1.00 in advance: $1.50 at the door. Sunday noon meal for a nominal fee. Flea markets, XYL activities. For information and advance tickets write Dale Beilsmith, WØKNF, Box 74, Slater, Missouri 65349, (816) 529-2173.

MARAC — ICHN 9th Annual County Hunters Convention June 30th — July 3rd, 1977. Rochester Holiday Inn downtown, 220 South Broadway on Highway 63, Rochester, Minnesota.

NEW JERSEY QSO PARTY — August 20-22. The Englewood Amateur Radio Association invites amateurs the world over to take part in this 18th Annual QSO party. Send SASE to the Englewood A.R.C., 303 Tenafly Road, Englewood, NJ 07631 for full rules.

TOTAH AMATEUR RADIO CLUB field day. July 30 & 31, 1977 at our Corners National Monument (NM, AZ, UT & CO), Mode — SSB, CW. Bands 15, 20 & 40 meters. All contacts confirmed upon request. For further information write R. Keith Miller, Totah Amateur Radio Club, Box 1991, Farmington, NM 87401.

ROME HAM FAMILY DAY — June 5, 1977. Bring the family to the Beeches, Rt. 26, Rome, NY for a full day of events. We have a giant outdoor flea market and over 5,000 sq. ft. of indoor display area. There will be exhibits, contests, new equipment displays, technical presentations and a demonstration of French cooking for the XYL. For info write PO Box 721, Rome, NY 13440.

AKRON, OH Goodyear Amateur Radio Club Hamfest and Family Picnic June 12, 1977 at Wingfoot Lake Park, 10 AM 6 PM. The park is southeast of Akron on County Road 87 near Route 43. For details — Advance Donations — Map — Program — etc. Contact Don Rogers, WA8SXJ, 161 S. Hawkins Ave., Akron, Ohio 44313. Phone: 216-864-3665.

THE WOOD COUNTY ARC ANNUAL Ham-A-Rama is Sunday, July 17, from 8:00 A.M. to 5:00 P.M., at the County Fairgrounds in Bowling Green, Ohio (about 25 miles south of Toledo). Free parking and admission. Donation $1.50 advance, $2.00 at the door. Tables $2.00. Talk-in on 146.52. Refreshments available. Write WCARC, 7929 Rudolph Road, Rudolph, Ohio 43462.

TRI-CLUB HAMFEST July 13, 1977 from 8:00 AM to 5:00 PM at Allentown Police Academy Pistol Range in Lehigh, Parkway South at Allentown, PA. Admission $1.00 to all, including sellers. Children free. Talk-in is .34-.94 and .52. For further information contact Fred Hermann, W3HYT at RD 1, Box 104, Emmaus, PA 18049.

HARRISBURG ARC HAMFEST — Monday July 4. Indian Echo Park, 2 miles West of Hershey. Admission $3.00. Sell no charge.

MCKEESPORT, PA Two Rivers Amateur Radio Club Hamfest Sunday, July 17, 1977 at Green Valley Fire Department fairgrounds off U.S. Route 30 near East McKeesport. Big Flea Market.

PENN CENTRAL HAMFEST June 5, 1977 at new location in Allenwood, Penna., rain or shine. Flea market, auction and contests. Registration $3.00 XYL and children free. Talk-in on 3940 MHz., 146.37-97, 146.13-73 and 52 simplex. For information contact Clair Yeagle WA3QXI, R.D.#1 P.O. Box 224, Turbotville, Pa. 17772. Phone 717-437-2595.

More Details? CHECK—OFF Page 126

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flea market

MCKEESPORT, PENNSYLVANIA: The Two Rivers Amateur Radio Club will hold its 13th annual Hamfest on Sunday, July 17, 1977, at the Green Valley Fire Dept. grounds off U.S. Route 30 near East McKeesport. Check-ins on 52/52 and 22/82. For information write Andrew Salitros, W30FM, 2901 Stewart Street, McKeesport, Pennsylvania 15132.

OKANAGAN INTERNATIONAL HAMFEST — GALLAGHER LAKE KOA — 8 miles North of Oliver, BC. Prizes, Games, Contests — Family Fun — Gabfest — Bring your Crafts Ladies Flea Market — POTLUCK DINNER on the SUNDAY AT NOON. For more info. contact: VE7ALV-Phil Wilkinson, Naramata, B.C. Canada.

ATLANTA HAMFESTIVAL and 1977 Georgia ARRL convention will be held June 18-19th in Atlanta at the Downtown Marriott Hotel. (Rooms: $18 single, $24 double). A pre-registration package will be mailed on May 1st to all who have attended the HamFestival within the past three years; if you have not received a package by May 10th, you may write to Atlanta HamFestival, 53 Old Stone Mill Road, Marietta, GA, 30067 or call area 404/971-HAMS anytime day or night. Doors open at 9AM, Saturday June 18th and Sunday June 19th. Contact the hotel directly toll-free at 1-800-228-9290 for reservations — and hurry!!

THE SIXTH ANNUAL INDIANAPOLIS HAMFEST will be held on Sunday July 10, 1977 at the Marion County Fairground. Gate admission $2, which includes in & out privileges, hourly prize drawings, and main prize drawing. Prizes will include complete low band transceiver, 2-meter synthesizer. & the possibility of low & high band beam antennas. Also large indoor flea market. Electricity available at most indoor flea market spaces. For further info write Indianapolis Hamfest, P.O. Box 1002, Indianapolis, IN 46206.

INDIANA LAKE COUNTY AMATEUR RADIO CLUB'S hamfest June 19th at the Isaac Walton League in Portage, Indiana. (Take I-94 to Ind. 249 exit, then north on Ind. 249 ½ mile) Tickets: $1.50 advance, $2.00 at gate. Write Herbert S. Brier W9AD (W9EGQ), 409 S. 14th St., Chesterton, Ind. 46304.

SOUTH MILWAUKEE AMATEUR Radio Club Swapfest "77" will be held Saturday, July 9, 1977 at Shepard Park (American Legion Post #434), 9327 South Shepard Avenue, Oak Creek, Wisconsin. Activities begin at 7:00 AM and run till 5:00 PM. Parking, picnic area, hot and cold sandwiches and liquid refreshments available on grounds. Overnite camping available. Admission $1.00. And includes a "Happy Hour" with free beverages. Prizes will be awarded. Talk-in on 146.94 MHz. FM. More details (inc. map) from: South Milwaukee Amateur Radio Club Inc., S. F. Schreiter, W9AKF, Sec., 104 Brookdale Drive, South Milwaukee, WI 53172.

MADISON, WI SWAPFEST. Rain or Shine Sunday, June 19. Dane Co. Expo Center Youth Building. Doors open 8 AM. Bring the whole family for delicious food and entertainment.

SPOKANE RADIO AMATEUR COUNCIL (SARC) Hamfest, July 16th & 17th in the Pence Union Blvd. at Eastern Wash. State College, Cheney WA. About 15 miles West of Spokane. Pre-registration is $9.50, includes banquet, seminars, flea market, XYL activities. Make pre-registration checks payable to SARC and send to Jim Johnson, WA7BWO, Drawer "A", Cheney, WA 99004.

HAM HOLLER-IN AWARD: Issue for contacting A.R.S. NC4NHC. Date-18 June 77. Time: 9 A.M. to 9 P.M. E.S.T. Frequencies: 14280-7280-3980. NC4NHC will be operating from The National Hollerin Contest grounds located at Spivey's Corner, N.C. The theme is "Communication- The Old and The New." The award is sponsored by The Cape Fear Amateur Radio Society, Methodist College, Box M-618, Fayetteville, N.C. 28301. The certificate will be issued to stations who make contact with NC4NHC and send $1.00 (postage and handling fee) to the Cape Fear Amateur Radio Society prior to 1 Sept. 1977.

Stolen Equipment

ICOM 230 SN 4889 inscribed on back — Fred WA4OFR SS#231 56 0569 — in Lanham, Md. March 13. Contact WA4OFR at 301-459-0798 with any information.

STOLEN from my car on February 28, 1977, Clegg FM 27B s/n sometime between 1:30 and 5:00 PM. Car was parked at 571 Fairfield Ave., Bridgeport, CT. Contact Robert H. Horen, 571 Fairfield Ave., Bridgeport, CT 06603 with any information.

More Details? CHECK—OFF Page 126

State of the art by K.V.G.

CRYSTAL FILTERS and DISCRIMINATORS

9.0 MHz FILTERS

XF9-A	2.5 kHz	SSB TX	$31.95
XF9-B	2.4 kHz	SSB RX/TX	$45.95
XF9-C	3.75 kHz	AM	$48.95
XF9-D	5.0 kHz	AM	$48.95
XF9-E	12.0 kHz	NBFM	$48.95
XF9-M	0.5 kHz	CW (4 pole)	$34.25
XF9-NB	0.5 kHz	CW (8 pole)	$63.95

9.0 MHz CRYSTALS (Hc25/u)

XF900	9000.0 kHz	Carrier	$4.00
XF901	8998.5 kHz	USB	$4.00
XF902	9001.5 kHz	LSB	$4.00
XF903	8999.0 kHz	BFO	$4.00
F-05	Hc25/u Socket		.50

Export Inquiries Invited

Shipping $1.25 per filter

ALSO AVAILABLE FROM KVG
10.7 MHz CRYSTAL FILTERS
OSCILLATOR CRYSTALS 50 kHz TO 150 MHz
Write for Details

432 MHz SSB TRANSVERTER, MMt432-28

Use your HF Transceiver on the 432 MHz band with the addition of the MMt432 linear Transverter. The MMt432 operates on all modes; SSB, CW, AM, FM. It contains BOTH the linear transmit up-converter and the receive down-converter. An internal PIN diode T/R connects to your Transceiver T/R line. The MMt432 is FT101 and similar HF rig compatible. Add the 70/MBM48 MULTIBEAM and operate direct into OSCAR 7 mode B. Write for application note.

Specifications:

Output Power	10 W PEP
Drive, 10 meters	½ W max
Receiver N.F.	3.0dB typ
Receiver gain	30 dB typ
Bandwidth	2 MHz
Prime Power	12 V D.C.
Price	**$239.95**

Shipping: $3.50

ALSO AVAILABLE:

MMt432-28 MK4	$254.95
MMt432-50	$249.95
MMt438-ATV	$274.95
QMt432-144	$164.95
FMt440-146	$164.95
MMt144-28	$199.95

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LINEAR AMPLIFIERS and TRANSVERTERS
by POLAR ELECTRONIC DEVELOPMENTS

Specifications: EDL144

Drive Power	20 W PEP max
Output Power	100 W PEP max
Rx Pre-Amp Gain	20 dB typ
N.F.	2.5 dB typ
Power Supply	115 V A.C.
Size:	10″ x 6″ x 7″

The EDL144 amplifier contains a high power transmit linear amplifier (5894 PA) and power supply (115v) together with a low noise receive pre-amplifier (2.5 dB NF). T/R switching is automatic by an internal VOX circuit; no changes are needed to your existing transceiver.

Specifications: EDL432P

Drive Power	10 W PEP max
Output Power	50 W PEP max
Power Supply	115 V A.C.
Size:	10″ x 6″ x 11″

The EDL432P amplifier contains a high power triode amplifier (2C39A) with matching power supply (115v). The cabinet also contains the cooling air blower, antenna relays and full metering. The RF section is also available as a complete sub-assembly, model EDL432, for use with an existing power supply etc.

Use your 10 meter transceiver with the EDT50-28 or EDT144-28 transverters to operate on the 6M or 2M bands. These transverters operate in all modes; they have the same style P.A. design as the EDL144 amplifier. Receiving is with a MMc50 or MMc144 style converter mounted inside the cabinet.

Specifications:

EDT50-28	50-52 MHz
EDT144-28	144-146 MHz
Drive Power, 10M	0.5 W max
Output Power	100 W PEP max
Rx Gain	30 dB typ
N.F.	2.5 dB typ
Size:	10″ x 6″ x 7″

An external power supply is required.

Write for full details and application notes. Visit us at Atlanta June 18, 19.

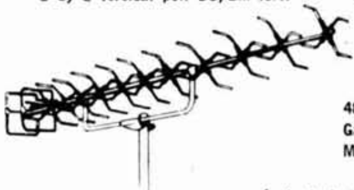
ANTENNAS
144-148 MHz

Model 8XY/2M	$39.95

Gain 9.5dB in each plane, 50Ω feed.

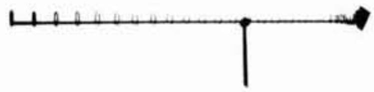
For OSCAR communications add Circular polarization harness, PMH/2C	$9.70
Also available for 2 meters:	
8 over 8 J-Slot, Model D8/2M	$39.95
8 by 8 vertical pol. D8/2M-vert.	$48.70

for CW AM FM SSB OSCAR



420-450 MHz

48 element J-Beam MULTIBEAM
Gain +15.7 dBd. Feed 50Ω coaxial.
Model 70/MBM48 $49.95



1296 — LY
1296 MHz LOOP-YAGI
GAIN +20dBi
FEED 50Ω COAXIAL
$54.45

1250-1340 MHz

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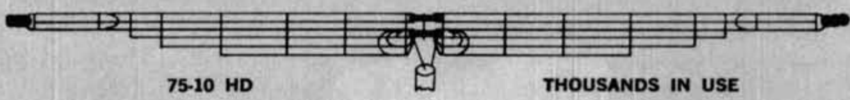

75-10 HD

THOUSANDS IN USE

MOR-GAIN HD DIPOLE SPECIFICATIONS

MODEL	BANDS (Meters)	LENGTH (feet)	PRICE
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75-40 HD	75/40	66	55.00
75-40 HD (SP)	75/40	66	57.50
75-20 HD	75/40/20	66	66.50
75-20 HD (SP)	75/40/20	66	66.50
75-10 HD	75/40/20/15/10	66	74.50
75-10 HD (SP)	75/40/20/15/10	66	74.50
80-10 HD	80/40/20/15/10	69	76.50

NOTE: 75 meter models are factory tuned to resonate at 3950 KHz. (SP) models are factory tuned to resonate at 3800 KHz. 80 meter models are factory tuned to resonate at 3650 KHz.

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- Lowest cost/highest performance antenna on the market today.
- Highest performance for the Novice as well as the Extra-Class Op.
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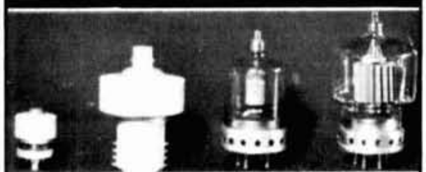

4CX150	4CX1000	4-65	4-250
4CX250	4CX1500	4-125A	4-400
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4CX350A	4CX5000		304TL
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1N914	100v	10mA	.05
1N4004	400v	1A	.08
1N4005	600v	1A	.08
1N4007	1000v	1A	.15
1N4148	75v	10mA	.03
1N753A	6.2v	z	.25
1N758A	10v	z	.25
1N759A	12v	z	.25
1N4733	5.1v	z	.25
1N5243	13v	z	.25
1N5244B	14v	z	.25
1N5245B	15v	z	.25

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8-pin	pcb	.25	ww	.45
14-pin	pcb	.25	ww	.40
16-pin	pcb	.25	ww	.40
18-pin	pcb	.25	ww	.75
22-pin	pcb	.45	ww	1.25
24-pin	pcb	.35	ww	1.25
28-pin	pcb	.35	ww	1.45
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2N3740 PNP 1A 60v	.25
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2N3054 NPN	.35
2N3055 NPN 15A 60v	.50
T1P125 PNP Darlington	.35
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4012	.20
4013	.40
4014	1.10
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4017	1.10
4018	1.10
4019	.70
4020	.85
4021	1.35
4022	.95
4023	.25
4024	.75
4025	.35
4026	1.95
4027	.50
4028	.95
4030	.35
4033	1.95
4034	2.45
4035	1.25
4040	1.35
4041	.69
4042	.95
4043	1.25
4044	.95
4046	1.50
4049	.80
4050	.60
4066	1.35
4069	.40
4071	.35
4082	.45

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7400	.15
7401	.15
7402	.20
7403	.20
7404	.15
7405	.25
7406	.35
7407	.55
7408	.25
7409	.15
7410	.10
7411	.25
7412	.30
7413	.45
7414	1.10
7416	.25
7417	.40
7420	.15
7426	.30
7427	.45
7430	.15
7432	.30
7437	.35
7438	.35
7440	.25
7441	1.15
7442	.55
7443	.85
7444	.45
7445	.80
7446	.95
7447	.95
7448	.95
7450	.25
7451	.25
7453	.20
7454	.25
7460	.40
7470	.45
7472	.45

7473	.25
7474	.35
7475	.35
7476	.30
7480	.55
7481	.75
7483	.95
7485	.95
7486	.30
7489	1.35
7490	.55
7491	.95
7492	.95
7493	.40
7494	1.25
7495	.60
7496	.80
74100	1.85
74107	.35
74121	.35
74122	.55
74123	.55
74125	.45
74126	.35
74132	1.35
74141	1.00
74150	1.00
74151	.75
74153	.95
74154	.75
74156	1.15
74157	.65
74161	.85
74163	.95
74164	.60
74165	1.50
74166	1.35
74175	.80

74176	1.25
74180	.85
74181	2.75
74182	.95
74190	1.75
74191	.35
74192	1.65
74193	.85
74194	1.25
74195	.95
74196	1.25
74197	1.25
74198	2.35
74221	1.00
74367	.85
75108A	.35
75110	.35
75491	.50
75492	.50
74H00	.25
74H01	.25
74H04	.25
74H05	.25
74H08	.35
74H10	.35
74H11	.25
74H15	.30
74H20	.30
74H21	.25
74H22	.40
74H30	.25
74H40	.25
74H50	.25
74H51	.25
74H52	.15
74H53J	.25
74H55	.25

74H72	.55
74H101	.75
74H103	.75
74H106	.95
74L00	.35
74L02	.35
74L03	.30
74L04	.35
74L10	.35
74L20	.35
74L30	.45
74L47	1.95
74L51	.45
74L55	.65
74L72	.45
74L73	.40
74L74	.45
74L75	.55
74L93	.55
74L123	.55
74S00	.55
74S02	.55
74S03	.40
74S04	.35
74S05	.35
74S08	.35
74S10	.35
74S11	.35
74S20	.35
74S40	.25
74S50	.25
74S51	.45
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74S112	.90
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74S133	.45
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74LS00	.45
74LS01	.45
74LS02	.45
74LS04	.45
74LS05	.55
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74LS10	.45
74LS11	.45
74LS20	.40
74LS21	.25
74LS22	.25
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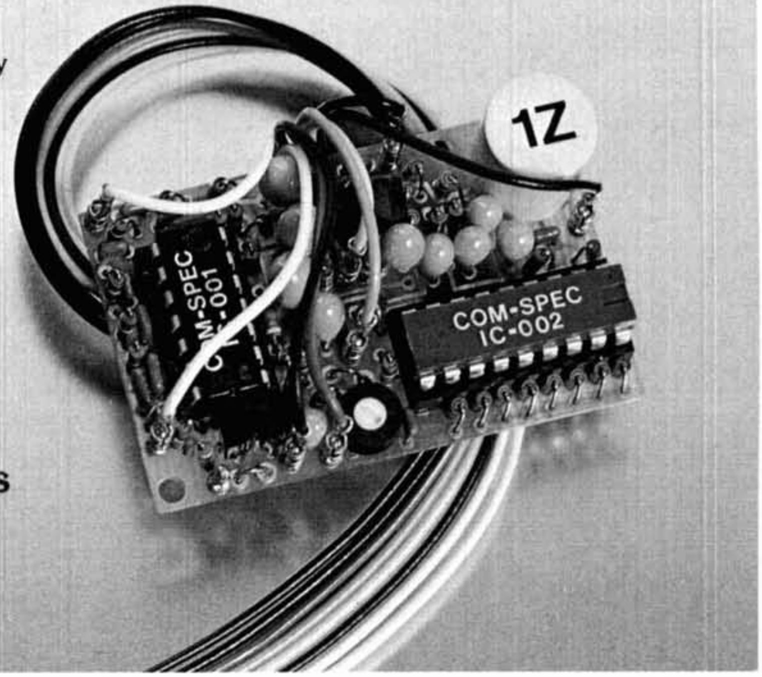

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Our very popular custom models are still available for those of you who have the following radios: Regency BTH BTL, HR2, HR6, Microcom, Aquaphone, MT25 or MT15, Genave Business Amateur and Marine, ICOM 230, 22, 22S, 30A, Midland 13-500 and 13-509, Heathkit HW202 and HW2036, and Standard SR-C806, SR-C826MA.

Porta-Pak started out as a very versatile accessory for the amateur, but, now the word is out and as a result 85% of our sales are now to the business radio services. Among enthusiastic users of the Porta-Pak are the C.A.P., ambulance services, fire departments, Red Cross disaster teams, towing services and private surveilance organizations. The Porta-Pak and VHF Marine radio have provided small craft with reliable communications where it was impractical a short time ago.

Be sure to state the make and model of your transceiver when ordering, and a custom model will be provided if one is made for it. The regular mobile mount holes are used to attach the custom models. No modification of your radio is necessary.

4.5 AMP HR Deluxe Porta-Pak, Universal or Custom Case $67.50 plus $1.75 shipping

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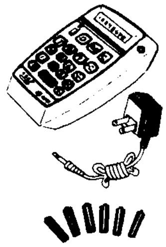

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C22736

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TYPE 1738D 2.7V @ .06A
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MDA962-4	10 Amps	400V	$5.00
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2E26	3.50	4CX250B	20.00	6146	3.00	7645	5.00
3B24WB	3.00	35T	15.00	6146W	4.00	8950	5.50
3B28	4.00	578	3.00	6336A	7.00	4600A	150.00
3D21WB	3.50	811	6.95	6360	5.00	4-125A	15.00
4X150A	13.00	832A	3.00	6528A	7.00		
		5894	25.00	6939	7.00		

I.C.'s							
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MM5262N	2.00	LM340K-24	1.50	74S200L	2.00	LM741CH	4/1.00
C8008-1	5.00	LM337	1.00	MC7908CP	1.00	LM747CH	4/1.00
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		AMS6003	3.00	MC2360G	2.00		

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1N914	30/1.00	1N1202	.40	1N4999	.65	1N5892B	5/1.00
1N1602	.65	1N754A	5/1.00	1N5271B	5/1.00	1N5894B	5/1.00
1N1828	.65	1N3050	1.00	1N5276B	5/1.00	1N5895B	5/1.00
1N2804B	1.00	1N4719	.60	1N5278B	5/1.00	1N4148	30/1.00
		1N4749A	5/1.00	1N5862B	5/1.00	1N823	5/1.00
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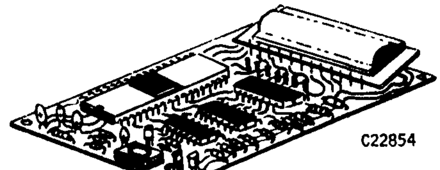

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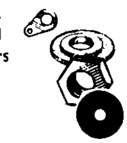

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2N3291	250	10	16DB	A	$ .85
2N3375/MM3375	100	28	11.6W	E	$ 3.50
2N3818	150	28	20W	E	$ 4.50
2N3866	400	28	1W	B	$ .65
2N3866 JAN	400	28	1W	B	$ 2.00
2N3866 JAN TX	400	28	1W	B	$ 3.00
2N3925	175	13.6	5W	D	$ 2.50
2N3948	200	20	1W	B	$ 1.25
2N3950	50	28	50W	E	$10.00
2N4072	175	13.6	.25W	A	$ 1.00
2N4957	1200	10	17DB	A	$ 5.00
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2N5179	900	6	15DB	A	$ .50
2N5589	175	13.6	3W	G	$ 4.60
2N5590	175	13.6	10W	G	$ 6.30
2N5591	175	13.6	25W	G	$10.35
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2N5862	150	27	75W	H	$49.00
2N5942	30	28	80PEP	I	$44.00
2N6081	175	12.5	15W	G	$ 8.50
2N6083	175	12.5	30W	G	$11.25
2N6084	175	12.5	40W	G	$14.95
2N6097	175	12.5	40W	H	$15.00
MM1500	1500	20	250mw	L	$15.00
MM1607/2N5842	1700	4		A	$ 6.00
MM4049	4000	5		A	$ 7.00
MM8006	1000	6	14DB	A	$ 1.25
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2SD235	2/$1.00		
MJE800	$ .65		
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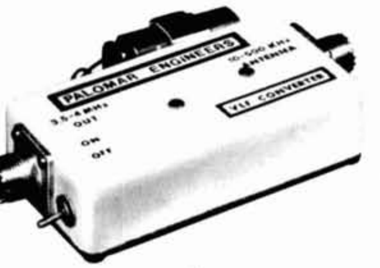
More Details? CHECK—OFF Page 126

VLF CONVERTER



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More Details? CHECK—OFF Page 126

GREAT VALUES Ham Radio's Back Issue Collector's Items

March 1968 (first issue)
FEATURING: 5-band SSB exciter, IC-regulated power supply. Remotely-tuned 10-meter beam, Transistor curve tracer, Double-balanced mixers.

May 1969
FEATURING: Potpourri of integrated-circuit applications, FM repeater receiver performance, RTTY converter, IC noise blanker, The ionospheric e-layer.

August 1969
FEATURING: Homebrew Parabolic Reflector, Solid-state Q-5er, Frequency calibrator with mos IC's, New multiband quad antenna, Troubleshooting with a scope.

September 1969
FEATURING: FM techniques and practices, IC power supplies, 1296-MHz varactor tripler, Tunable bandpass filters, Amateur microwave standards.

October 1969
FEATURING: Hot Carrier Diodes, Low-cost linear IC's, Diversity antennas, solid-state 432-MHz exciter, Tropospheric-duct communications.

November 1969
FEATURING: Op Amps . . . theory, selection & application, WWV receiver, Multi-band antenna, Electronic key, Six-meter collinear.

June 1970
FEATURING: Communication experiments with light emitting diodes, FM modulation standards, Designing phase-shift networks, Transistor frequency multipliers, RTTY frequency-shift meter.

October 1970
FEATURING: An SWR meter for accurate RF power measurements, Direct-conversion receiver, IC voltage regulators, 432 MHz converter, Introduction to thyristors.

December 1970
FEATURING: SSB generator, RF interference, Antenna bridge, QRP transmitter, AFSK oscillator.

June 1971
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FEATURING: A phasing-type single-sideband transmitter, Slim-line touch-tone, UHF prescaler, Crystal oscillators, Noise-figure measurements.

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Iambic squeeze key operation with dot and dash insertion lets you form characters with minimal wrist movement for comfortable, fatique-free sending.

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This NEW MFJ Super Antenna Tuner . . .

matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines. Up to 200 watts RF OUTPUT. Built-in balun, too!

$69.95

With the NEW MFJ Super Antenna Tuner you can run your full transceiver power output — up to 200 watts RF power output — and match your transmitter to **any** feedline from 160 thru 10 Meters whether you have coax cable, balance line, or random wire.

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

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

Increase the usable bandwidth of your mobile whip by tuning out the SWR **from inside your car.** Works great with all solid state rigs (like the Atlas) and with all tube type rigs.

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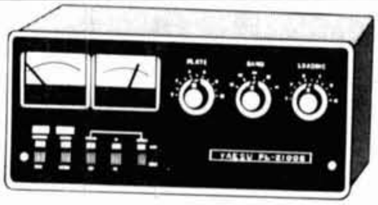

YAESU
FL-2100B linear amplifier
The FL-2100B features: • 2100 watts PEP • Input on 80 to 10 meters • Easy primary voltage change from 117 to 234 V AC • Dual meters allow continuous monitoring of plate current and voltage • Built-in front panel adjustable SWR meter • Drive requirement is 30 to 100W.

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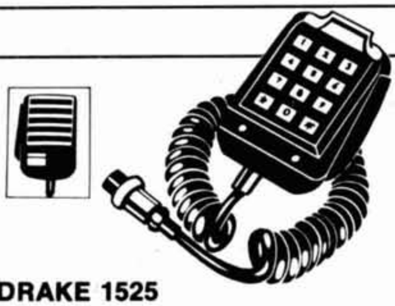

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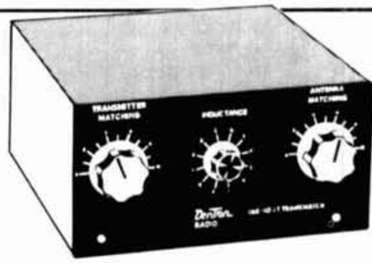

DENTRON
160-10AT super tuner
Features: • Continuous tuning 1.7 to 30 mc • 3 antenna inputs • 500 watts DC, 1000 W PEP • Built-in heavy-duty Balun (3½" dia. x 3" H), 2-core • Harmonic attenuation • Capacitor spacing: 3000 volt.

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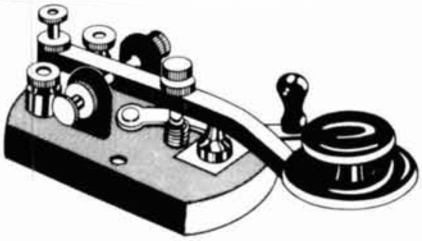

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Speed-x 114-320-003 key
This heady-duty key is constructed on a heavy die-cast base. The hardware is nickel-plated. Has smooth adjustable bearings and heavy-duty coin silver contacts. Black wrinkle base finish, switch and Navy knob.

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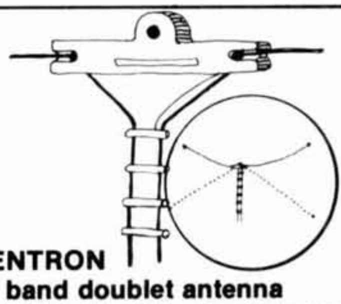

DENTRON
all band doublet antenna
This all band doublet or inverted antenna covers 160 thru 10 meters. Has total length of 130 ft. of 14 ga. stranded copper wire. The doublet is tuned & center fed thru 100 ft. of 470 ohm PVC covered transmission line. Assembly is complete.

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Enjoy Two Meters the economical way! FTV-250 plus your HF Transceiver and you are there with USB, LSB, CW & AM

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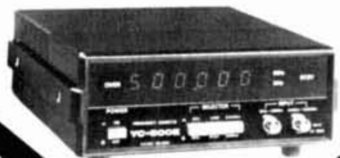

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