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incorporating HAM RADIO HORIZONS

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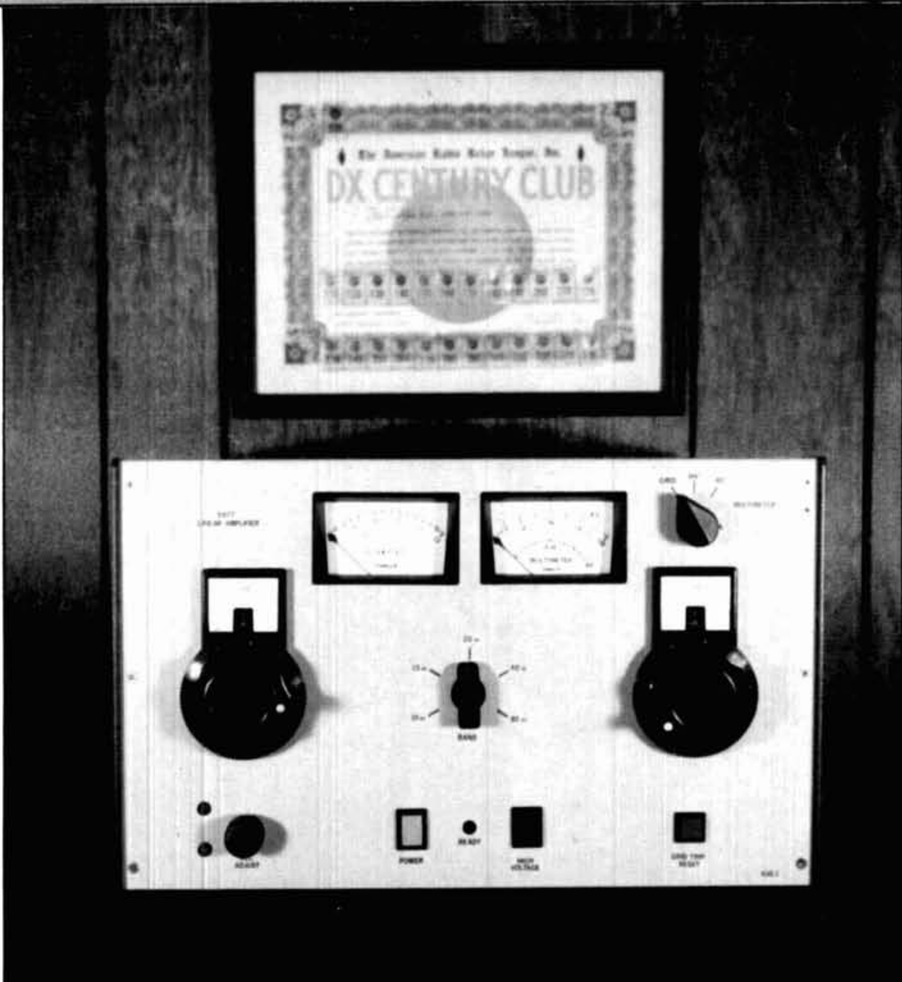
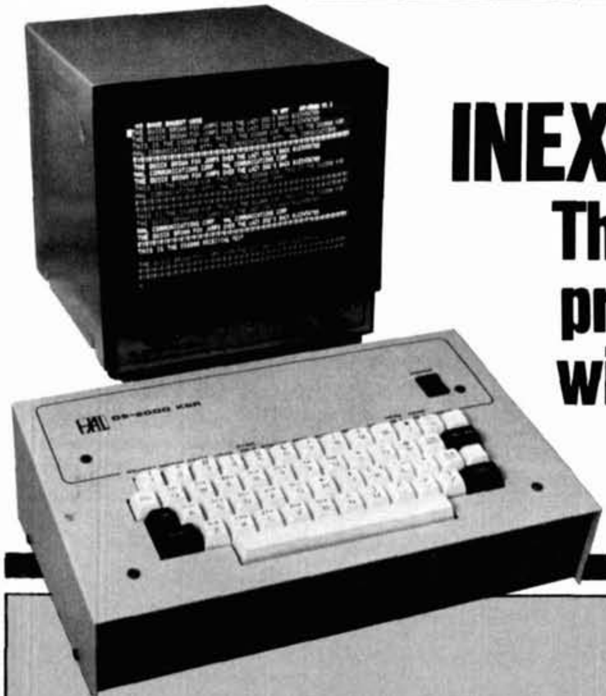


modular linear *amplifier*



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More Details? CHECK—OFF Page 110

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JANUARY 1981
volume 14, number 1

ham radio magazine
is published monthly by
Communications Technology, Inc
Greenville, New Hampshire 03048
Telephone: 603-878-1441

subscription rates
United States: one year, $15.00
two years, $26.00; three years, $35.00
Canada and other countries (via Surface Mail)
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contents

publisher's log

by Skip Tenney

The year 1981 is a year of great change and opportunity here at *ham radio*. We start off by welcoming as new readers the many thousands of subscribers to *ham radio's* former sister magazine, *Ham Radio Horizons*. These readers will be receiving *ham radio* each month on an issue-for-issue basis for the balance of their subscriptions (and, we hope, for many years after that). If you are one of our loyal supporters who took both magazines, then your *ham radio* subscription will be extended on the same issue-for-issue basis.

What are our plans for the combined magazine? Alf Wilson, W6NIF, our editor, will be explaining on page 6 in greater detail, but basically we intend to devote a major segment of each issue to keeping *ham radio* the pre-eminent technical publication in the Amateur field. At the same time, we plan to offer an interesting variety of new features derived from successful ideas that first appeared in *Ham Radio Horizons* — plus a few that are new to both publications. All of this material will be carefully chosen to make *ham radio* the most interesting and useful Amateur Radio magazine, both for the technically involved reader and the active but less technically oriented Amateur.

It all adds up to a big challenge for Alf and his staff, but he is well geared up for this project and I'm sure you're going to be seeing some of the best issues of *ham radio* that we've ever put out. Your comments will be of great value to us during this period of change. Don't hesitate to let us know your likes and dislikes, and also any other suggestions you may have.

Many folks have been wondering why we stopped HRH, at a time when it was getting better and better with each issue and was showing such great promise. It wasn't an easy decision to make, especially in view of the fact that our readership kept growing stronger every month after our redirection of the magazine early last year. Our weakness was in the advertising column. As the newest of the major Amateur monthlies, *Horizons* never really got its share of the Amateur Radio advertising action. A new advertising sales team, which took over here last summer and had great luck in making *ham radio* grow, could not — despite their extensive efforts — keep the outlook for *Horizons* from deteriorating.

The answer soon became obvious. With the combination of our two magazines, *ham radio* now has the largest circulation of any of the independent Amateur magazines, and our new economics of operation allow us to give the advertiser a value unmatched anywhere else.

Everyone should benefit, because a strong magazine means a better magazine. Suddenly everyone here in the *ham radio* organization can double his efforts toward our one magazine.

This issue represents just the beginning of our new direction. It will take a few issues to really get the new format in place. We have a lot of good ideas, and by the time spring rolls around you'll be seeing the most interesting and best balanced magazine that has ever been offered to the Radio Amateur.

Skip Tenney, **W1NLB**

2A Versatility

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Observation & Opinion

As the new editor of *ham radio*, I've seen a number of changes occur in this organization. Some of these changes reflect difficult choices and decisions, but all of them have been made with the continued welfare of *ham radio* and its readers in mind.

The structure of the Ham Radio Publishing Group has changed. One of the best Amateur magazines in the world, *Ham Radio Horizons*, has been discontinued. The decision to drop *Horizons* was made with great reluctance. However, business is business, and *Horizons* just didn't pull its weight in advertising revenue, although the magazine flourished in the Amateur community. It provided many newcomers with down-to-earth information on operating techniques, easy-to-build construction projects, and an opportunity to get the feel of what Amateur Radio is all about.

ham radio, on the other hand, provides advanced Amateurs with the latest state-of-the-art developments in technology. It will continue to do so. So we have a dilemma: How do we keep *Horizons* readers interested in our product without turning off the old guard *ham radio* reader? It's a real challenge — one that is rare in the technical-publications business.

Our decision has been to expand *ham radio*, still retaining the best in communications technology but also including material for readers who have enjoyed *Ham Radio Horizons*.

In this issue you'll find two features that were standbys in *Horizons*: Bill Orr's "Ham Radio Techniques" and "DX Forecaster." The latter is a two-page summary of what's happening for the month in the DX world, including a propagation chart based on scientific observations. Bill Orr's column, which explores Amateur Radio from top to bottom, will be welcomed by every active Amateur, old and new.

Also in this issue are some subtle changes in magazine graphics. Our graphics designer, Bill Scarborough, has instituted some interesting methods of portraying the editorial material so that it's easier to read and much more pleasing to the eye. We think you'll like it.

Here's a brief description of the other new features you'll find in future issues of the new *ham radio*. "Questions and Answers" was extremely popular in *Ham Radio Horizons*. We plan to continue this feature. We also plan to include our "Owner's Survey," which is compiled from reader responses to questions about popular pieces of ham gear. This is probably one of the most unbiased reviews ever published in an Amateur Radio magazine. It's the kind of report that will never be prepared by a laboratory or in a manufacturer's test facility — it tells it like it is: no holds barred.

The DXer is very much a part of Amateur Radio. This fellow is unique. He will do just about anything to work a new country. In *Ham Radio Horizons*, Bob Locher, W9KNI, wrote on the trials and tribulations of the DX operator in his series, "DXer's Diary." Bob's easy-going literary style was much appreciated by *Horizons* readers. In a future issue of *ham radio*, we'll publish another of Bob's articles on DXing. We'd like to hear from readers as to how they like this feature. If the response is positive, we'll continue "DXer's Diary"; if not, we'll drop it. Let us know what you think. It's your magazine. We'll do our best to publish articles that appeal to the most readers.

We're trying to provide the best balance of articles ever published in an Amateur Radio magazine. Let us know your desires and needs. We welcome all suggestions.

Alf Wilson, W6NIF
Editor

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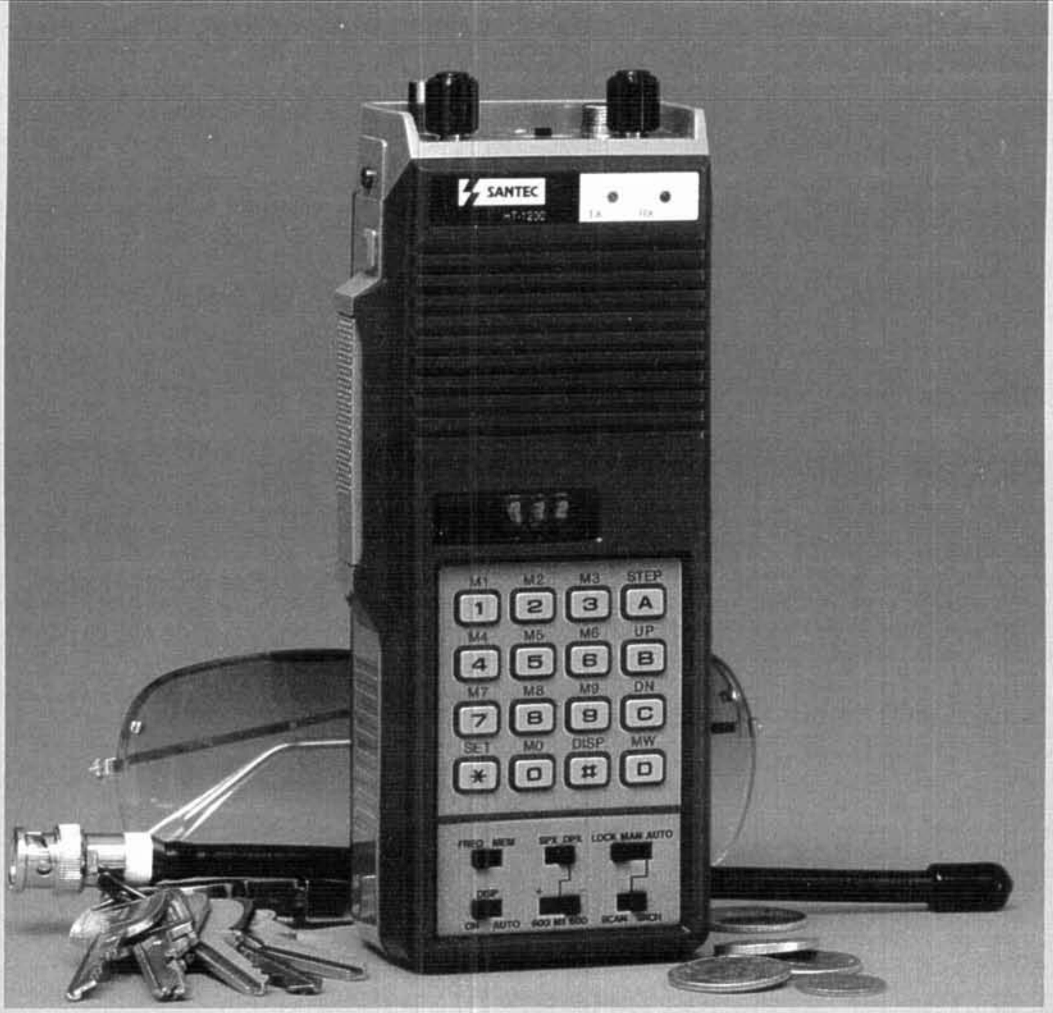

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presstop

HAM RADIO HORIZONS ceased publication with the December, 1980, issue. Though Ham Radio Horizons had succeeded in building a loyal and ever-increasing following of Amateur Radio readers, it had not achieved sufficient popularity with advertisers to put it into a very solid financial position. That, coupled with Horizons' pulling of some ad revenue from its parent publication, Ham Radio, and the current spottiness of the Amateur Radio marketplace, led to the decision to concentrate both editorial and marketing efforts of the staff on the older, more successful and better known magazine.

Ham Radio Horizons Subscribers will receive Ham Radio on an issue for issue basis; the 25 per cent or so who now receive both publications will have their Ham Radio subscriptions extended proportionally. Adding the unduplicated Horizons subscribers to Ham Radio will increase that magazine's subscriber list to more than 75,000. As a result, Ham Radio advertisers will, for the moment, be getting the best bargain in the Amateur Radio field.

Though The Shutdown after four years of publication is a disappointment to both Ham Radio Horizons' staff and its many loyal readers, it provides the necessary resources to ensure that Ham Radio will maintain its well entrenched position as Amateur Radio's top technical publication. In addition, Ham Radio's editorial coverage will be broadened somewhat by adding to it some of the more popular features for Ham Radio Horizons.

SIGNIFICANT IMPLICATIONS FOR AMATEUR RADIO can be inferred from Ronald Reagan's smashing victory in the presidential elections. The effects will be felt almost entirely through the FCC, which Washington observers feel will retreat sharply from the recent liberal philosophy under Chairman Ferris that led to directions blurring the distinctions between Amateur Radio and CB.

The Transition Won't Occur quickly, however; Dick Wiley was FCC Chairman for almost a year after Jimmy Carter won the presidency in 1976, while the new president tended first to higher priority changes in the Washington hierarchy. During the interim period, expect the Commission to move very slowly, avoiding any appearance of "making waves."

Though This May Mean Delays in implementing some needed rule making and even access to the new 10.1 MHz band, the more conservative Reagan administration is expected to return Amateur Radio's direction to more traditional paths than it has recently been on.

220 MHZ IS AGAIN BEING EYED by other services, according to current rumors from Washington. Users of the inland waterways system are reported to be looking at 220-225 MHz to relieve the congestion they're presently experiencing on the 160-MHz marine band, possibly at the suggestion of someone within the FCC!

Placing Inland Waterways Communications on 220-225 MHz came up earlier this year in FCC General Docket 80-1, an NPRM that was issued in response to the waterways users' request for more 150-170 MHz channels, suggesting 216-220 or 806-890 MHz for alternatives. The problem with that NPRM came in its summary, which cited 216-225 MHz instead of 216-220 MHz for expansion. After discussing the discrepancy with Amateur Radio representatives (and a storm of protests from 220 users), the FCC did issue a correction, but there was then and certainly is now some suspicion that inclusion of the 220 MHz Amateur band in the FCC's proposal was more a trial balloon than an error.

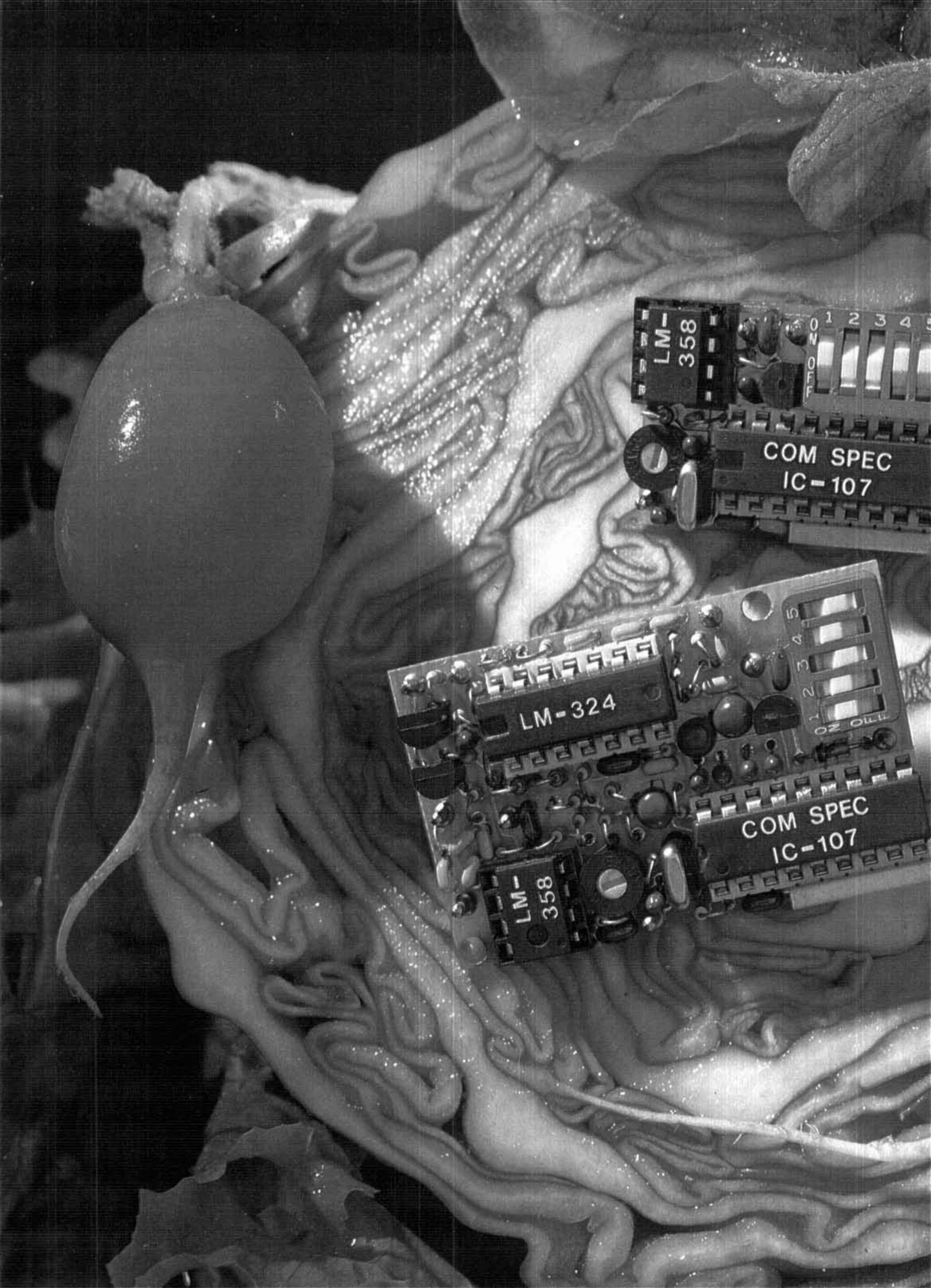
The ARRL Is Well Aware of this new threat, and is planning a determined campaign to quash it. Still another threat is brought out in the December, 1980, issue of Popular Electronics, in an article on cordless phones which says that manufacturers are looking to the FCC for new frequencies in either the 27 or 220-225 MHz band!

A PROBE INTO AMATEUR LICENSING IMPROPRIETIES dating back to 1976, when the August 15 issue of the Indianapolis Star ran a long article headlined "FBI PROBES HAM RADIO SCANDAL," is finally coming to a head. In a November 6 release, the Private Radio Bureau Chief ordered 13 Amateurs, 10 of them from Indianapolis or vicinity, to show cause why their station licenses should not be revoked, and suspended their operators licenses. Another, whose license is up for renewal, is under review.

All Of The 14 Are Alleged to have "fraudulently obtained or attempted to obtain new or upgraded Amateur licenses without examination" or "actively participated to fraudulently obtain or attempt to obtain new or upgraded Amateur licenses without examination" by the Commission in PR Dockets 80-668 through 80-682.

PLAIN LANGUAGE AMATEUR RULES GOT the Commissioners' nod recently along with some nice comments about both the Amateur Service and the staff's work on the rewrite job. The proposed new Part 97 has four parts: Part A, The Amateur Radio Service; Part B, The Radio Amateur Civil Emergency Service; Part C, The Amateur Satellite Service; and Part D, Technical Standards common to all three services.

Three Important Changes are in the proposed new rules. One: All previous log keeping requirements are deleted. Two: The Communications Act Section 303 requirement that an Amateur make his station available to the Commission for inspection will now be incorporated in the new Part 97. Three: Every Amateur would be required to keep a copy of Part 97 in his station.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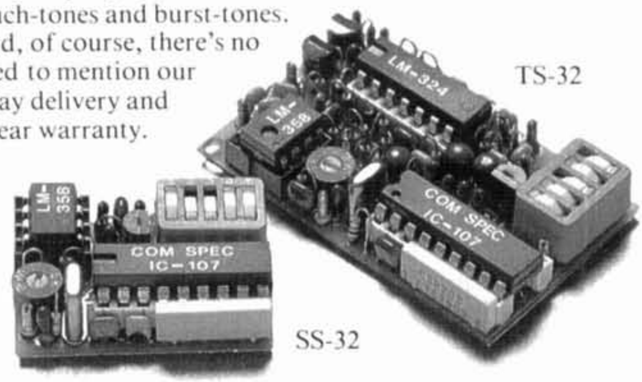

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71.9 XA	94.8 ZA	123.0 3Z	162.2 5B
74.4 WA	97.4 ZB	127.3 3A	167.9 6Z
77.0 XB	100.0 1Z	131.8 3B	173.8 6A
79.7 SP	103.5 1A	136.5 4Z	179.9 6B
82.5 YZ	107.2 1B	141.3 4A	186.2 7Z
85.4 YA	110.9 2Z	146.2 4B	192.8 7A
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1500	852	1477	1700	1950	2250	2500
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2805			1800	2100	2350	

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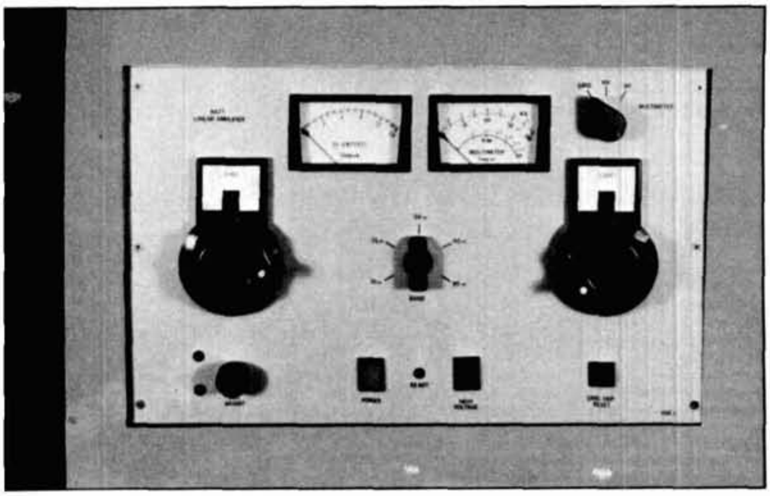


modular linear amplifier

for the high-frequency Amateur bands

Genesis of an 8877 amplifier — an example of Amateur craftsmanship

Many articles have been published on linear-amplifier construction. The reason is that an rf linear amplifier is one relatively simple, technical project the average Amateur can build that can rival the best commercial units on the market. The amplifier described here was inspired by the excellent series on amplifier design by Bill Orr, W6SAI,[1] combined with my desire to try some new ideas. The results of the project are illustrated and discussed below.

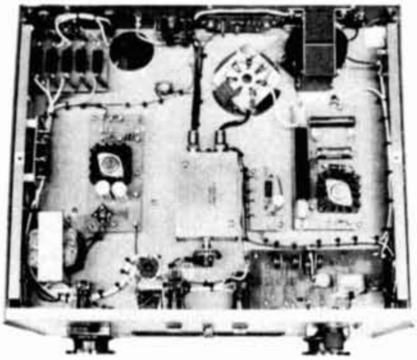
The amplifier was built for a friend, Steve, K8EJ, who wanted a "clean" amplifier built around the Eimac 8877 ceramic metal triode. Steve supplied all financial backing in return for a design that provided excellent performance and maintenance as well as reliability. These objectives have been met. I hope that the results of this project will encourage others.

preliminary building tips

After designing and building several different amplifiers, I've come to some basic conclusions and recommendations which I feel are worth consideration. First of all, build the amplifier *right* the first time! If you compromise anywhere, you'll not be happy with the product in the long run. Granted, it will cost more money and take more time and effort, but you'll be more gratified with the final results. For example, buy a commercial cabinet. Spend $35.00 each for nice meters with lighting kits. Buy a good, heavy filament transformer and quality pushbutton switches. Use Teflon wire and plenty of cable ties to bundle the cable harnesses.

Most important, *plan* your design. This is extremely important. (I spent time on this project drawing circuits, laying out the front panel, rear panel, and chassis.) Build the amplifier in *your mind* before you drill the first hole. Also, modularize the design and test as many of the modules as possible as they're constructed. Take the time to learn how the unit works. Study the technical details, and enjoy the project as a creation to be proud of, developed by your own hands.

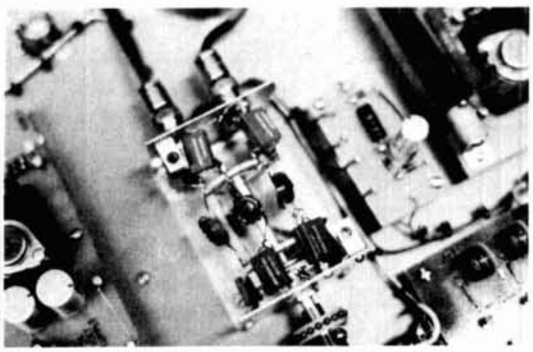
By Jerry Pittenger, K8RA, 2165 Sumac Loop South, Columbus, Ohio 43229



Amplifier bottom view. The modularization is evident here. Refer to fig. 3 for identification of each module.

circuit design

This amplifier **(fig. 1A)** is designed around the popular Eimac 8877 high-mu power triode. The 8877 was selected for its low drive requirements (that is, 50 watts drive for 2 kW PEP), its compactness, and the clean operating characteristics* exhibited by the tube.



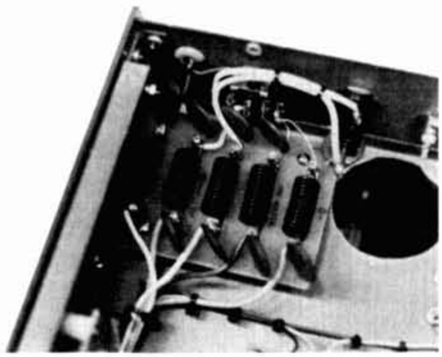
Tuned input network. Component values are:

band	CT (pF)	LT (μH)
80	860	3.23
40	445	1.67
20	225	0.84
15	150	0.56
10	112	0.42

band	core	wire size	turns
80	T50-2	22	20
40	T50-2	18	14
20	T50-6	18	10
15	T50-6	16	8
10	T50-6	16	6

*Refer to Eimac Bulletin 3950, which is the technical data sheet for the 8877 (3CX1500A7).

The amplifier uses a tuned input, which is ganged to the main band switch, and an effective ALC circuit **(fig. 1B)** to prevent overdriving the tube. The ALC level is adjustable from the front panel. Drive conditions can also be monitored by a pair of LEDs located on the front panel lower left corner. One LED is green and begins to flash with approxiately 50 mA of grid current. The other LED, red, begins to flash at approximately 150 mA grid current; thus, the LEDs provide a convenient way to monitor instantaneous grid current, which cannot be accomplished with metering circuits because of the slow response of meter movements.



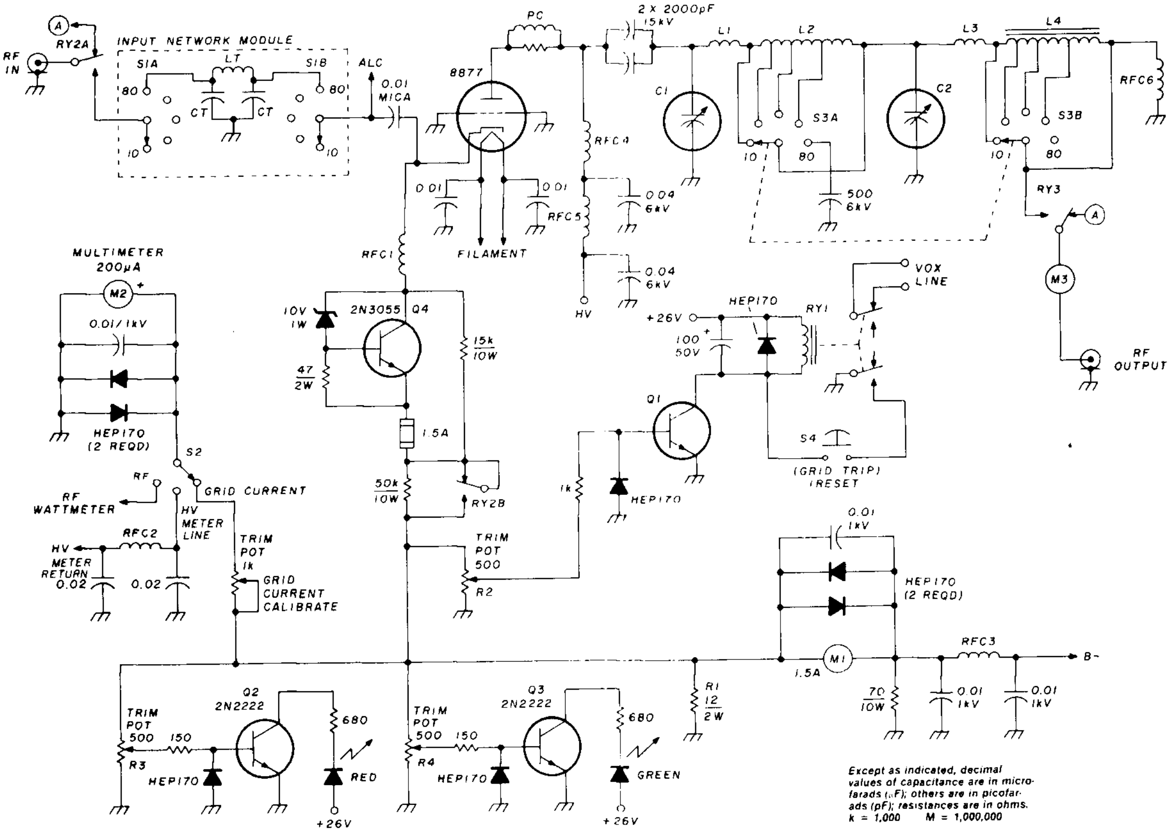
Lead filtering. All leads entering and leaving the amplifier rf deck are filtered using a pi-section network. Leads include 110 Vac coming to the amplifier, and 110 Vac returning to the power supply; a high-voltage metering line, a B- lead, and a chassis ground lead.

The amplifier also has a very effective grid-trip circuit, which snaps the amplifier out of the transmit line should the grid current exceed 200 mA. This feature protects the tube if the antenna is disconnected or if the tube should flash over for any reason. The grid trip circuit is reset for normal operation by pushing the GRID TRIP RESET button on the lower front panel.

The amplifier uses a vacuum relay on the output because of its quiet operating characteristics, compactness, and power-handling capability. Also, vacuum variable capacitors are used for both tune and loading controls in a pi-L output matching network.

Included also is a built-in rf wattmeter and a time-delay circuit, which keeps the blower on for about two minutes after the amplifier is turned off to ensure cooling.

The amplifier shown in the photos was built in a



C1	370-pF, 15-kV Jennings vacuum variable	PC	three 150-ohm, 2-watt carbon resistors in parallel shunted across one turn of 3/8-inch (9.5-mm) plate strap
C2	1000-pF, 3-kV Jennings vacuum variable	Q1	NPN — 2N3054
CT/LT	tuned input network*	RFC1	110 turns No. 22 (0.6-mm) enameled, on 1/2-inch (12.7-mm) diameter nylon rod (~28 μH)
cabinet	Intra Fab Inc., 10" × 17" × 14" (25.4 × 43.2 × 35.6 cm)	RFC2, RFC3	twelve turns No. 16 (1.3-mm) enameled, 3/8-inch (9.5-mm) diameter
chimney	Eimac SK2216	RFC4	200 turns No. 24 (0.5-mm) enameled, 7/8 inch (22-mm) diameter nylon rod (~180 μH)
Counter dials	Bauman TC3-S	RFC5	twelve turns No. 14 (1.6-mm) enameled, 3/8-inch (9.5-mm) diameter
L1	10-meter, 1/4-inch (6.5-mm) silver-plated copper tubing 1 3/4" diameter five turns tapped on third turn.	RFC6	1-mH, 0.3A rf choke
L2	15-80 meter, 1/4-inch (6.5-mm) silver-plated copper tubing 3 1/8" (10.6 cm) diameter nineteen turns tapped — one turn, 15 meters; five turns, 20 meters; twelve turns, 40 meters; nineteen turns, 80 meters.	RY1	4pdt Potter & Brumfield KHU17D11, 24 Vdc coil
L3	10-meter L-coil No. 14 (1.6 mm) wire — five turns, 3/4-inch (19-mm) diameter	RY2	KA11DG Potter & Brumfield dpdt, 24 Vdc coil
L4	toroid 2/T200-6 cores No. 14 (1.6 mm) Teflon sleeved wire, seventeen turns tapped — one turn, 15 meters; three turns, 20 meters; nine turns, 40 meters; seventeen turns, 80 meters.	RY3	PB2 Jennings vacuum relay
M1	Simpson "Wide VUE" 3 1/2-inch (8.9-cm) 1.5A model 1327 including 01253 mounting bezel and 01165 illumination kit	S1	two-pole, six-position, 60° indexing Centralab 2045
M2	Simpson "Wide VUE" 3 1/2-inch (8.9-cm) 200-μA model 1327 including 01253 mounting bezel and 01165 illumination kit	S2	one-pole, three-position, Centralab 2003
M3	Wattmeter (see fig. 5).	S3	radio switch model 86, two-pole, six-position, 60° indexing with sleeve bearings
		S4	Arrow Hart momentary pushbutton switch actuator AH83502, contact block AH83500-30, lense AH83500-84
		socket	see text or substitute Eimac SK2210

fig. 1. Schematic of the modularized linear amplifier rf deck (A).

*Only one network shown for clarity. See table on page 13 for CT/LT values.

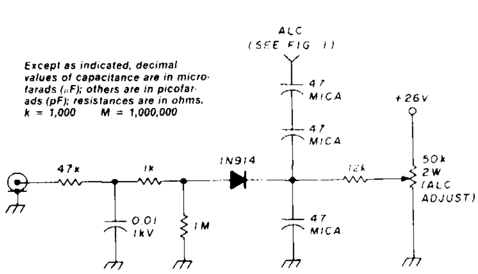


fig. 1. Automatic level control schematic (B).

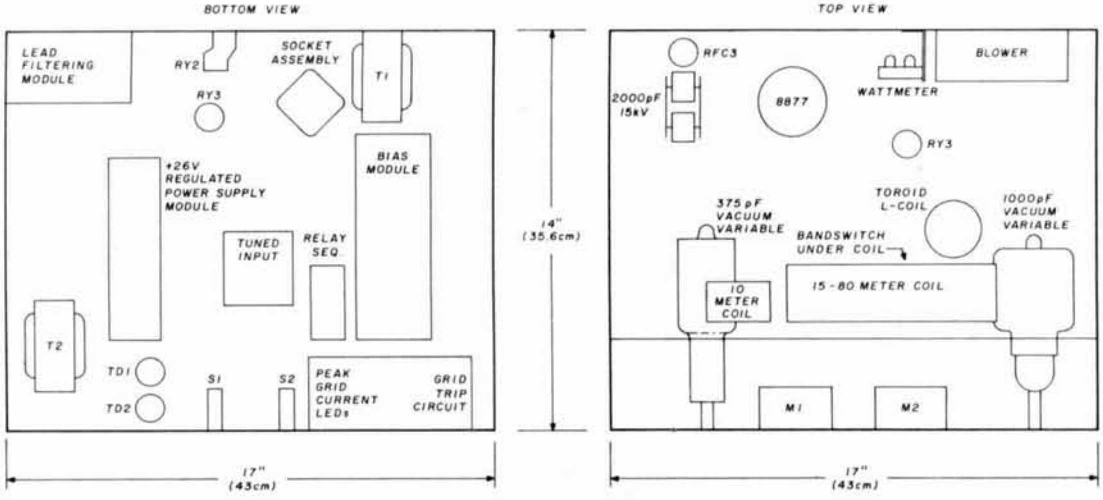
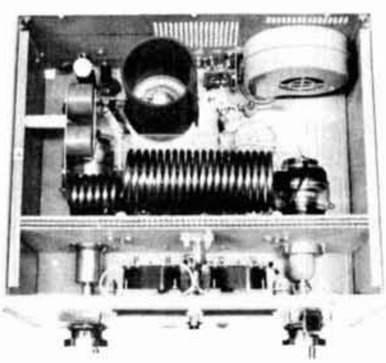


fig. 2. Modular component layout.

modularized fashion. **Fig. 2** shows the location of each module below and above the amplifier chassis. The major modules are discussed.

amplifier control circuit

The control circuit (**fig. 3**) provides the power control for operation of the amplifier. Ac power is received from the amplifier power supply in this design. Actuating the POWER ON button on the front panel turns on the blower, the tube filaments, and initiates a three-minute warm-up sequence for the 8877. Time delay is accomplished using an Amperite time delay tube, which provides a reliable timing circuit. The Amperite delay tube has a normally open set of contacts, which close after the designed period of time after 110 Vac has been applied to the delay-tube coil. The POWER ON button also activates the 26 Vdc regulated power supply to light the meters and POWER ON front panel push button. Note that the amplifier can't be keyed up until the required warm up time has elapsed (S2B). When three minutes have elapsed, the READY LED, located at the bottom center on the amplifier front panel, lights. The HIGH VOLTAGE front-panel pushbutton can then be pressed, which results in 110 Vac applied to the power supply to actuate the step start relays, which turns on the high voltage. Note that activating the HIGH VOLTAGE switch before the READY light comes on results in no action.



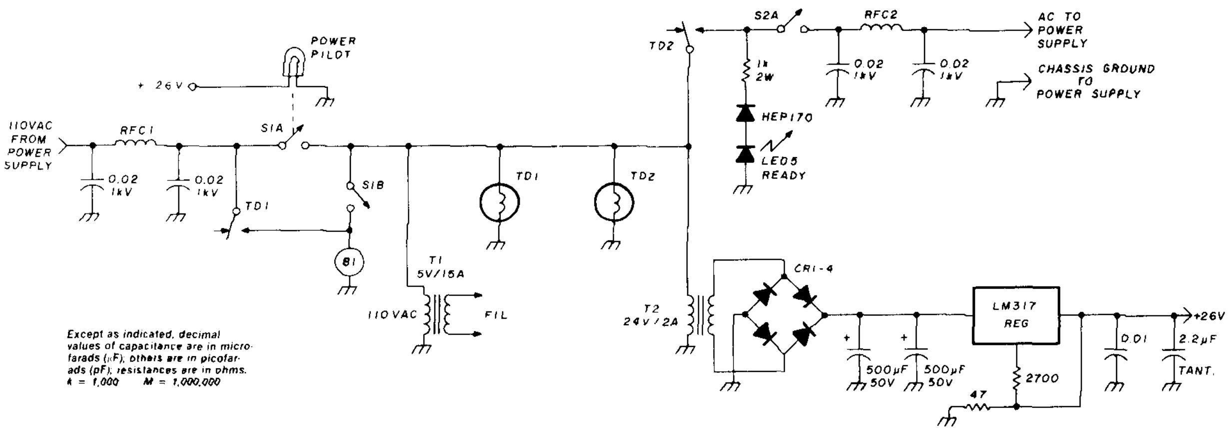
Amplifier top view. All metering circuits are isolated from the rf section using a subpanel. The subpanel is used to mount the bandswitch and the two vacuum variables.

The 26-Vdc power supply uses an LM317 voltage regulator chip, which is readily available. A regulated supply was used to avoid having the pilot lights and meter lights dim when the amplifier is keyed up.

input network

The amplifier input network is designed for a $Q=1$ and is enclosed in a self-contained shielded module. The design departs somewhat from conventional designs in that toroids are used instead of slug-tuned coils. Initially, there had been concern that the ferrite cores would saturate at the given drive powers, but this proved not to be a problem. I've used this same module in two other amplifiers with great success. The advantages are obvious. The toroids are compact and can be suspended right into the network bandswitch for easy mounting.

The rf enters and leaves the input network through BNC connectors. Since the nominal input impedance



B1	Dayton 1C180 blower	S1,S2	Arrow Hart pushbutton lighted switch actuator AH 83504 contact block AH-83500-30 lense AH 83500-72 two contact blocks stacked for two pole A,B sections
CR1-4	Motorola MDA970-2 or equivalent	T1	5V, 15A Stancor P6433
LED 5	Red light-emitting diode 5V, 15A	T2	24V, 2A Stancor P8662
RFC1, RFC2	twelve turns No. 14 (1.6-mm) enameled 3/8-inch (9.5-mm) diameter	TD1,TD2	Amperite time delay tube 115N0180 (180 seconds)

fig. 3. Amplifier control circuit.

of the 8877 is approximately 60 ohms, the module can be tested separately by connecting it in series with a 50-ohm dummy load. Initially, I used plenty of inductance on each core, then removed one turn at a time to obtain a flat SWR across each band with the fixed micas used. The input network is connected to the main bandswitch using a bead chain sprocket.*

grid-trip circuit

The grid-trip circuit automatically causes the amplifier to drop out of the rf line when the grid current reaches a specific level. It's a great feature and operates as follows.

Grid current is drawn through the 12-ohm, 2-watt resistor, R1, developing a voltage drop across the resistor, which turns on transistor switch Q1 (**fig. 1**). The amount of grid current drawn through R1 to turn on Q1 is determined by the voltage divider created by trim pot R2. When Q1 turns on, current is drawn through the grid-trip relay coil, RY1, which actuates the relay. Three sets of relay contact on RY1 are used. One set sends 26 Vdc to the pilot light on the GRID TRIP RESET normally closed momentary switch located on the front panel, which indicates that the grid trip has been actuated. A second set of contacts (normally closed) break the VOX line, resulting in amplifier standby mode. The third set of contacts provide a path to ground for RY1, so it remains closed. The GRID TRIP RESET switch deactivates RY1, which puts the amplifier back into the ready mode. Of course, you should check loading conditions to determine the problem that resulted in the excessive grid current.

peak LED grid current indicators

The peak grid current indicators operate on the same principle as the grid trip reset circuit discussed previously. The voltage generated across R1 turns on transistors Q2 and Q3, which operates the red and green LEDs respectively. Voltage dividers R3 and R4 provide a way to adjust the grid current levels at which the LEDs turn on. I recommend that the LEDs be turned on at approximately 150 mA (red) and 50 mA (green) respectively. In proper operation, the drive power should be controlled using the ALC adjustment to allow the green LED to just start flashing on voice peaks. Any flashing of the red LED indicates either excessive drive power or improper loading-control adjustment in the pi-L circuit.

lead filtering

All control cables between the amplifier rf deck and the power supply are filtered before entering or leaving the rf deck. This will minimize any rf leakage, which might cause RFI. The filters are mounted on a separate module and use a pi-network section. Make the leads from each pi section as short as possible to the rear-panel connector.

operating bias circuit

Operating bias is generated using a high-power 2N3055 NPN transistor (Q4) which, with the 1-watt zener, acts as a high-power zener (see **fig. 1**). This circuit includes readily available components and

*Available from Small Parts, Inc., 6901 N.E. Third Avenue, Miami, Florida 33128. Ask for their catalog. They have a wealth of hard-to-find parts available in small quantities.

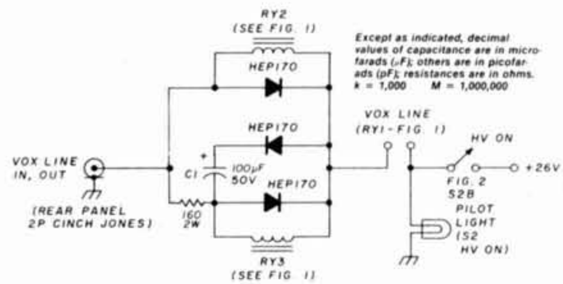
provides an easy way to adjust the bias voltage merely by changing the 1-watt zener between the collector and base of Q4. The bias should be adjusted to give approximately 180 mA idle current for the plate voltage used.

blower delay circuit

When the amplifier is turned off, the blower remains on for about two minutes. This is accomplished with an Amperite three-minute time delay relay (TD1). The Amperite time delay operates by heating a resistive coil, which acts as a filament, to heat a thin strip of metal, causing it to bend, thus closing two relay contacts. I found that the three-minute Amperite unit required about two minutes to cool after the power had been removed to the point that the relay contacts open. This delay keeps the blower on after the amplifier is turned off. Note that S1B turns on the blower immediately with the filaments.

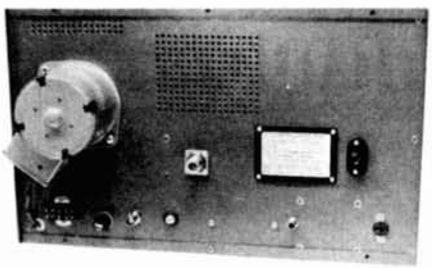
in/out relay sequencing

When keyed up, the amplifier closes both an input open-frame relay and an output vacuum relay. The relays must be sequenced to ensure that the output vacuum relay is closed before any drive power reaching the tube. Should the output relay close second, the amplifier would be placed in operation for a short period without a proper 50-ohm load. This condition could be harmful to the 8877 tube. **Fig. 4** illustrates the timing circuit that accomplishes proper relay sequencing. Capacitor C1 determines the length of delay on the input relay. Usually the delay can be detected by your ear.



RY2	KA11DG P&B dpdt, 24 Vdc coil
RY3	RB2 Jennings vacuum relay

fig. 4. Amplifier relay sequencing schematic.



Amplifier rear view. The holes in the rear panel are for air intake to the blower. The 10-pin Cinch Jones connector is for the control cable running to the high-voltage power supply. Rf input is through the BNC connector, and the rf output is through the type N connector. The high-voltage connector is a Millen 37001. The phono plug provides ALC to the exciter. The two-pin Cinch Jones connector is the VOX line to control keying up the amplifier in transmit mode.

pi-L matching network

The pi-L network was selected because of its improved harmonic attenuation characteristics over a regular pi-network. The inductance values depend on the tube plate impedance, which is based on the operating conditions of the tube and the selected network Q. The plate impedance is determined by:

$$I = \frac{V_p}{(1.6)(I_p)} \tag{1}$$

where I = plate impedance (ohms)
V_p = plate voltage (volts)
I_p = plate current (amperes)

Given the plate impedance, I, component values in the pi-L network are available for a Q of 12 from tables published in the literature.[2]

I recommend that the L coil be shielded from the pi coil for most efficient operation. Therefore, a toroid coil was used for the L section in this amplifier. A toroid has the inherent advantage of being self-shielding. The toroid L coil also has the advantages of compactness. The pi-coil is wound from 1/4-inch (6.4 mm) diameter soft copper tubing that has been silver plated.*

Taps on the coils always seem to be a troublesome detail for many builders. A way I've found to be successful in determining tank-circuit inductance is this:

1. Mount the coils into the amplifier but disconnect the tube end of the coil network and the plate and load vacuum-tuning capacitors.

2. Obtain a high-tolerance mica capacitor (selected for its low-inductance properties). Connect one lead to the tube-end of the network. With a very short piece of wire, connect the other lead to the coil turn at which the inductance is to be determined.

*Custom-made coils may be obtained at nominal charge. Send a SASE for details to the author.

3. Use a grid dip meter to determine network resonant frequency.

4. Determine the coil-tap inductance from:

$$L = \frac{10^6}{2\pi^2 f^2 C} \qquad \textbf{(2)}$$

where L = inductance (μH)
f = frequency (kilohertz)
C = capacitance (pF)

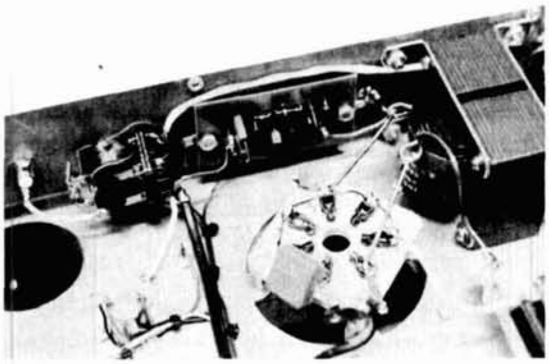
5. Set taps according to inductance values for the circuit Q selected ($Q = 12$ is recommended).

Remember that these data provide approximate inductance values, since the leads to the bandswitch will also affect inductance. I've found that slightly less inductance from the values determined by the procedure above should be used for 10 and 15 meters.

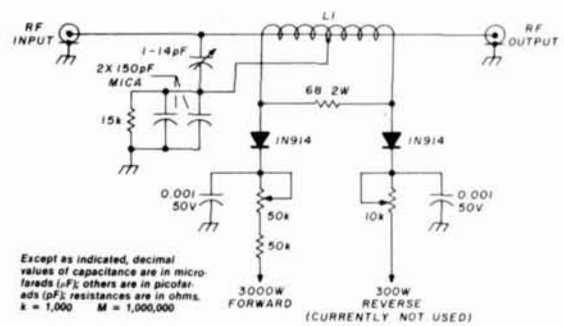
cooling

I have only one recommendation concerning amplifier cooling, and that is use plenty of it! Excessive heat will destroy a vacuum tube in short order. Refer to Bill Orr's comments[1] for a discussion of a representative air-cooling system for the 8877 tube.

Note the air duct in the photo that directs the air out of the top of the cabinet. The air flow is ducted to avoid the blower from short circuiting the hot air into the blower intake. The air duct is made from a 1/8-inch (3-mm) Neoprene sheet. The Neoprene sheet was first wrapped around the tube and held in place with a large cable tie. The seam was then stitched with a large needle and thread. I made a slit at the bottom of the neoprene tube to allow the plate strap



8877 socket assembly. The tube socket is a Johnson 248 submounted 1/2 inch (12.7 mm) below the chassis. Grid clips were made from a stiff copper stock and were silver plated. (An Eimac socket assembly, SK-2210, can be purchased directly from Eimac.) The Eimac Technical Data Sheet for the SK-2200 and SK-2210 air system gives detailed mounting information. If only the grid clips are needed, order Eimac part 149-842.



L1 twenty-two turns No. 22 (0.6-mm) enameled, on T-50-6 ferrite toroid, centertapped

fig. 5. Rf wattmeter schematic.

to exit. The blower air intake is through the back panel.

meters

The meters in this amplifier are 1.5-ampere for plate current and 200 microampere for the multimeter. A microammeter was selected for the multimeter because of sensitivity requirements of the built-in rf wattmeter. The plate-current meter was used as purchased. However, the 200 μA multimeter required work relabeling the scale. The 0-200 μA scale was used for grid current. All other lettering was made with dry transfers available from most electronic suppliers. Note that any labeling that comes on a meter scale can usually be removed with a pencil eraser and a little determination. The additional scale for rf power was added to the meter using a set of dry transfers contained in a kit.*

rf wattmeter

The rf wattmeter (**fig. 5**) is taken directly from the Drake L4B. Any of the popular wattmeter designs will work well. An rf wattmeter was selected over a relative output meter to eliminate the sensitivity control, which must be adjusted at different frequencies.

concluding remarks

Results from using this amplifier have been gratifying. The amplifier loafs at 1 kW dc input. It should give many years of trouble-free service. The project has resulted in a piece of equipment that is of better quality than most commercial amplifiers on the market and can be built at a competitive price.

reference

1. William I. Orr, W6SAI, "Design Considerations for Linear Amplifiers," *ham radio*, June, 1979, page 12.
2. Irvin M. Hoff, W6FFC, "Pi Network Design for High-Frequency Power Amplifiers," *ham radio*, June, 1978, page 52.

ham radio

*DATAK Corporation, 65 71st Street, Guttenberg, New Jersey 07092.

elliptic lowpass filters for transistor amplifiers

Advances in computer-aided filter design provide increased exactness and number of designs for Amateur use

The typical solid-state rf amplifier usually requires lowpass filtering to attenuate harmonics to an acceptable level. A selection of eleven Chebyshev lowpass filters was recently published for the 160-10-meter Amateur bands, but because of the relatively gradual rise in attenuation, the second harmonic was attenuated only 27-40 dB.[1] For those desiring greater attenuation at the second-harmonic frequency, the elliptic filter may be preferred, because its two resonant circuits can be tuned very closely to the second and third harmonics of the transmitter fundamental output frequency. This technique was first used in 1967 in the design and construction of a 10-meter harmonic TVI filter.[2]

background

In his March, 1974, *ham radio* article, G. Kent Shubert, WA0JYK, improved on the harmonic filtering application and presented six elliptic designs of seven elements, which required only standard-value capacitors.[3] Shubert designed one filter for each band using the well-known and frequently referenced elliptic tables of Saal and Zverev.[4,5] Because the normalized filter design data in these references was available only for the commonly published eleven values of reflection coefficient, Shubert was restricted in his design selections. To minimize the effect of this restriction, the exactness of the designs was compromised slightly to make all the capacitor values come out as standard values, thus simplifying filter construction.

Recently it has become possible to calculate the component values of elliptic filters for *any* desired reflection coefficient using a small computer such as the Radio Shack TRS-80.[6] Thus it's no longer necessary to restrict the elliptic designs to only the published normalized data. Because of this advancement in computer programming, it's now possible to make

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significant improvements on Shubert's designs. This article demonstrates how these advancements in computer-aided filter design can be used to increase both the exactness and number of elliptic lowpass filter designs for Amateur Radio applications.

filter designs

Table 1 lists thirty-one 50-ohm elliptic lowpass filter designs that have the unusual feature in which the three shunt capacitors (C1, C3 and C5) are all of the standard 5 per cent value. These designs were especially selected for transmitter output filtering in the six Amateur radio-frequency bands of 160 through 10 meters. Only those designs having reflection coefficients less than 13 per cent (VSWR = 1.3 and $A_p = 0.074$ *dB*) were selected to minimize the maximum VSWR and to make the designs less sensitive to component-value variation. **Fig. 1** shows the filter schematic diagram; **fig. 2** shows the attenuation response.

The first four columns of **table 1** list the filter design identification number and the frequencies (in MHz) at the A_p, *3-dB* and A_s attenuation levels. The significance of the frequency/attenuation parameters is illustrated in **fig. 2**. Column five lists the minimum value of the stopband attenuation, A_s, in dB. The remaining columns list the reflection coefficient, component values, and the *F2* and *F4* frequencies of maximum attenuation. The column headings of the capacitor and inductor listings are associated with the similarly labeled components in **fig. 1**.

In Shubert's designs, the values of C2 and C4 (capacitors that resonate the inductors to *F2* and *F4*) were given standard values. In **table 1**, however, the values of C2 and C4 are listed exactly as calculated and no attempt has been made to force these two values to be standard.

Standard values for these two capacitors are unnecessary, because these two circuits should be individually tuned to the exact design frequency either by using trimmer capacitors or by slightly varying the inductor turns. For convenience in construction, the standard value closest to the listed C2 and C4 capacitor values may be used if the inductor is adjusted to tune the circuit to the listed frequency. If this is done, the resulting attenuation will differ only slightly from the exact design value.

For example, in filter design 4, if the nonstandard C4 value (375 pF) is replaced by either standard values of 360 pF or 390 pF, and if L4 is adjusted to resonate at 4.04 MHz, the new attenuation response between 2-4 MHz will differ by only ±0.5 dB, respectively from the exact design response.

A similar procedure may be used for C2 and L2. This characteristic of the elliptic filter demonstrates the importance of *accurately* tuning the resonant circuits to the listed frequencies. In comparison, the exact values of C and L of the resonant circuit are less important. This tuning may be made with a calibrated grid-dip meter.

computer-calculated designs

The computer-calculated filter designs of **table 1** have several advantages over the manually calculated designs.[3] The main advantage in using a computer to perform the design calculations is that several designs can be easily obtained for each band, so that one can choose a design that may be optimum for a particular application; or perhaps some capacitor values will be more convenient to obtain than others. The minimum stopband attenuation, A_s, will usually be higher in the computer-designed filters than in those manually calculated. For example, five of Shubert's designs[3] have A_s values of 41 dB or less. In comparison, the computer-designed filters have A_s values between 47-63 dB. The computer-designed filters have a maximum passband ripple amplitude, A_p, of 0.074 dB, which corresponds to a maximum VSWR of 1.3. In comparison, all of Shubert's designs had A_p values of 0.1 dB or more, implying higher VSWRs.

To demonstrate the application of the tabulated data in constructing an actual filter, design 18 was assembled and its insertion loss was measured with a spectrum analyzer. **Fig. 3** shows the assembled filter with the inductor specifications. A plot of the insertion loss is shown in **fig. 4**.

L2 = 0.675 μH, T44-10, 13-turns No. 22
(adjust turns to resonate at 43.3 MHz)
L4 = 0.562 μH, T44-10, 12-turns No. 22
(adjust turns to resonate at 28.1 MHz)

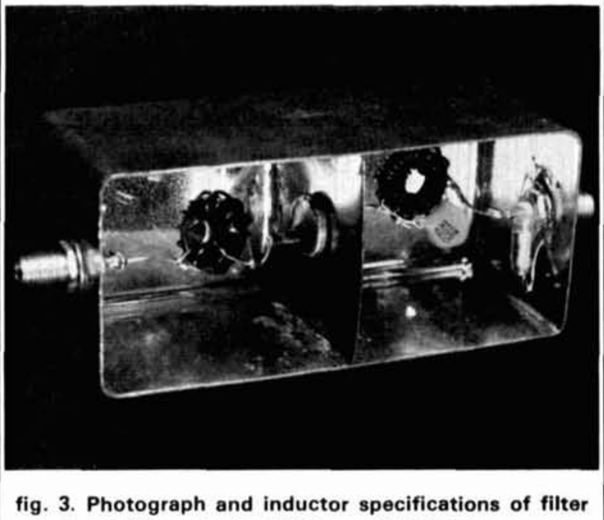

fig. 3. Photograph and inductor specifications of filter constructed from design 18 (F_{A_p} = 14.9 MHz).

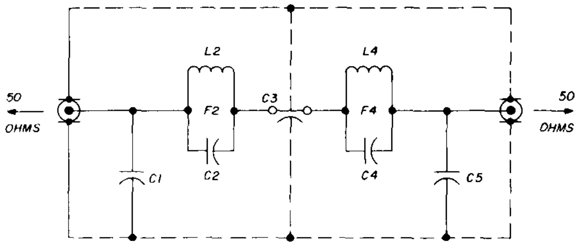

fig. 1. Filter schematic diagram.

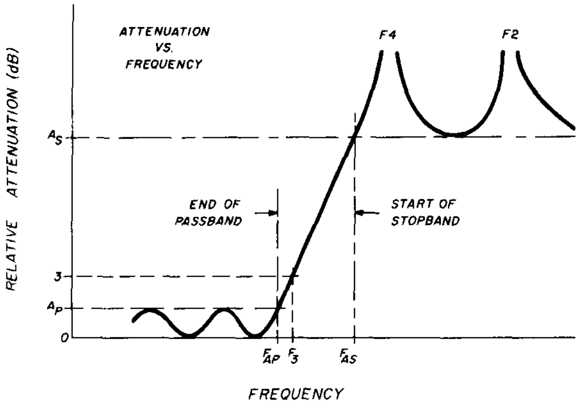

fig. 2. Typical attenuation response.

table 1. Elliptic lowpass filters selected for Amateur transmitter output filtering, 160 through 10 meters.

FLT NO.	F-AP (MHZ)	F-3DB (MHZ)	F-AS (MHZ)	AS (DB)	R.C. (%)	C1 (PF)	C3 (PF)	C5 (PF)	C2 (PF)	L2 (UH)	F2 (MHZ)	C4 (PF)	L4 (UH)	F4 (MHZ)
1	1.87	2.47	5.31	62.4	3.32	1100	2400	1000	63.0	5.15	8.84	173	4.74	5.56
2	1.89	2.19	3.68	53.3	9.50	1500	2700	1300	132	5.39	5.96	369	4.65	3.84
3	1.92	2.35	3.80	49.3	5.41	1200	2400	1000	130	5.10	6.17	372	4.32	3.97
4	2.05	2.40	3.87	50.5	8.22	1300	2400	1100	133	4.90	6.24	375	4.15	4.04
5	2.09	2.55	4.07	48.6	5.39	1100	2200	910	124	4.68	6.60	355	3.94	4.25
6	2.21	2.73	5.62	61.9	5.69	1100	2200	1000	63.8	4.59	9.30	175	4.19	5.88
7	3.90	4.41	6.63	47.6	10.82	750	1300	620	88.4	2.57	10.6	252	2.10	6.91
8	3.91	4.67	7.29	47.8	6.38	620	1200	510	72.6	2.53	11.8	208	2.10	7.61
9	4.15	5.01	9.89	60.5	6.80	620	1200	560	38.5	2.47	16.3	106	2.23	10.4
10	4.16	4.74	7.14	47.3	9.95	680	1200	560	81.3	2.40	11.4	233	1.97	7.44
11	7.07	8.34	13.2	48.9	7.26	360	680	300	39.7	1.41	21.3	113	1.18	13.8
12	7.30	8.34	15.9	60.9	11.71	430	750	390	26.1	1.43	26.1	71.3	1.28	16.6
13	7.55	8.59	13.5	50.2	10.77	390	680	330	40.4	1.34	21.7	114	1.12	14.1
14	7.58	9.56	19.2	60.1	4.66	300	620	270	19.1	1.31	31.8	52.6	1.19	20.1
15	8.26	9.33	14.2	48.4	11.22	360	620	300	40.7	1.22	22.6	116	1.00	14.8
16	8.28	9.84	19.6	61.7	8.03	330	620	300	19.3	1.25	32.4	52.8	1.14	20.6
17	14.7	17.0	32.0	59.5	9.91	200	360	180	13.0	.707	52.4	35.7	.633	33.5
18	14.9	17.3	27.0	48.8	8.40	180	330	150	20.0	.675	43.3	57.0	.562	28.1
19	15.6	18.2	33.2	57.3	8.91	180	330	160	13.0	.661	54.2	36.0	.585	34.7
20	17.5	19.5	29.2	48.2	12.98	180	300	150	20.6	.572	46.4	58.5	.468	30.4
21	22.1	27.3	56.2	61.9	5.69	110	220	100	6.4	.459	93.0	17.5	.419	58.8
22	22.3	25.9	40.4	48.8	8.40	120	220	100	13.3	.450	65.0	38.0	.375	42.2
23	22.9	27.9	43.9	47.7	5.37	100	200	82	11.8	.426	71.0	33.9	.356	45.8
24	24.5	28.4	43.6	48.0	8.69	110	200	91	12.7	.408	69.9	36.4	.337	45.4
25	26.2	29.2	43.8	48.2	12.98	120	200	100	13.7	.381	69.6	39.0	.312	45.6
26	28.7	37.1	77.9	61.7	3.84	75	160	68	4.4	.342	129	12.1	.313	81.6
27	29.3	35.2	55.9	48.6	6.14	82	160	68	9.2	.337	90.4	26.3	.283	58.3
28	30.7	37.3	59.7	48.5	5.39	75	150	62	8.5	.319	96.7	24.3	.268	62.3
29	31.3	35.6	53.6	47.4	10.12	91	160	75	10.8	.320	85.5	31.0	.262	55.9
30	32.4	40.0	81.3	61.4	5.68	75	150	68	4.5	.313	135	12.2	.285	85.1
31	32.6	37.9	58.5	48.2	8.46	82	150	68	9.4	.307	93.9	26.7	.254	61.0

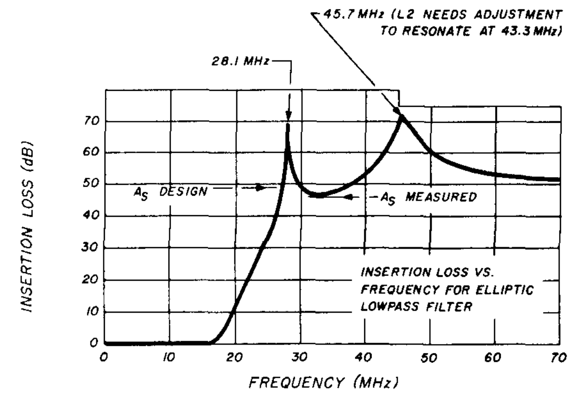

fig. 4. Measured insertion loss of elliptic filter constructed from design 18.

The plotted response shows that the circuit of L4C4 was accurately tuned to the design frequency of 28.1 MHz, but the L2C2 circuit resonance was slightly in error. The effect of this mistuning was not significant (only a 2-dB lower-than-normal level of A_s at 33 MHz), so no attempt was made to adjust the circuit to the correct frequency. This design was intended for a 20-meter filter application, and the attenuation at the second and third harmonics is greater than 50 and 60 dB respectively. The Micrometals T44-10 toroidal cores were on hand, and to expedite construction they were used; if the filter is to be used in filtering transmitter outputs greater than a few watts, however, the larger T80 cores used by Shubert are recommended.

construction notes

The filter was assembled in a 1.578 by 3.5 by 1.547 inch (40 by 89 by 39 mm) Hudson tinned steel box [12] (HU-6570-1.547ST). The center partition and the Button®-mica capacitor ($C = 330\ pF$) are recommended if maximum stopband attenuation is required above 50 MHz. If the filter cutoff frequency is less than about 10 MHz, the center partition probably can be omitted. The capacitor voltage ratings should be 500 V or greater, and the capacitors can be either polystyrene or mica. See references 7, 8, and 9 for information regarding the winding and application of toroidal inductors. References 10 and 11 give two distributors of the Micrometal cores.

Those familiar with modern filter design can confirm the correctness of the designs in **table 1** by using one of the published normalized designs of Saal or Zverev to calculate independently the component values of filter design 16. This is possible only because design 16 has a reflection coefficient of 8.03 per cent, which is almost identical with the 8 per cent standard published value. Also, the A_s value for design 16 is 61.7 dB, which is practically identical to the published value of 61.6 dB. Therefore, if the published normalized data for an elliptic design of C0508, $\theta = 25$ (see page 78 of Saal's book) is used, an independent calculation using the $F - A_p$ cutoff frequency of design 16 (8.28 MHz) should give the same component values listed in **table 1**.

The published normalized values of Saal for C1, C3, and C5 for $A_s = 61.6\ dB$ and for an 8 per cent reflection coefficient are: 0.8574 F, 1.612 F, and 0.7790 F. Capacitor scaling factor, $C_s = 1/(Rw) = 1/(50 \cdot 2\pi \cdot 8.28 \cdot 10^6) = 384.4 \cdot 10^{-12}$. Calculating the filter component values for C1, 3, and 5: $C1 = C_s (0.8574) = 330\ pF$, $C3 = C_s(1.612) = 620\ pF$, and $C5 = C_s(0.7790) = 300\ pF$.

These independently calculated values are identical to the computer-calculated values, and the validity of the computer program used to generate **table 1** is therefore confirmed. In a similar way, the other component values of design 16 may be confirmed.

I gratefully acknowledge the assistance of the following: Philip Geffe of Lynch Communications Systems for providing the computer program needed to calculate the elliptic normalized values, Roosevelt Townsend of Honeywell for his assistance in modifying the computer program to generate **table 1**, Charles Miller of Honeywell for his assistance in plotting the filter insertion loss, and Joseph Gutowski of EWC and Rex Cox of Honeywell for their review of the manuscript.

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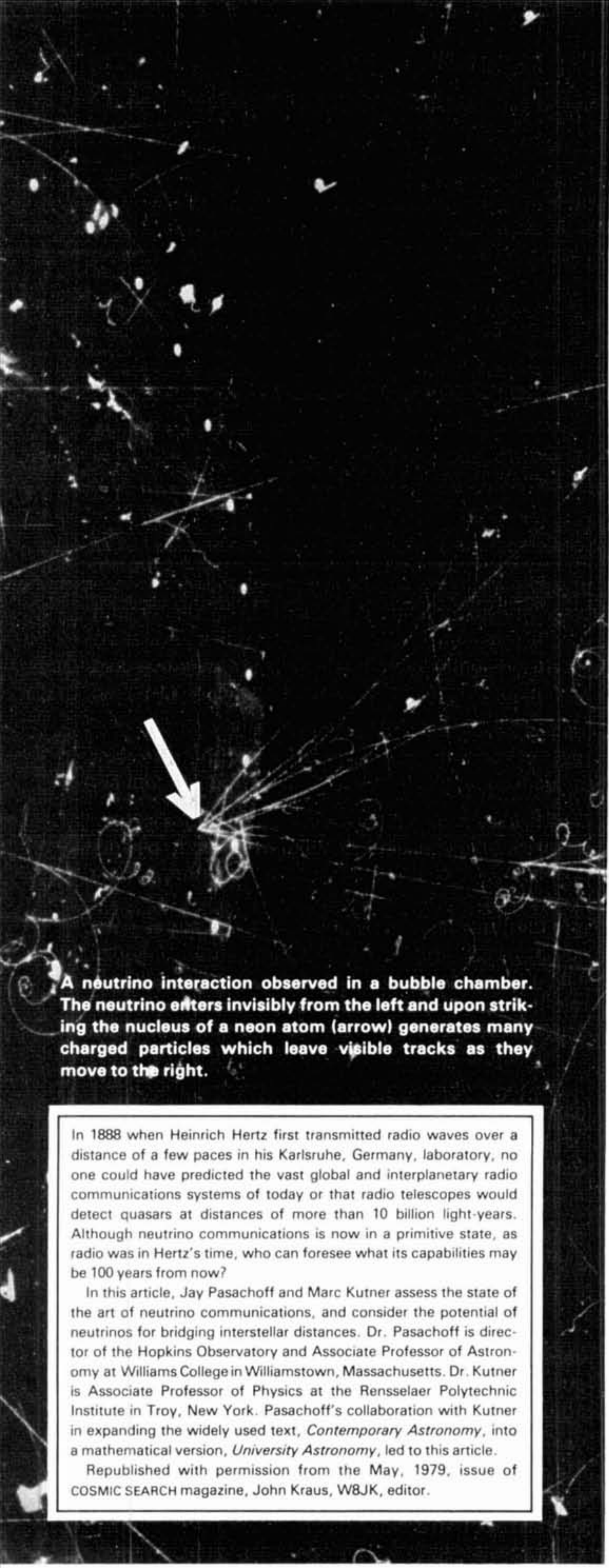

A neutrino interaction observed in a bubble chamber. The neutrino enters invisibly from the left and upon striking the nucleus of a neon atom (arrow) generates many charged particles which leave visible tracks as they move to the right.

In 1888 when Heinrich Hertz first transmitted radio waves over a distance of a few paces in his Karlsruhe, Germany, laboratory, no one could have predicted the vast global and interplanetary radio communications systems of today or that radio telescopes would detect quasars at distances of more than 10 billion light-years. Although neutrino communications is now in a primitive state, as radio was in Hertz's time, who can foresee what its capabilities may be 100 years from now?

In this article, Jay Pasachoff and Marc Kutner assess the state of the art of neutrino communications, and consider the potential of neutrinos for bridging interstellar distances. Dr. Pasachoff is director of the Hopkins Observatory and Associate Professor of Astronomy at Williams College in Williamstown, Massachusetts. Dr. Kutner is Associate Professor of Physics at the Rensselaer Polytechnic Institute in Troy, New York. Pasachoff's collaboration with Kutner in expanding the widely used text, *Contemporary Astronomy*, into a mathematical version, *University Astronomy*, led to this article.

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neutrino communications

A discussion of the use of neutrinos for interstellar communications and the search for extraterrestrial intelligence

Though no interstellar communication is going to be quick, it is obviously better to send messages that travel at the highest possible velocity, all other things being equal. This usually leads to the conclusion that we should communicate with radio or other electromagnetic waves, but one nuclear particle — the neutrino — travels at the speed of light, and the possibilities of using neutrinos as the vehicle for interstellar communication should be carefully explored.

It turns out that neutrinos have many advantages for this purpose, along with their major difficulties, and we think that investigation of neutrino communications should not be ignored. We note that we are here exploring possibilities whose realization would be in the distant future; nonetheless, since most other communicating civilizations would be far more advanced than our own, it is an interesting exercise

By Jay M. Pasachoff and Marc L. Kutner

to examine alternative means by which interstellar communication may have evolved.

Though neutrinos are very difficult to detect, the scientific benefits of studying the neutrinos that emanate from stellar interiors or supernovae have made it worthwhile. Continued improvements in the sensitivity of neutrino detection devices can be expected.

advantages of neutrino communications

Much discussion of the Search for Extra-Terrestrial Intelligence (SETI) has been taken up with finding a suitable frequency for radio communications. Interesting arguments have been advanced for a wavelength of 21 centimeters, the water hole, and other wavelengths. It is hard to reason satisfactorily on this subject; only the detection of a signal will tell us whether or not we are right. Neutrino detection schemes, on the other hand, are broadband; that is, the apparatus is sensitive to neutrinos over a wide energy range. The fact that neutrinos pass through the earth would also be an advantage, because detectors would be omnidirectional. Thus, the whole sky can be covered by a single detector. It is perhaps reasonable to search for messages from extraterrestrial civilizations by looking for the neutrinos they are transmitting, and then switch to electromagnetic means for further conversations.

Though the means of detecting neutrinos are now relatively insensitive, we must remember that we are in the dawn of the age of neutrino detection. It is only 23 years since the first antineutrinos were detected; on the other hand it is almost a hundred years since the detection of radio waves, and our sensitivity to them has increased manyfold. We can expect that in the next decades we will be better able to detect neutrinos from space, both those arising naturally from astronomical phenomena and those carrying messages from extraterrestrial civilizations. It may well be that neutrinos are considered by extraterrestrial civilizations to be a more advanced method of communications than radio waves.

Neutrino beams are currently generated on earth by proton beams impinging on targets in large accelerators. Proton beams are highly directional, and neutrino beams are also highly directional. The present neutrino beams generated at the Fermi National Accelerator Laboratory (Fermilab) in Batavia, Illinois, has a beamwidth as narrow as one-tenth of a degree. The beams are so narrow because of the way they are formed. As the incident proton beams become stronger in energy, the resulting neutrino beams become even more directional.

The doubling in energy expected within two years at Fermilab, to be matched by the accelerator at the European Center for Nuclear Research (CERN), near Geneva, Switzerland, will lead to a neutrino beam about one-twentieth of a degree in width. This is twice as sharp as the beam of the Arecibo radio telescope when used at 21 centimeters. At the distance of the star *Tau Ceti* (12 light years), the neutrino beam will be broad enough* to irradiate all the planets of a system like our solar system. Since we have no idea where any planets might be, we would have to cover roughly this scale to be sure we are not missing planets, not to mention the difficulties in the high precision tracking that would be necessary.

The tunnel that composes the "Main Ring" at Fermilab. The proton beam travels in an evacuated tube (inside square housing) around the 6 kilometer circumference ring 50,000 times per second being constrained to stay in the pipe by over one thousand magnets. On each revolution the beam is given a 3 million electron-volt "kick" from a radio transmitter boosting the proton beam energy to 200 billion electron volts or higher. Pipes carry cooling water to the electromagnet coils located at intervals along the square housing.

neutrinos

In the early 1930s, Wolfgang Pauli suggested that a new nuclear particle might exist to explain the

*Beamwidth 600 astronomical units (AU); the earth's orbit is two AU in diameter and Pluto's is about 80 AU.

apparent lack of conservation of energy and spin in beta decay, a radioactive process that involves the emission of an electron (a beta particle). Soon Enrico Fermi worked out a theory for the new particle, which he called *neutrino*, Italian for "little neutral one (since the particle has no charge and to distinguish it from the neutron, which had recently been discovered). The neutrino is a particle with no charge but with a certain energy and a certain spin; it travels at the speed of light, and to do so, it must have no rest mass.

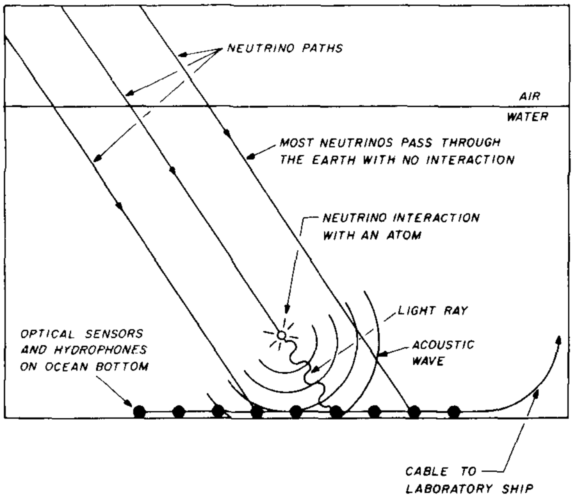

fig. 1. In the Deep Underwater Muon and Neutrino Detection system the light flash and accoustic "ping" resulting from a neutrino interaction with the nucleus of an atom would be picked up by optical sensors and hydrophones (underwater microphones) deployed on the bottom of the ocean.

Neutrinos are the most elusive atomic particles of which we know. They have a very low probability of interacting with any other matter. On the average, in fact, neutrinos can penetate four light years of lead before being stopped. Rather than try to detect individual neutrinos, we normally resort to capturing a tiny fraction of a huge number of neutrinos. In this way, Frederick Reines and Clyde L. Cowan, Jr., detected *anti*neutrinos in 1956 in the flood of particles coming from the Savannah River reactor. The existence of antineutrinos shows that neutrinos too must exist.

The disadvantage that neutrinos interact so weakly with matter is also an advantage, because this property allows neutrinos to escape from the center of a star. The nuclear fusion cycles that fuel our sun lead to the emission of neutrinos of relatively low energy. These neutrinos come straight out of the solar core, reaching the radius of the earth's orbit in 8 minutes, thus providing us with our only direct link to the solar core.

Raymond Davis, Jr., of the Brookhaven National Laboratory, has set up a 400,000-liter tank of perchloroethylene (a type of cleaning fluid, C_2Cl_4) to capture solar neutrinos. His tank is located deep underground in a gold mine at Lead, South Dakota, where it is shielded by earth and rock from other particles that would cause reactions in his material. He is now detecting 1.8 ($\pm$0.4) Solar Neutrino Units (SNUs)* of neutrinos. This amounts to about one interaction every six days.

Davis's results are lower by a factor of almost three than the best current predictions of the solar neutrino outflow, but within about twice the error (standard deviation) in the theoretical predictions, which give 4.7 ($\pm$1.4) SNU. Gallium is more sensitive to lower-energy neutrinos than chlorine, and Davis is planning an experiment using gallium; a Soviet group is also planning a gallium experiment.

Still higher-energy neutrinos should be emitted in supernova explosions. Such neutrinos might be detected by an apparatus being considered in Project DUMAND (Deep Underwater Muon and Neutrino Detection). One plan involves setting up acoustic or optical detectors over an area of several square kilometers on the floor of the Pacific Ocean near Hawaii to detect the Cerenkov radiation emitted when neutrinos interact with nucleons in the water. (Cerenkov radiation is emitted when a particle enters a medium traveling faster than the speed of light in that medium and is somewhat analogous to a sonic boom.)

A Very Big Accelerator (VBA) is under discussion as a possible international project on a large scale. This accelerator would generate a proton beam of 10 times higher energy than the new Fermilab and CERN beams. If directed at our solar system from *Tau Ceti*, this beam would encompass the planets out to Neptune.

neutrino generator and detection

Presently the proton beam generated at Fermilab in the 2 kilometer-diameter ring has energies of 400 billion electron volts. Neutrinos are produced in pulses of 10 billion per pulse with pulses occurring every eight seconds. The proton beam impinges on a target, resulting in muons, particles of mass intermediate between electrons and protons. The muons travel in a forward direction and their decay products include neutrinos. This forward beam travels through a "filter" of 1 kilometer of earth, which excludes all

*One SNU corresponds to 10^{-36} captures per chlorine atom per day, or one capture per day among a trillion, trillion, trillion atoms.

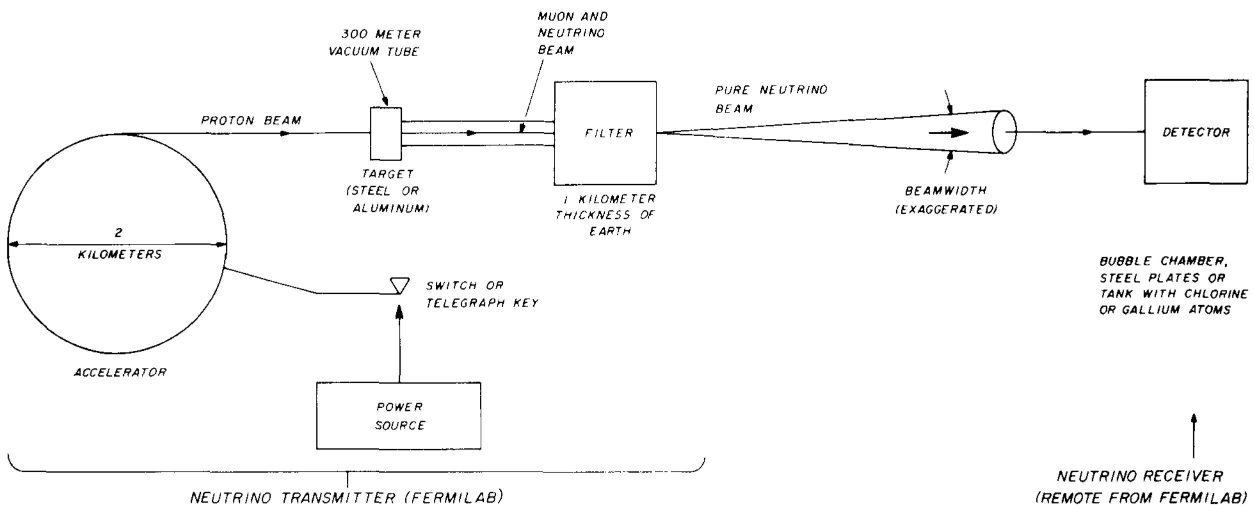

fig. 2. A neutrino "communications system." A guidance system to steer the beam is not shown but probably would employ magnetic deflection of the proton beam.

the particles except neutrinos. To detect the neutrinos, a bubble chamber or an apparatus containing approximately 1000 tons of steel is used. Even that much steel produces only about 1 interaction per pulse. A person can (and often does) stand in the neutrino beam without any effect on either the person or the beam.

As the proton beam is brought to higher energies, the number of protons per pulse goes up proportionally to the energy. Thus the particle flow (or flux) at a given distance from the ring goes up with two factors of energy from the increase in the number of protons and, thus, neutrinos per pulse. This means that the particle flux goes up with the cube of the energy. For a given neutrino beam, the number of interactions in the target is also roughly proportional to the energy. Thus for a one-trillion-electron-volt proton beam, we have calculated that about 50 trillion, trillion kilograms (5×10^{25} kilograms or ten times the mass of the earth) of material would have to be located at the distance of *Tau Ceti* to obtain one interaction per day with a neutrino from earth. This is about the rate of Davis's solar neutrino experiment. We are currently comparing prospective fluxes with the natural neutrino background.

With the technology we expect to have within a few years, we would need ten times the mass of the earth as detector material. However, we must remember that the neutrino pulses are *on* only about 0.00025 per cent of the time. If this percentage can be raised to even one per cent, and the proton beam energy can be brought up by another factor of 10, then the detector mass required is less than one millionth the mass of the earth to detect one neutrino event per day. This is equivalent to a cubic volume of 100 kilometers on a side, not too different from the scale of detectors being proposed for Project DUMAND.

We must also allow for further technological development. Already a detection scheme has been suggested that detects muons simultaneously with the Cerenkov radiation. Detecting muons can be achieved by a smaller apparatus. Even though a large detector mass may be required for interaction, the active part (or electronic equipment) involved could be much smaller.

Some investigation of neutrino communications has been underway for sending a neutrino signal through the earth to nuclear submarines on the other side of the earth, but this will require the reduction of the detection apparatus from a cube 100 kilometers on a side to the size of a submarine.

present data

Though solar neutrino schemes involve much lower energies than Project DUMAND methods, solar neutrinos represent our longest string of data on interstellar neutrinos. We have examined the Davis records for any sign of obvious messages, and have found none. It seems reasonable that any civilization that chooses to generate strong neutrino beams and point them at a given star would also calculate the appropriate number of neutrinos that inhabitants associated with the star would need to detect the incoming signal. Davis's data involve runs of a month or a few months, with interactions occurring once per day. There is no reason to believe that a neutrino signal need be any shorter in duration than months, nor indeed to be even that short. It seems worthwhile to continue collecting data on a regular basis to have long enough data runs to properly search for an extraterrestrial signal.

conclusions

Current plans for neutrino generation, coupled with prospective abilities for neutrino detection, are beginning to make feasible neutrino messages we might send to be detected at the distances of nearby stars. Though neutrino beams are expensive to generate, this disadvantage may be outweighed by such advantages as very narrow focusing of the transmitted neutrino beam, and the ability to receive simultaneously from the whole sky over a broad range of neutrino energies.

On earth, neutrino communications have lagged radio communications by about 100 years, during which time our ability to send and receive radio signals has increased dramatically. We can hope for similar improvements in neutrino communications in the next decades.

Technologically, the transmission of strong modulated neutrino signals would be a sure sign of an advanced civilization, and we should be looking for such a signal. Since our transmission of a neutrino signal would involve finding a way of making the neutrino beam track the stars to a high degree of accuracy, the transmission of neutrino signals from earth seems farther in the future.

It may be that neutrino communications is best suited for finding another civilization, after which dialogue could be carried out by electromagnetic means. In any case, if our reasoning of the advantages of neutrino communications is correct, then much of the information now crisscrossing the galaxy could be in this mode.

glossary

Muon:
A particle with a mass between that of an electron and a proton. It has a negative charge and a mean lifetime of 2.2×10^{-6} second.

Neutrino:
Either of two massless, electrically neutral, stable particles. One is named "electron's neutrino"; the other "muon's neutrino."

ham radio

More Details? CHECK—OFF Page 110

pulse-position control of the CDE Tailtwister rotor

Many Amateurs worry or become frustrated when they put up a large antenna array and then notice how unevenly the system moves when turned by one of the commercially available rotors. Large moments of inertia build up in long-boom hf Yagis or large multi-antenna vhf arrays used in EME communications. These antennas must be rotated at a slower rate than that which most commercial rotors provide. To relieve this frustration, a method is described for incremental pulse-position modulation of a Cornell-Dubilier Tailtwister rotor.

background

I became aware of the large moments of inertia that build up in an antenna system when I began the design of my 160-element collinear array for 144 MHz EME (or moonbounce, as others call it). This array has a 33-foot (10-meter) boom of 3-inch (7.6-cm) OD aluminum tubing that supports four 24-foot (7.3-meter) secondary booms of 1-1/4 inch (3.2 cm) aluminum tubing. Eight Cushcraft DX-120 collinear antennas are mounted on the secondary booms. The antenna array, excluding the elevation mount, weighs about 125 pounds (56.7 kg).

I could picture in my mind (as well as on paper) the jerking and twisting that would occur if this array were rotated by a conventional rotor. I'd planned to use a CDE Tailtwister.

Other complications arise when one wants only to move the array a few degrees in azimuth. A 1-degree change for an array with a 33-foot (10-meter) boom is only about 3.6 inches (9 cm) of arc movement at the end of the boom. The rotor doesn't need to be activated very long to give this change. Also, every time you push a control switch, the rotor tries to turn the array at full speed: quite an acceleration for a long-boom, large-mass array. We must also allow for array coastdown before releasing the brake switch, which complicates matters further.

theory

Could I add speed control to the motor used in the Tailtwister rotor? This question turned out to be easier asked than answered, as the motor is an induction type. It starts to lose torque quickly if its applied voltage is reduced. This problem rules out the use of Variacs and typical triac motor speed controls. After much meditation, I decided to take a different approach. Why not try to pulse-position modulate the rotor motor? I reasoned that if I could pulse the motor with full voltage, I could make it turn only a few revolutions and have practically full torque available. The rotor gearing would reduce these revolutions to a small incremental step in actual rotor position. Two major benefits arise:

1. The small incremental steps can be only parts of a degree, so that positioning the antenna to a particular azimuth setting would be no problem at all.

2. The slow motion of the long boom never allows any appreciable velocity buildup, so no moment-of-inertia problem occurs.

Now all I had to do was develop some hardware to do what all these words are saying, which turned out to be the least difficult part of this project. The pulse-control unit, as I call it, is surprisingly simple. A block diagram of the result is shown in **fig. 1**.

Basically two items were needed. The first was an electronic switch that could be opened and closed for small periods of time to apply power to the rotor motor. The motor would have full voltage applied to it during each pulse. Second, a pulse generator was needed to send constant-width pulses at a variable rate to the electronic switch. By adjusting the pulse width, I could vary the number of revolutions that the motor would turn for a given load, thus varying the incremental movement of the rotor. Also, the pulse rate could be adjusted to vary the number of incremental steps that occurred each second or, in other words, vary the rotor speed.

the circuit

The schematic of the pulse-control unit, integrated with the circuit of a CDE Tailtwister rotor, appears in **fig. 2**. All control-unit parts are numbered starting with X20 so they won't be confused with those that are part of the Tailtwister rotor.

Power supply. Parts CR20, C20, C21, and VR20

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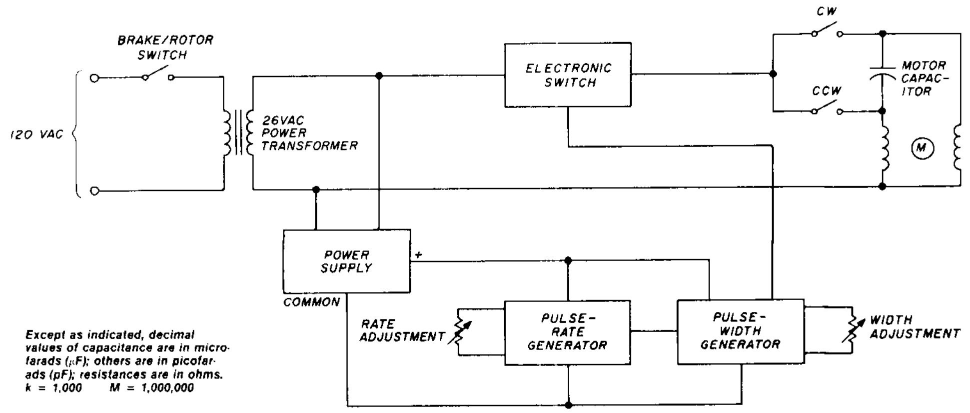


fig. 1. Block diagram of the pulse-control unit added to the Tailtwister motor-control circuit. Brake solenoid and end-of-travel limit switches have been omitted for simplicity.

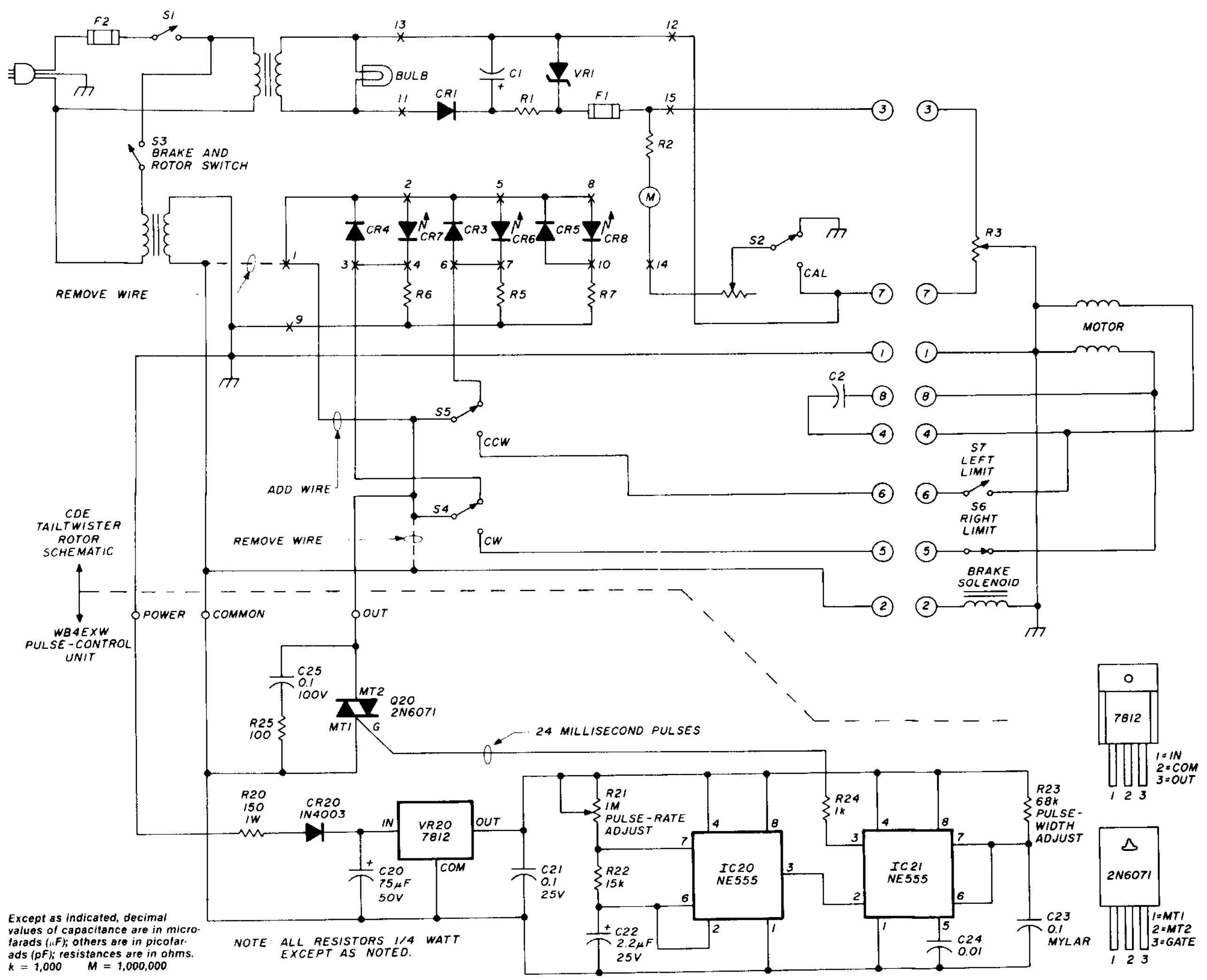
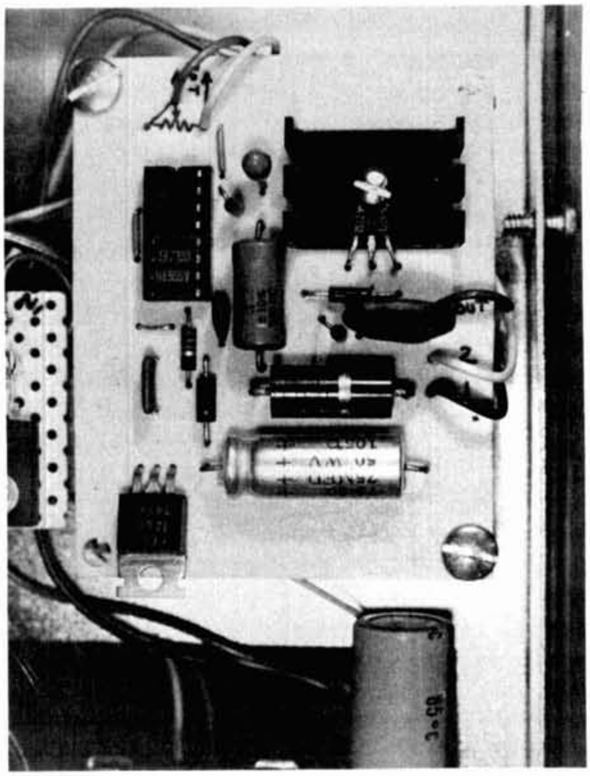


fig. 2. Schematic of the pulse-control unit and Tailtwister control box. The PCU schematic is below the dashed line that separates the two circuits. Note wiring changes to control box.



Close-up of the pulse-control unit showing parts placement. Unit is mounted on the underside of the Tailtwister control box chassis above the power transformer, as shown. Note that VR20 leans slightly to clear the bottom cover when installed.

define a regulated 12 Vdc power supply for the PCU. Resistor R20 decreases the 26 Vac from the rotor power transformer somewhat, so that the rectified input voltage to VR20 will be below its 35-volt maximum. *Do not* omit R20. You may notice that the ac input leads to the power supply section seem to be hooked up backward. The lead from R20 (PWR) connects to terminal 1 of the eight-terminal output connector of the rotor control box. It is grounded to the chassis. The pulser unit common lead is connected to terminal 2. These power leads *must* be wired in this manner to provide proper phasing to the electronic switch.

Pulse-rate generator. The pulse rate generator is made from IC20, an NE-555 timer wired as an astable multivibrator. Resistors R21, R22, and capacitor C22 determine the frequency of oscillation of IC20. R21 is a 1-meg pot, which can be a panel-mount type for varying the pulse rate. Or it could be a trimpot mounted on the PC wiring board for those who like set-it-and-forget-it controls. The minimum pulse rate was set to about 0.5 Hz.

It's possible to allow IC20 output to provide the control signal to the electronic switch. But if you investigate the inner workings of an NE-555 used in the astable mode, you'll find that the duty cycle also varies as the frequency of oscillation is changed. I wanted constant-width pulses applied to the electronic switch, so that the incremental movement of the rotor could be controlled. Thus, IC21, another NE-555, was added as a monostable multivibrator.

Each time IC20 output (pin 3) goes low, it triggers IC21. IC21 output (pin 3) then goes high and turns on the electronic switch. IC21 pin 3 stays high for a period determined by R23 and C23. The output pulse width is determined by:

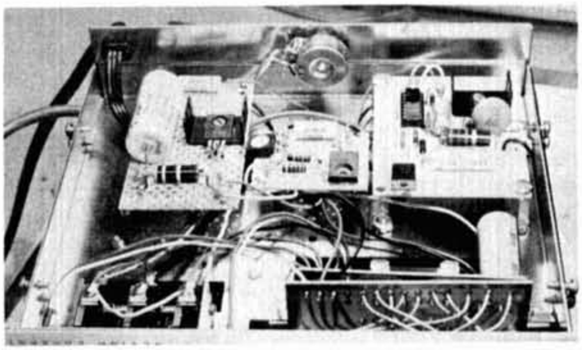
$$T = 1.1\,RC \qquad (1)$$

where T = Output pulse width (seconds)
R = R23 resistance (ohms)
C = C23 capacitance (farads)

Realize that the tolerances of C23 and R23 may make **eq. 1** seem invalid; actually, the formula is quite accurate.

Pulse-width calculations. The pulse width in my unit is about 24 milliseconds. The 0.1-μF cap I used for C23 supposedly had a tolerance of $\pm$10 per cent. Using **eq. 1**, I should be using a 200k resistor for R23. However, I ended up with a 68k to obtain the 24-millisecond pulse width. (The capacitor was way out of tolerance or incorrectly labeled.) Rough measurements indicate that this pulse width gives an incremental rotor travel of about 0.1 degree for my antenna system. By varying the R23 value, you can change the pulse width to suit your needs.

Electronic switch. The last part of the pulse control



Overall view of the PCU and regulator module installed in the control box. The smaller board next to the PCU is the regulator module. Also shown is the external rate-adjustment pot mounted on the rear panel. Azimuth-indicating meter in this control box has been modified to a digital readout for improved accuracy.

unit to be described is the electronic switch. A triac, Q20, gave simple, full-wave ac control to the rotor motor. Output pin 3 of IC21 going high will turn on the triac. The triac will pass current to the motor so long as it has gate drive. When gate drive is removed, the triac will turn off the next time that the ac current through it passes through zero. The suppression network, made up of R25 and C25, which is across the main terminals of the triac, was added to prevent triac false triggering. In some cases, where a triac is used to control an inductive load, there's enough phase difference between the voltage and current waveforms to cause the triac to retrigger when the gate drive is removed. Later I found that the suppression network was not needed in this application, but I left it in the circuit as a precaution.

I also found that I didn't need additional circuitry to make the triac switch at the zero points of the ac voltage waveform. No radio frequency interference has been noted.

printed wiring board

A full-size layout of a printed wiring board for the pulse control unit is shown in **fig. 3**. It's designed to mount on two 8-32 by 1/2-inch (M4 by 12.5 mm) threaded spacers that are screwed down onto the existing mounting bolts for the power transformer in the Tailtwister control box (see photo). Solder pads are provided for external connections. Don't forget to install the three jumpers on the printed wiring board.

The two NE-555 timers mount in one 16-pin IC socket. Both are oriented in the same direction as shown in the parts layout (**fig.4**). The heatsink used for the triac is a Thermalloy 6107B. A similar heatsink can be made from a piece of scrap aluminum. If a substitute is made for the 2N6071 triac, be sure that it has the same pinout and triggering characteristics.

Notice that the voltage regulator must be tilted somewhat to clear the bottom cover of the rotor control box. It does not require a heatsink. Resistors R22 and R25 mount vertically. C22 is a small dipped tantalum capacitor, and C23 should be a Mylar capacitor for stability.

The pulse-rate generator control pot can be a printed wiring board type such as a CTS X201 or can be a panel-mount pot as shown in the photo. If an external pot is used, wire it so that clockwise rotation of the shaft decreases the total resistance. Then the pulse frequency and rotor speed will increase with clockwise rotation.

installation

Only five wires must be soldered to the printed wiring board (three if you mount R21 on the board) for

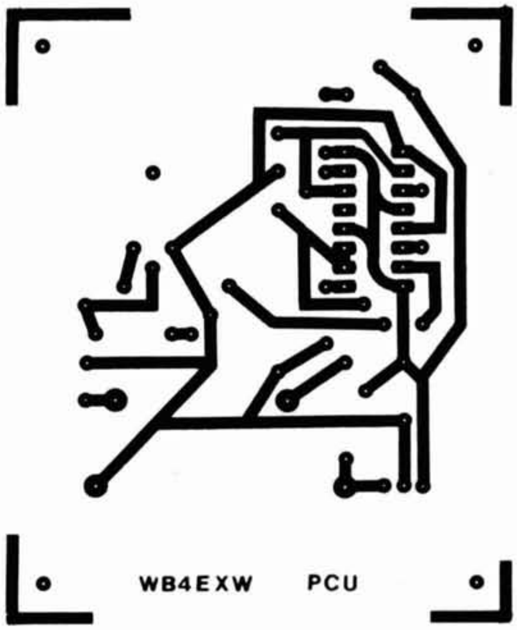


fig. 3. Full-size printed wiring board layout for the pulse-control unit. Wire side is shown.

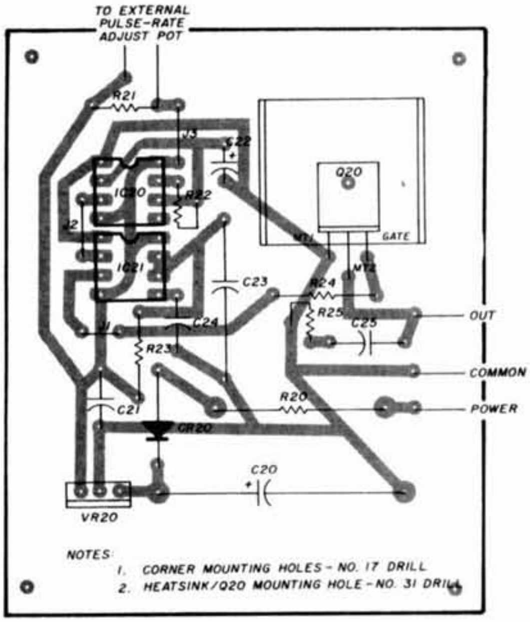


fig. 4. Parts placement for the pulse-control unit. IC20 and IC21 mount in a single 16-pin dual inline socket. R22 and R25 mount vertically. Do not omit the three jumpers (labeled J1, J2, J3). A trimpot can be used for R21, or leads can be extended to a panel-mounted pot.

external connections. Solder two wires, preferably color-coded, to the pads marked PWR and COM. Make sure the wires are long enough to reach to the rotor control box eight-terminal connector block. Solder another wire, that will reach to the CW and CCW control switches, to the pad marked OUT. The last two wires are for the external pulse-rate adjust pot; their length depends on where the pot is mounted.

Before permanently mounting the pulse-control unit, you must make two minor wiring changes in the Tailtwister control box. You must first remove the existing wire that connects terminal 2 of the eight-terminal connector block to the common terminals of the CW (S4) and CCW (S5) control switches. Leave the jumper that connects the common terminals of S4 and S5 together in place.

Next unsolder the wire from terminal 2 of the connector whose opposite end is soldered to point 1 of the Meter/LED printed wiring board. Connect the end of this wire that was on terminal 2 to the common terminal of either S4 or S5. This wiring change avoids a triac gating problem due to the LED indicator circuitry. The LEDs will still be functional but in a slightly different manner, as discussed later.

Mount the printed wiring board on its spacers over the power transformer. Solder the PWR wire to terminal 1 and the COM wire to terminal 2 of the eight-terminal block. Connect the OUT wire to the common terminal of either the S4 or S5 control switch. The last two wires connect to the pulse-rate adjust potentiometer. This completes the hookup of the unit.

checkout

Check all wiring against **fig. 2** and the instructions in the previous section to uncover any possible wiring errors. Next, set the rotor-control box on its side so you have access to both pulse-control unit and eight-terminal connector. Set the pulse-rate adjust pot to mid position.

Turn on the control-box power switch. This should cause the meter to indicate existing rotor position and become illuminated. Connect the common probe of an ac voltmeter to terminal 1 of the eight-terminal connector, and connect the input probe to terminal 2. Press the BRAKE RELEASE switch on the control box. This should cause the brake to release; about 26 Vac will be indicated on the voltmeter. The green LED that normally indicates brake release should light, but it should flash on and off. (It will flash at the pulse rate set by R21.) The brake solenoid now is receiving full power, but the LED flashes so you can set the desired pulse rate before applying pulsed power to the rotor motor. Vary the pulse rate adjust pot to see that the LED flashes from about 0.5 Hz to full on. If this checks out, proceed to the next step.

Leave the common voltmeter probe connected to terminal 1 and connect the input probe to terminal 6. Press the BRAKE RELEASE switch, then press the CCW control switch. Pulsed 26-Vac should show on the meter (the voltmeter will not actually indicate 26 Vac because of the short pulses). Both green brake and red CCW LED indicators should flash with the 26-Vac pulses. Your antenna should rotate counterclockwise at a much slower rate than before.

Next, remove the probe from terminal 6 and connect it to terminal 5. Press the BRAKE RELEASE switch again, then press the CW control switch. Pulsed 26-Vac should show on the voltmeter; the brake and CW LEDs should flash, and your antenna should slowly rotate clockwise.

If you have problems, recheck all wiring and use the schematic and circuit-board description as a troubleshooting guide. When checking the operation of the pulse-control unit, be sure to use terminal 2 and the *common point* for voltage measurements. Also note that the pulse-control unit receives no power unless the BRAKE RELEASE switch is operated.

This modification is made to a CDE Tailtwister rotor, but is also applicable to HAM-type rotors. In fact, I initially checked out the pulse-control unit in my HAM-2 rotor. All wiring changes in the rotor-control box are the same except for the LED indicator wire change. Since a HAM-2 or -3 has no LED indicators, this change doesn't have to be made.

position circuit modification

You've probably noticed the small circuit board mounted next to the pulse-control unit (photo). It is a 13-Vdc regulator module that replaces the zener diode and current-limit resistor on the Tailtwister printed wiring board. This circuit offers a more stable regulated source for the azimuth indication circuitry. The schematic (**fig. 5**) uses a simple 8-volt, 3-terminal regulator. The 500-ohm trimpot allows the output to be adjusted to 13 volts. The filtered dc supply is about 30 volts and is close to the maximum input voltage rating of the 7808 regulator, so a 150-ohm, 1/2-watt resistor was added to reduce the regulator input voltage. An added benefit is reduced power dissipation. Mine was built on a small piece of perf board, mounted on a standoff in an existing hole in the control-box chassis.

To wire the regulator module, first remove R1, the 390-ohm, 2-watt resistor, and VR1, the 13-volt zener from the rotor printed wiring board. Don't use much heat, as the traces on the printed wiring board are fragile. Next, connect the module IN terminal to the pad left by R1 that connects to the positive lead of C1. Connect the OUT terminal to the pad left by R1

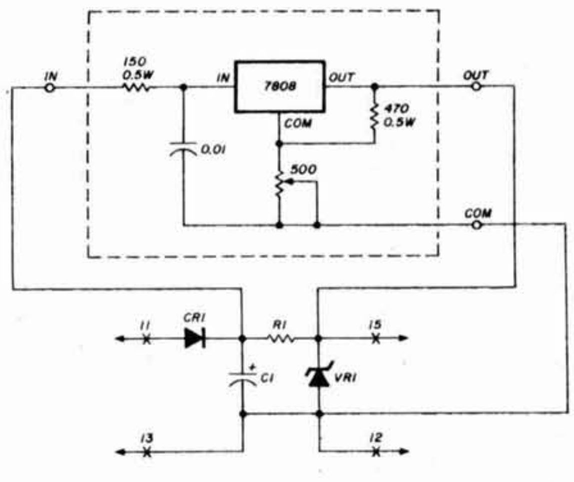


fig. 5. Schematic of the 13-volt regulator module for the rotor position is contained in the dashed box. Also shown are the connections from the module to the original circuit. R1 and VR1 are removed before connections are made. Adjust the trimpot for 13 Vdc across terminals 3 and 7 of the control box.

that connected to VR1 cathode. Solder the COM terminal to the pad that previously contained the VR1 anode lead. Turn the control box on, and adjust the trimpot for 13 volts across terminals 3 and 7 of the connector block.

conclusion

This modification to the Tailtwister rotor system has made controlling my large EME array a dream. Variable speed control, from 360 degrees per minute down to much less than one degree per second, allows turning torque and positioning overshoot to be minimized. The system would now lend itself well to microprocessor control; thus, this modification should also be of interest to Amateur satellite users.

My rotor-control box has also been incorporated into a digital readout similar to that described by K1DG.[1] Positioning accuracy has been increased, as errors of 5 degrees or more existed in the indication circuitry alone in my control box before the digital indicator was added. Now my position accuracy seems to be limited mainly by a) the potentiometer linearity, and b) the accuracy to which I can set the antenna to a calibrated point.

reference

1. Doug Grant, K1DG, "Digital Readout for the HAM-3 Rotator, *ham radio*, January, 1979, pages 56-59.

ham radio

microcomputer-based contest keyer

During the last CW contest, did you wish you had a memory keyer to send the contest exchange and give you more time for log and dupe-sheet entry? Did you wish you could automatically transmit the proper QSO number and generate any contest exchange? The microcomputer-based contest keyer may be the answer. This fully iambic keyer has ten programmable messages, a four-digit QSO counter, and the flexibility to transmit almost any contest exchange. An added feature is the display of the current code speed setting.

Several years ago I used tape-recorded tones to drive a relay and key my transmitter during a CW sweepstakes. Since this first attempt at automation, I built several memory keyers,[1] a TTL CW QSO number generator, and an 8080 microprocessor keyer that required three circuit boards. Because of the decreasing cost of microcomputer chips, I thought the time had arrived to design a state-of-the-art contest keyer.

Several operational objectives were set for the design. First, it should be an iambic keyer having dot and dash memory plus the optional forced-letter space of the popular WB4VVF *Accu-Keyer.*[2] For simple programming, the user should be able to load messages directly in code from the keyer paddle. Next, there should be several options such as automatic QSO number and RST generation. Last and most important, the keyer should be simple and natural to use but be flexible enough for any contest exchange. The result is the microcomputer keyer presented here.

circuit description

Keyer construction is shown in the lead photo and **fig. 1**. Instead of a large keyboard, I selected a simple, twelve-button keypad for control. A potentiometer is included to adjust code speed and a four-digit LED display is included for QSO number and other alpha-numeric keyer messages. Simplicity results from using an Intel 8748 programmable microcomputer. The advantage of such a device is the ability to perform complex tasks with minimum circuitry. Software may be reprogrammed for other features or options. The following sections describe the design, discuss operation, and outline software routines for the keyer functions.

The keyer circuit, **fig. 2**, requires only seven integrated circuits and six transistors. The 8748 microcomputer, U1, contains 1024 bytes of electrically programmable read-only memory (EPROM), 64 bytes of random access memory (RAM), a programmable counter, and three 8-bit input/output ports. Since a complete description of the 8748 was recently published, it will not be repeated here. Readers should review the article by N6TY[3] and the Intel *MCS-48 Microcomputer Users Manual*.[4]

The 8748 has limited RAM for this design. U2, an Intel 8155, is used to expand the message buffer area. The 8155 contains 256 bytes of RAM, three input/output ports, and a 14-bit counter; only the RAM section is used. While this may seem a waste of internal functions, the 8155 will directly interface with the 8748's multiplexed data and address lines, DB0 to DB7. Additional interface circuitry is not required, saving at least two additional chips and considerable wiring.

U3 is an Intel 8279 programmable keyboard/display interface.* This versatile device can drive up to thirty-two multiplexed 4-bit display digits and simultaneously decode a 64-button keyboard. This design uses the interface to drive four 7-segment LED displays for alpha-numeric messages to the user. It also scans the 12-button keypad, decoding the key and signalling U1's IRQ (Interrupt Request) input with a logic 1. An internal first-in, first-out (FIFO) buffer stores up to eight keypresses; the user can "type ahead" without losing inputs.

Display multiplexing of the common-anode 7-segment display reduces the number of interconnect wires from 32 to 11. Anodes are switched by Q3 to Q6 from U6 inverting SL0 to SL3 from U3. Scan out-

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*U2, U3 *functions* are programmed by commands from U1; the EPROM within U1 contains the actual program for the entire keyer.

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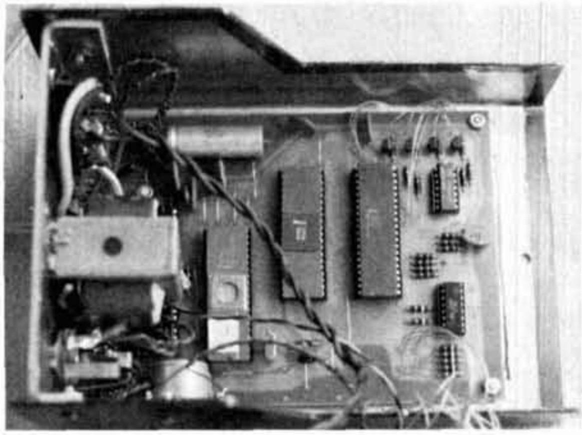

fig. 1. The keyer electronics are contained on one single-sided printed circuit board.

puts SL0 to SL3 are also used for the switch rows in the keypad. Segment driving is accomplished by A0 to A2, B0 to B3 from U3 through U7. Scanning is invisible to the user.

Keypad switch columns are connected to RL0, RL2, RL4, and RL6 of U3 for key decoding. U3 provides debouncing and key rollover detection internally. The keypad may be a conventional telephone type with separate switch contacts.

command and control

U1, U2, and U3 are all connected to the same 8-bit data/address bus, DB0 to DB7, and the same Read (RD) and Write (WR) control lines. Data input or output for the microcomputer will be enabled by a logic 0 on either U1-36 (chip select for U3) or U1-37 (chip select for U2). U1-35 is the Control/Data (C/D) input for U3; Control mode (logic 1) is for display scanning, Data mode (logic 0) reads the keypad and loads the display.

The keyboard/display interface requires a clock input for display multiplexing and other timing operations. This is supplied by U1's Address Latch Enable (ALE) and is equal to the crystal frequency divide by 15. This signal eliminates a separate display oscillator. It should also be noted that U1 contains an internal clock oscillator circuit, requiring only an external crystal.

Dual timer U4 functions as both code-speed oscillator and sidetone generator. The code-speed section, U4A, is enabled by a logic 0 on U1-34. The same enable signal energizes clock indicator CR1 to indicate that the code clock is on. Code speed output, U4-5, is connected to U1's internal 8-bit counter. The microcomputer will set the length of one dot as 100 code speed pulses during CW operations. U4A oscillates at 100 times dot rate and is set by R1; speed range may be altered from the keypad.

Transmitter keying occurs when U1-33 is logic 0 (U5-10 at logic 1). The Q1, Q2 circuit is used with grid-block keying on my transmitter. Key-up voltage at the TX line is about −30 Vdc. Key-down voltage is close to ground. Other keying methods will require changing the Q1, Q2 circuit.

The microcomputer interrogates the key dot and dash paddle switches through I/O ports at U1-27 and U1-28. Internal pull-up resistors allow direct key connection.

U1 will automatically reset to program start when turned on. A manual reset push button, S1, is included for convenience. CR3 and CR4 are program status indicators and are described later.

Sidetone oscillator U4B is optional. The speaker may be any small unit with 8-ohm impedance or higher. CR2 indicates key down and should be retained; a keyer option function allows holding key down for tuning.

software routines

Microcomputer hardware is useless without a program to make it "come alive". Software routines are described, but the detailed program is too long to include here. A complete program listing is available from the author.†

A flowchart of the main software polling loop appears in **fig. 3**. At power on or RESET, the keyer status flags are initialized and -HI- appears on the 4-digit display. The program enters a loop that continually checks the key paddle switches and keypad switch status, waiting for some request. Any switch closure will enable a branch to a routine that will generate the proper message or code element.

For example, assume the dot paddle switch is closed. The microcomputer turns on the transmitter for one dot period. It also checks the dash paddle switch and sets the dash memory flag if the dash switch was closed before the dot was completed. If so, a dash will begin after the dot space. Holding either dot or dash paddle closed will generate successive dots or dashes with proper spacing.

If no key paddle switches are closed, the microcomputer checks the letter space option. If enabled, no paddle inputs are processed for two dot periods, corresponding to one letter space interval. Key paddles are still tested during this interval. If either is closed during the interval, the proper flag is set and the appropriate dot or dash is sent after interval com-

†A copy may be obtained from the author for $3.00, which covers reproduction and postage costs.

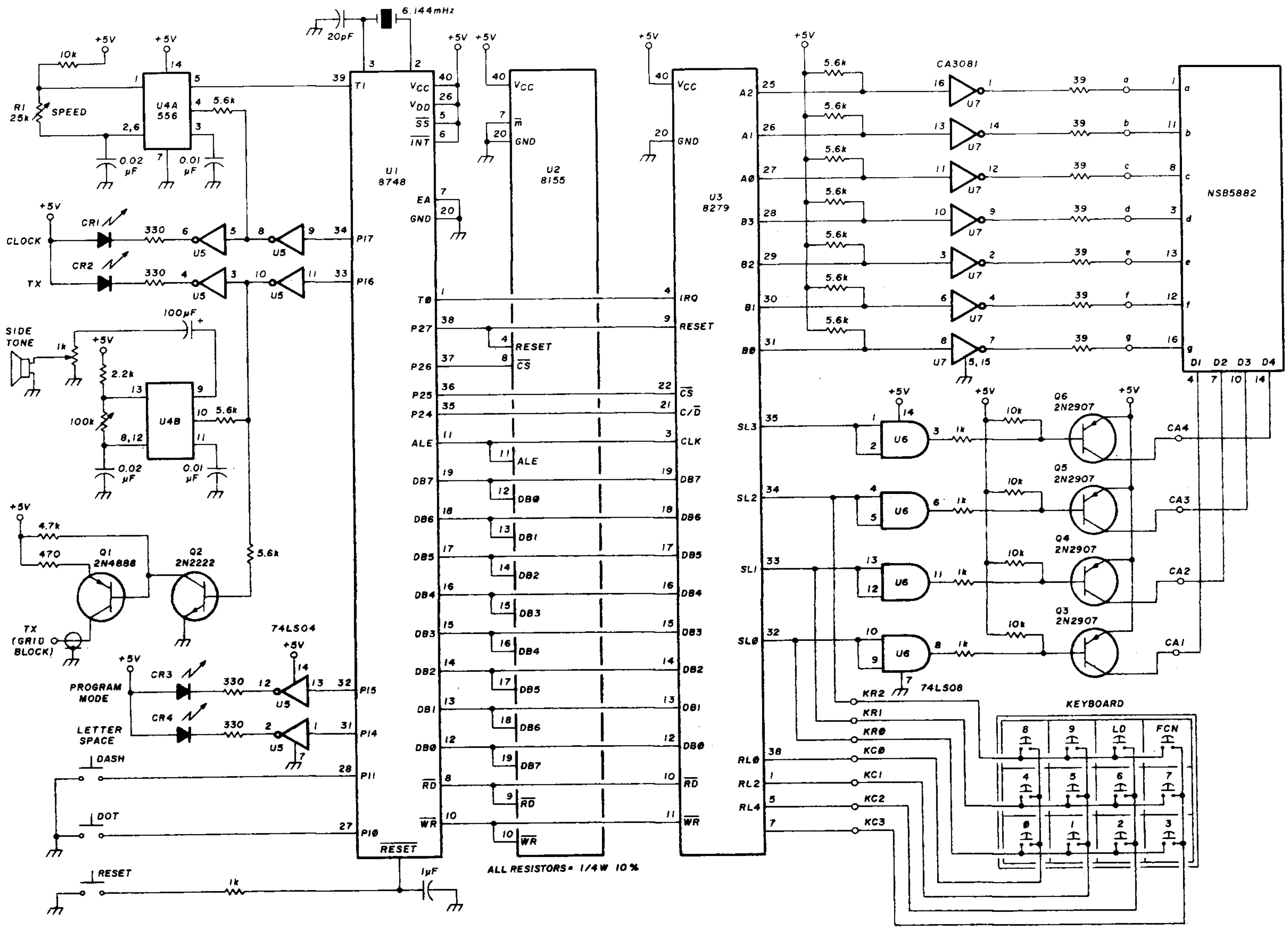

fig. 2. The keyer schematic. The four-digit LED display module may be replaced with four individual seven-segment common-anode digits.

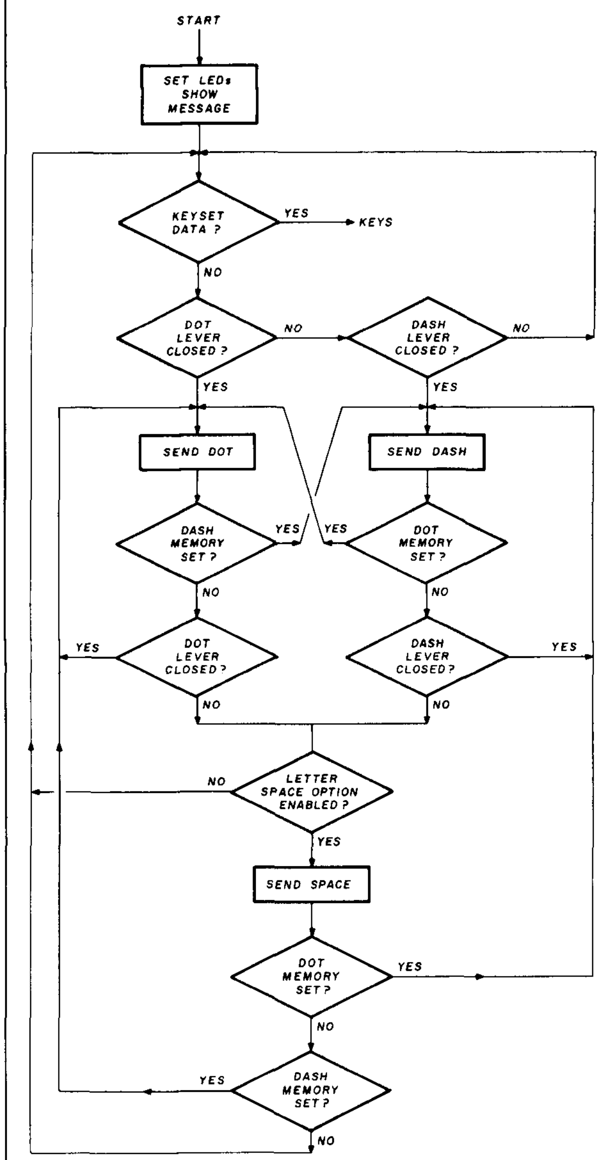

fig. 3. A flow chart for the microcomputer keyer program. The keyer is fully iambic and has an optional forced letter space.

pletion.* If no requests are pending, the program returns to the main polling loop and waits for the next input.

Pressing one of the keyboard switches calls a keypad processing routine. This routine decodes the keypad and executes one of several routines shown in the flow chart of **fig. 4**.

function keys

To set various keyer options, the keypad F (FUNCTION) key is pressed, followed by 0 through 9 for the selected option. Pressing F will display -F- with the third digit blank to indicate an option number must be entered next. Entering the option will fill in the display and execute the function.

Table 1 summarizes function-key options. The first four, F-0 through F-3, will display, increment, decrement, or load a four-digit QSO counter. If QSO count load is selected, the number keys are used to enter an initial count. Once this initial count is entered, pressing the F or L (load) key will load it into memory as the current QSO count.

table 1. Function key options are activated by first entering F on the keypad, then the option number. Function F-9 may be terminated by pressing any key or key lever.

keypad entry	description
F-0	display QSO total
F-1	increment and display QSO total
F-2	decrement and display QSO total
F-3	load the QSO total
F-4	toggle forced letter space option
F-5	toggle 3-digit QSO number option
F-6	toggle -HI-/QSO-total display option
F-7	display code speed in wpm
F-8	toggle code speed range
F-9	tune transmitter (key-down hold)

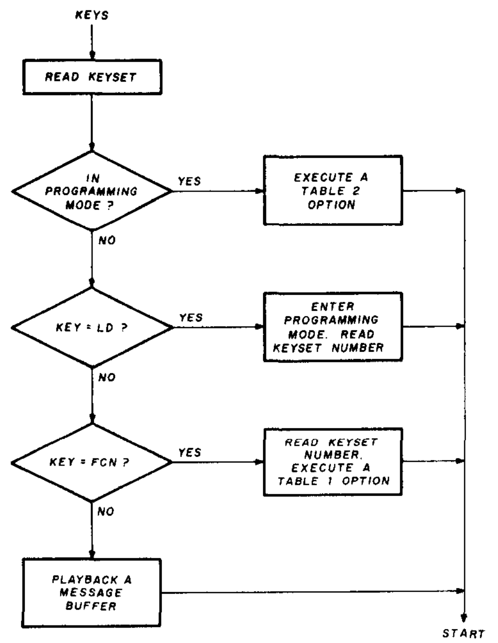

fig. 4. The operation performed by each key depends on the current keyer mode. For example in the message programming mode, the table 2 options are selected.

*The microcomputer is able to do several things seemingly at once due to execution speed and internal IC circuitry.

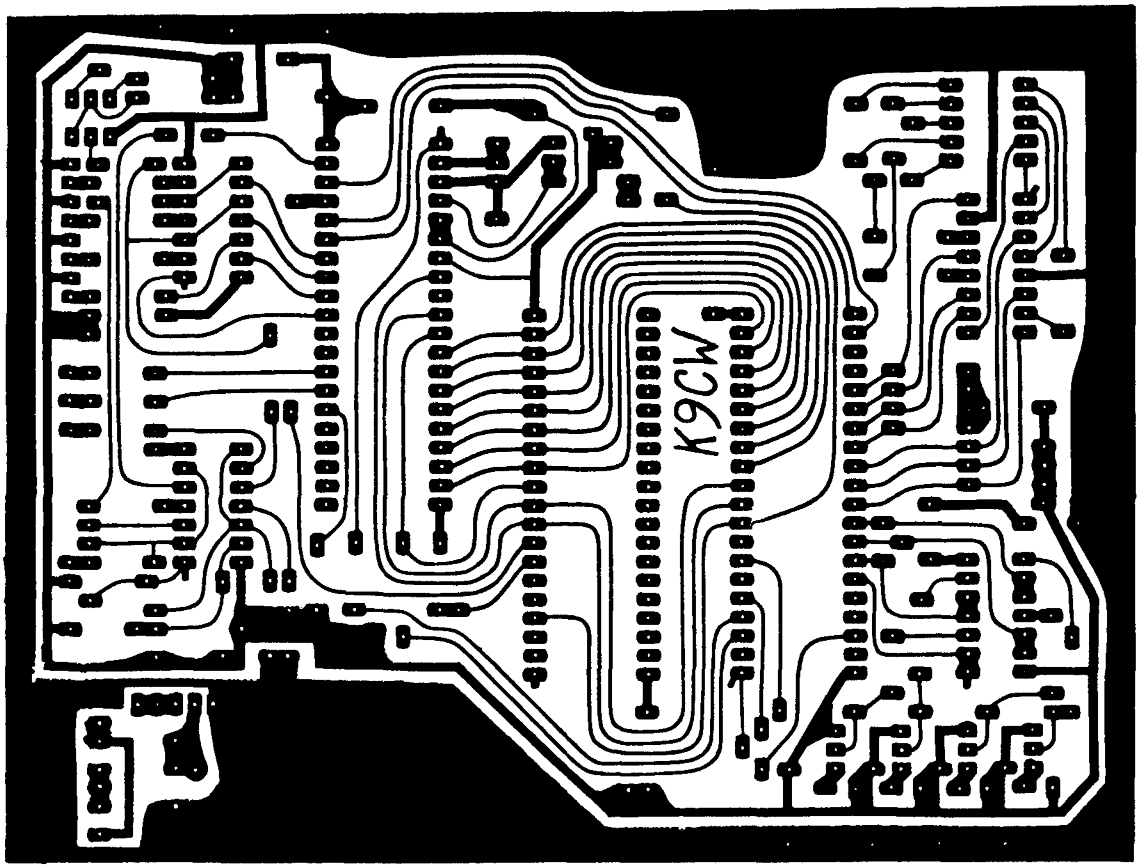

fig. 5. Printed circuit board pattern for the keyer shown from the foil side.

Options F-4, F-5, and F-6 select various keyer operating modes. Letter space option F-4 will force a letter space if neither dot nor dash is closed upon completion of a code element. Front panel indicator CR4 will light if this function is enabled. Option F-5 determines if QSO numbers should always be transmitted as three digits or if the leading zeros in the number should be suppressed. Option F-6 allows display of QSO total instead of -HI- during the input waiting period. Default values at turn-on are: Letter space enabled, suppress leading zeros, display -HI-.

Current code speed in words per minute is displayed by the F-7 option. Since the microcomputer is continuously measuring speed, speed control R1 can be adjusted through digital readout. Pressing any key or paddle lever will exit this function.

Option F-8 selects speed range, either 16-60 wpm or 5-30 wpm. Default value is 16-60 wpm. Option F-9 will key down the transmitter for tuning purposes; key up occurs on pressing any key or the paddle. Transmit indicator CR2 will be on and -F9- displayed during key down.

loading messages

To load one of the message buffers, the L key is pressed, followed by a selected message number of 0 through 9. Entering only L will display -L- with a blank third digit prompting the user for a message number. Entering the number will fill in the display such as -L5- for message number five.

Once the load identification is complete, the manually sent code is stored in memory until either F or L is pressed to terminate the message. The number of memory bytes per message is displayed during message entry. Message buffers 0 through 9 are contiguous in U2 so that individual messages can be up to 25 bytes long without interfering with the next higher message number.

Twenty-five bytes can store 100 dots, dashes, or spaces, but this is not a limit. Because the keyer allows variable message length, one long message containing over 1000 dots, dashes, or spaces is possible; this is an average of 200 letters at three dots or dashes per letter.

In addition to manually entered code, several special functions can be loaded into a message. **Table 2** gives the function options and a specific display for user confirmation. CR3 will be on during message loading.

table 2. Message loading options are activated by a single keypad entry. These options insert special functions into the message during message loading.

keypad	display	description
0	-C0-	send QSO total
1	-C1-	increment and send QSO total
2	-59-	send 5NN or 5 (keypad) 9
3	-IP-	send keypad numbers
4	-SP-	insert a letter space
5 - 8		ignored; reserved future use
9	-00-	restart message loading
L or F		terminate a message load

special message options

During playback, option 0 transmits the current QSO total as two to four numbers depending on leading-zero suppression previously set by F-5. Option 1 first increments the QSO total then sends the number.

Option 2 transmits an RST in one of two formats. If a key has been typed during message playback but before this option is encountered, that keypad number is sent as the middle digit (signal strength) in the RST. If no keypad number has been pressed, 5NN is transmitted.

Option 3 allows transmitting a number directly from the keypad. The keypad FIFO buffer is checked when this option is encountered. If any keypad numbers are stored (maximum of 8), they are sent, the FIFO empties, and message playback resumes.

Option 4 simply inserts one letter space each time it is entered. Option 9 allows the user to restart loading in case of error. Options 5 through 8 are currently ignored, reserved for future additions.

These options allow almost any contest message to be loaded. For example, 599001 can be loaded by typing 2, then 1. Keypad 2 calls for the 5NN; keypad 1 calls for the QSO total (incremented from zero, no suppression of leading zeros). NR 100 IL is loaded by manually keying NR, pressing 1 on the keypad, pressing 4 twice for two letter spaces, then keying IL manually.

During message loading, a routine measures the time between manually keyed dots and dashes. It corrects most spacing errors. No more than one word space interval will be automatically stored in memory. The user can stop loading without filling the memory with spaces. No word space is automatically inserted after loading options. Inserting spaces (option 4) permits storage of perfect Morse code.

message playback for transmission

A stored message is played back by simply pressing the appropriate keypad number. If message number seven is selected, the display shows -P7-. If one of the QSO number options is encountered in a message, the current or incremented QSO count is displayed. The keyer returns to the main polling loop when a message is complete and displays -HI- or the current QSO count (if F-6 is set).

Manually sent code can be inserted at any time during a message playback. When a paddle switch is closed, the keyer stops and waits for a word space interval. If the paddle switch is still closed, manual code transmission begins. When manual code ceases, the stored message playback continues. Message playback can be halted at any time by pressing the RESET button or briefly tapping the dot or dash levers.

A common method of storing Morse code in a digital memory is to use a length equivalent to each element. A Morse dot would appear in memory as binary 010. A Morse dash would be binary 01110. This keyer uses two bits for each element: A letter space is binary 00, a dot is binary 01, a dash is binary 10. Binary 11 is used as an option indicator for one of the message options. This assignment allows four code elements per memory byte.

construction

The circuit may be built on the PC pattern of **figs. 5** and **6** or with wire-wrap techniques. Sockets are suggested for either case to avoid IC damage. Space has been provided on the PC layout for diode rectifiers and a filter capacitor. A three-terminal voltage regulator should be used with an output rating of +5 Vdc at 0.5 amperes. Mount it on the cabinet for heat sinking.

Several 0.1 μF disc capacitors are used on the PC board to bypass the +5 volt line at each IC and reduce logic switching noise. A wire-wrap version should have these capacitors installed first with minimum lead length.

The finished keyer can be mounted in a commercial case or a homemade one such as in the title photo. Some users may want to separate the keypad to move the main keyer circuitry off the desk. In either configuration, be certain to rf bypass all leads into or out of the keyer cabinet.

Locating components isn't difficult.* Most of the

*Drilled and etched PC boards and some parts are available from RADIO-KIT, Box 411, Greenville, N. H. 03048.

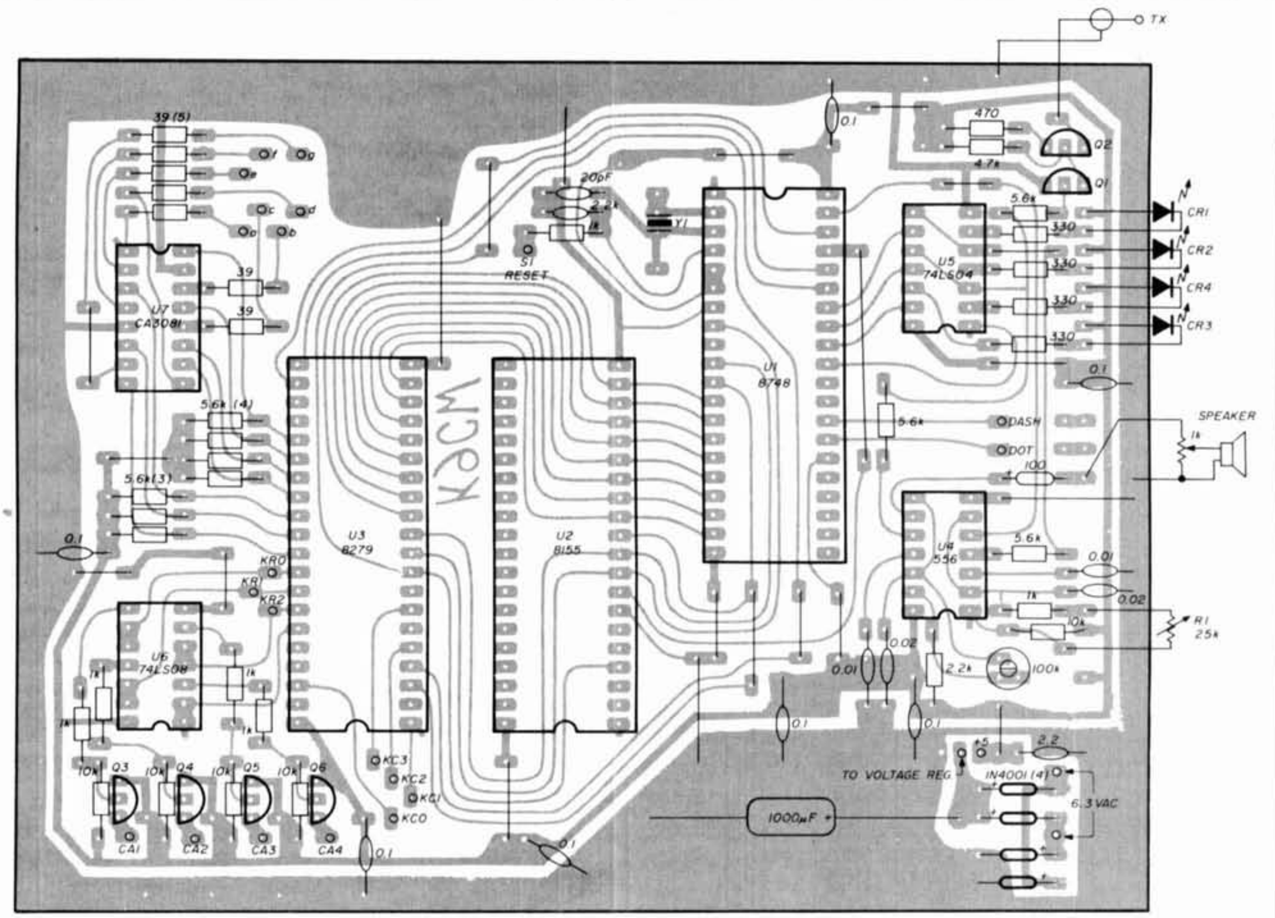

fig. 6. Component layout for the keyer shown from the component side of the board. Pads are provided for several 0.1 μF bypass capacitors on the +5 volt supply line.

computer hobbyist advertisers carry the Intel components. The four-digit LED display is a National Semiconductor NBS7882, but four separate seven-segment, common-anode displays may be substituted in place of that unit.

The 8748-8 is a lower-speed version that can use a 3.58 MHz color TV crystal; the program listing includes modification for this option. Either crystal should be series-resonant with series impedance less than 75 ohms for 6.144 MHz; less than 180 ohms for 3.58 MHz.

The 8748 is supplied unprogrammed. You must locate a nearby programming facility with an Intel compatible EPROM programmer. EPROM programming of the 8748 requires more care than a conventional hobby computer PROM burner. Intel supplies all information on EPROM programming in reference 4.

conclusion

This keyer has been used in only a few contests so far. I find the automatic number generation and multiple message playback features to be useful operating aids. Automatic message transmission during each exchange is an advantage that allows filling in the dupe sheet and checking off section multipliers. In addition, errors are never made in transmitting the QSO number or other parts of the contest exchange.

My objective in writing this article is not only to describe a powerful, inexpensive contest keyer, but also to start people thinking about new applications for microcomputers in Amateur Radio. I'd be interested in your ideas for using this new technology. Microprocessors and computers are already appearing in common ham equipment. Several manufacturers sell combination RTTY and Morse keyboards and displays based on microprocessors.

Perhaps a future version of the contest keyer will be able to scan the band, copy a call, check the dupe list, and make a contest contact automatically. While that operation may be feasible, I don't believe it's desirable. After all, what would remain for the contest operator to do?

references

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2. J. Garret, WB4VVF, "The WB4VVF Accu-Keyer," *QST*, August 1973, page 19.
3. J. Beaston, N6TY, "CW Trainer/Keyer Using a Single Chip Microcomputer," *ham radio*, August 1979, page 16.
4. *MCS-48 MICROCOMPUTER USER'S MANUAL*, Intel Corporation, 1977.

ham radio

More Details? CHECK—OFF Page 110

analog-to-digital display converter
for the visually handicapped

The circuit described in this article was designed for a visually impaired Amateur to replace his analog panel meter with a large digital display. A variation of the circuit can be used to drive a tactile decimal readout for the blind. This is the primary use of this circuit, but with a little ingenuity it can be adapted to convert any analog device (such as the frequency-readout dial on Amateur gear) to a digital display. For instance, it could be used as an azimuthal readout for an antenna rotor, calibrated directly in degrees of bearing. If so desired, using a thermistor, it could monitor the temperature of a final amplifier or linear amplifier tube compartment, crystal oven, or anything where temperature is critical.

theory of operation

All analog meters are designed with a specific internal resistance and are constructed to display a particular current as "full scale." It is easy to calculate the voltage necessary to produce a full-scale reading, using Ohm's Law, $E = IR$. Thus a 100-ohm, 1-mA

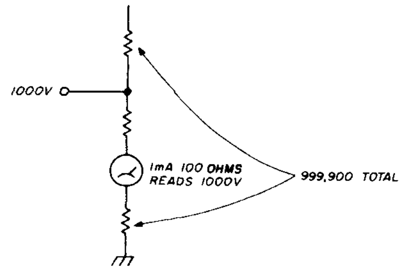

fig. 1. Example using a meter that reads 1 mA full scale. To obtain a full-scale reading of 500 mA with this meter, a 0.2-ohm resistor is connected across the meter, which has an internal resistance of 100 ohms. *Eq. 1* explains how to do it.

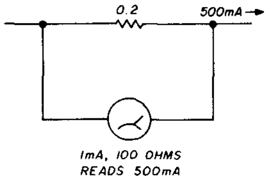

fig. 2. Another example showing how to obtain a full-scale range of 1kV using a voltage divider, with the meter supplying part of the resistance.

full-scale meter requires $100\ ohms \times 0.001\ ampere = 0.1\ volt$ for full scale; a 1000-ohm, 100-micro-ampere full-scale meter requires $1000\ ohms \times 0.0001\ ampere = 0.1\ volt$, and so on.

To obtain a higher full-scale range for a meter, construct a divider network or shunt network into which the meter is connected. Thus, to obtain a full-scale reading of 500 mA, using a 100 ohm/1 mA full-scale meter, calculate the resistance, to be placed in series with the current through the meter, that will develop a voltage drop of 0.1 volt (the full-scale voltage of the meter) at 500 mA. Using Ohm's Law again, $R = E/I = 0.1V/0.5A = 0.2\ ohm$. (See the circuit in **fig. 1**.) Likewise, to obtain a full-scale range of 1000 volts, construct a voltage divider with the meter as part of the divider. Again, $R = E/I = 1000V/0.001A = 1\ megohm$. See **fig. 2**.

If we replace the meter in **fig. 2** with a 100-ohm resistor, we can measure the voltage produced across the 100-ohm resistor and construct our voltmeter so that 0.1 volt reads 1000. That's basically what we're

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doing with this converter circuit, using a special analog-to-digital voltmeter.

It's beyond the scope of this article to go into the various methods to convert an analog signal to a digital output. Numerous texts are available covering the subject. Some of the best are the Motorola publications listed at the end of this article.

device selection

This circuit uses the Motorola MC14433 3-1/2 digit A/D converter. The MC14433 is a low-power linear and digital CMOS monolithic dual-slope A/D converter packaged in a 24-pin IC. It provides an accuracy of ±0.05 per cent of reading plus-or-minus one count and provides up to twenty-five conversions per second. The input resistance is 1000 megohms, and its outputs are standard B-series CMOS that will drive one low-power Schottky load. Power consumption is only 8.0 mW typical at 5.0 Vdc. For those interested, the data sheet contains schematics for several voltmeters, including an auto-ranging multimeter with ac and dc full-scale ranges from 200 mV to 200 V; 2 mA to 2 A; and 2 k to 2 megohms.

This circuit uses a finely adjustable reference voltage for the MC14433 to force it into giving any full-scale reading desired (up to 1999). By switching in a number of reference voltages, we can obtain a meter with *any* number of full-scale ranges.

Methods are available to allow any number of digits to be displayed, but this circuit is limited to displaying three and one-half digits to reduce cost and circuit complexity.

circuit description

Referring to **fig. 3**, the MC14433 uses two external resistors and two external capacitors to set its internal clock and ramp. Since analog ground, V_{AG} must be at least 2.8 volts above circuit ground, V_{EE}, we

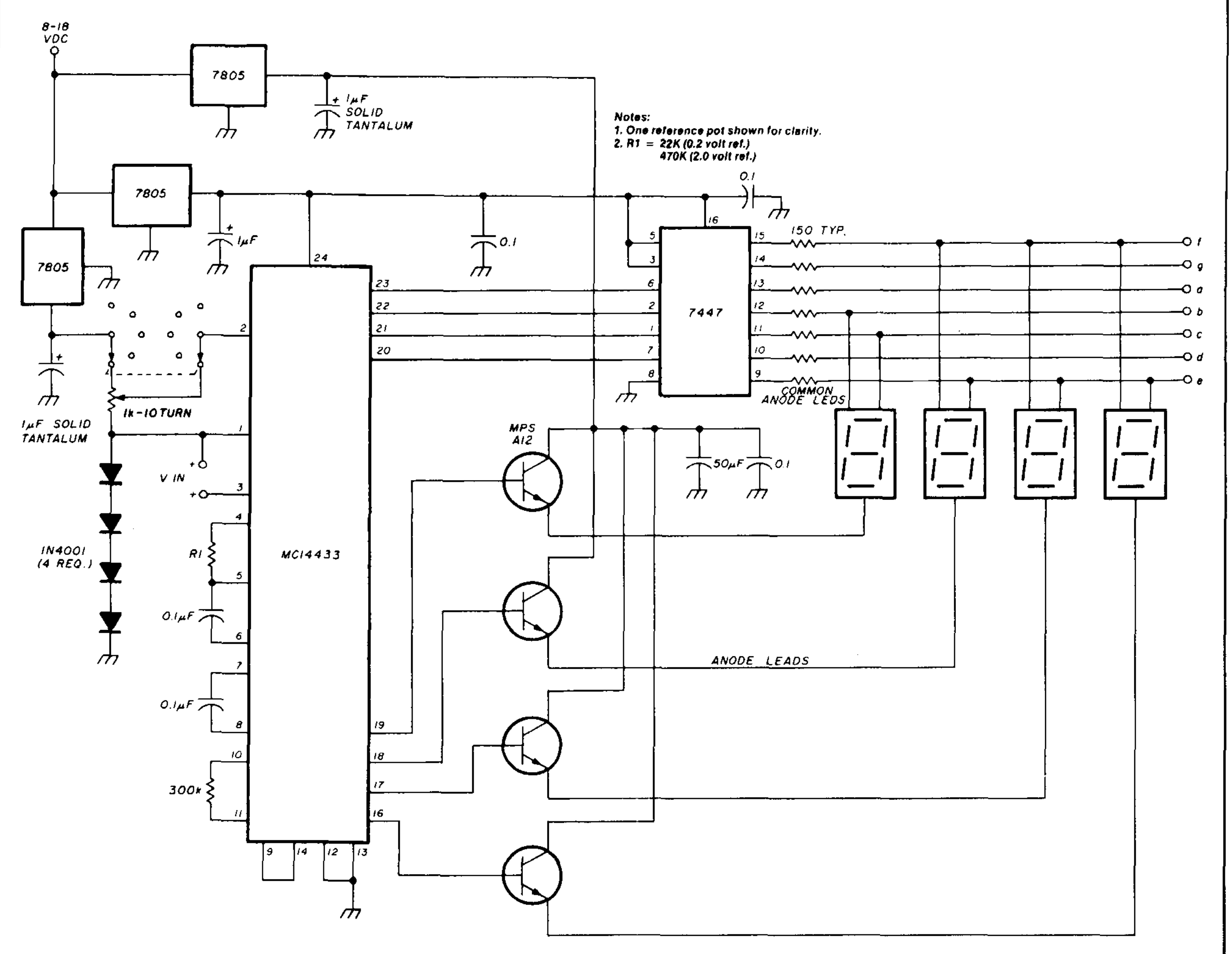

fig. 3. Schematic of the A/D converter using the Motorola MC14433 to drive a MC7447 seven-segment decoder-driver for the LED readouts.

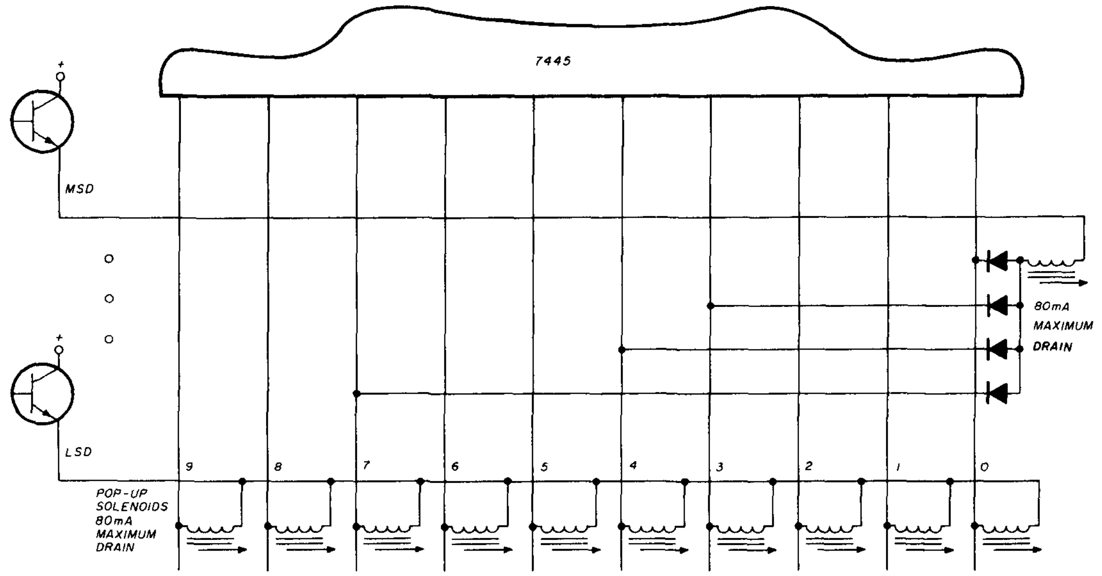

fig. 4. Alternative scheme using an MC7445 ten-line decoder IC for pop-up solenoids (tactile readouts).

use four silicon diodes between V_{AG} and ground (V_{EE} and V_{SS}), giving a voltage drop/separation of 2.8 volts (4 × 0.7 volt). We need either 2 or 0.2 volts for V_{REF}, depending on the range we must measure and the value we select for R1, so we use a 7805 regulator (5 volt) as an accurate voltage source and connect it through a 1k, ten-turn pot (available as surplus in many places) to V_{AG}. Since V_{AG} is 2.8 volts above ground, we have 2.2 volts available across the pot, (5.0V − 2.8V) or 0.22 volt per turn of the ten-turn pot. This voltage is adjusted during calibration for new full-scale readings.

The Q outputs of the MC14433 (pins 20-23) are fed to a 7447 seven-segment decoder-driver for LED readouts or to a 7445 ten-line decoder-driver for tactile readouts. (See **fig. 4.**) Note that only segments b and c are connected to the MSD (most significant digit) readout, **fig. 3.** The digit selects (pins 16-19) then enables each digit in turn through Motorola MPS A12 NPN transistor switches. You may use as large a digital display as desired, as the LEDs or solenoids are mounted external to the decoding and voltmeter circuitry for convenience.

Constuction of the circuit is straightforward, using normal PC techniques.* Care must be taken to shield the circuit and rf bypass the input and output leads, as the circuit can generate small amounts of rf hash due to its clock rate. I haven't found this to be objectionable except during very weak signal reception, so shielding is recommended.

The input voltage to the 7805s can be anywhere from 8 to 18 Vdc. I use a rectified and filtered filament line on my rig. You will experience less dissipation from the 7805s if you use a voltage closer to 8 than 18 volts.

The finished board can be mounted anywhere in the rig, as long as it is shielded from rf-sensitive areas. If at all possible, mount the new range switches onto the existing meter range switch. It's worth replacing the old switch with one that will switch everything.

calibration and installation

To replace an existing panel meter, determine the internal resistance and full-scale current of the meter.

1. *As accurately as possible* measure a resistor of the same nominal value as the internal resistance of the meter for example, 100 ohm, 1 k, and so on.
2. Disconnect the meter from the circuit and connect the leads to the new resistor. If the meter is not removed from the panel, single-lug terminal strips mounted to the meter binding posts make excellent mounts for the new resistor.
3. Now, *as accurately as possible,* measure the

*I will supply PC boards, kits, or completed units to those submitting proof of visual impairment, at my cost. The number and values of scales desired, and the replaced meter information (resistance and full-scale current), must be specified when inquiring about completed units and kits. Please send a self-addressed stamped envelope with your inquiry — **KB7JW** voltage produced across the new resistor at each setting of the meter range switch.

4. Use **eq. 1** to calculate the correct digital reading:

$$\left[\frac{V_{meas}}{(R_{repl})\ (I_{meter})}\right] Range_{fs} = digital\ reading \quad \textbf{(1)}$$

where V_{meas} = measured voltage (volts)

R_{repl} = meter replacement resistance (ohms)

I_{meter} = meter full-scale current (amperes)

$Range_{fs}$ = meter full-scale range reading

Adjust each reference potentiometer to obtain the correct reading at each range setting.

Thus, using a 100-ohm resistor to replace a 1-mA full-scale meter, a voltage reading of 0.07 volt should read:

$$\left[\frac{0.07}{(100)\ (0.001)}\right] 500 = 0350$$

for a full-scale range of 500 mA. Similarily, if your resistor is 100 ohms and replaces a 100-microampere meter, and your range is 500 mA full scale, 0.05 volt will read:

$$\left[\frac{0.05}{(1000)\ (0.001)}\right] 500 = 0250$$

(Leading zeros are not blanked.)

As you can see from **eq. 1**, the accuracy of your new readout will be no better than the accuracy with which you measure the new resistor and the voltages produced across it.

conclusion

Although intended primarily as a panel-meter replacement, this circuit may find almost unlimited uses in the Amateur station. If you design a PC board, do as I did and leave at least one blank spot on the function switch so you can add another function switch for expansion as you find more and more uses for the circuit. I hope you find this device as useful as I have.

bibliography

Motorola publications:

MC14433 Data sheet
AN471 Analog-to-Digital Conversion Techniques
AN557 Analog-to-Digital Cyclic Converter
AN559 Simple Ramp A/D Converter
AN702 High-Speed Digital-to-Analog and Analog-to-Digital Techniques
AN713 Binary D/A Converters Can Provide BCD-Coded Conversion
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the Kenscan 74

An inexpensive scanner and preset memory adapter for the Kenwood TR-7400A transceiver

Vhf transceiver design began with single-crystal control, then frequency synthesizers were added to cover an entire band. The newest radios have preset frequency memories and microprocessor scanning control. One purpose of Amateur Radio is to improve the state of the art. Another is building your own circuits. Some of us aren't designers, so I'd like to share this simple scanning module with those who want to improve the Kenwood TR-7400A transceiver. The module is inexpensive and has improved the operation at my station. I call it the Kenscan 74.

functional description

The Kenscan 74 module can store up to sixteen preset frequencies. Each can be selected individually in the step mode or scanned in the scan mode. Scanning will stop when the receiver squelch indicates an active frequency or when the transmitter is keyed. Scan rate is nominally four per second.

Scan is held for about seven seconds after a receiver frequency becomes clear. This pause allows you to pick up the microphone and press the transmit button. Scanning is then stopped and normal transmit/receive operations are possible. Scanning will remain at one frequency until reset by the Kenscan 74 module CLEAR button. Scanning will resume if the microphone button is untouched.

Step mode allows single-button manual scan. Preset channels are selected in groups of 2, 4, 8, or 16 frequencies. Either mode is especially convenient when operating mobile in heavy traffic or when changing a predetermined frequency.

Frequencies to be preset are first selected by the TR-7400A controls. The Kenscan 74 module memory is then loaded by using a single pushbutton. Memory contents are made available to the TR-7400A display and internal PLL frequency control. Once loaded, each frequency setting is held in separate memory locations and recalled with a single button.

Kenscan 74 module circuits are placed between the Kenwood panel controls and internal PLL circuit. Conversion doesn't alter existing Kenwood circuit boards; module wiring isn't critical. Parts count is minimal: only fourteen integrated circuits and two transistors. Layout isn't critical. The design offers many advantages at little expense.

memory and memory control

Fig. 1 is a block diagram of the Kenscan 74 mod-

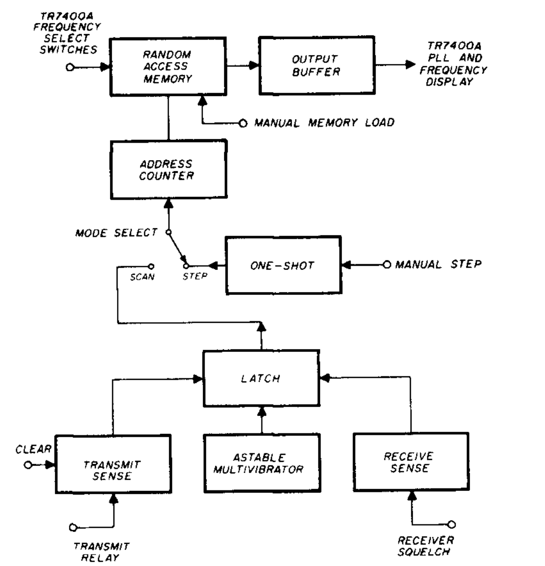


fig. 1. Kenscan 74 block diagram.

By Kenneth R. Fletcher, WB7QYB, 4080 SW 193rd, Aloha, Oregon 97007

ule. Three random access memory (RAM) ICs store the frequency switch settings. Each RAM is four bits by sixteen words; three RAMs are required to hold sixteen ten-bit words.*

The 74189 bipolar RAM memory IC (**fig. 2**) has inverted outputs. An inverting buffer is placed between RAM output and Kenwood control circuits to restore the original switch data. The RAMs are normally in read mode; a frequency word is always present at the output, depending on the memory address from the counter.

The sixteen-state address counter is set by either step-or scan-mode clocking. RAM addressing is thus one of sixteen stored frequency settings. Setting a particular address, then selecting a frequency and pressing the LOAD switch (**fig. 2**) will write the frequency setting into the RAM. Releasing the LOAD switch will set the Kenwood frequency control and display.

Fig. 2 is the Kenscan 74 module schematic. U5, U6, and U7 are the RAMs. U8 and U9 are the CMOS RAM output buffers. U3 is the RAM address counter. U3 is CMOS, so noninverting buffer U4 is used to provide sufficient drive for three TTL RAMs.

The number of address states produced by U3 is controlled by the CHANNEL SELECT switch, S2. In the 2 position, reset pin 9 of U3 is connected to the second stage. Counter output is 0000, then 0001. On reaching 0010, the second-stage high state causes a reset to all-zero; address count is limited to two. Positions 4 and 8 will reset on third- and fourth-stage high states, respectively, while the 16 positon allows a full count.

mode control

The counter is clocked in the step mode by one-shot U2. U2 is triggered by STEP switch S3, and the counter advances once. Scan-mode clocking is controlled by U1, a 555 timer used as an astable multivibrator with a rate of 4 Hz.

While STEP/SCAN switch S1 selects the counter clock, U1 output must pass through latch U10, which will stop the scan-mode clock when U12, Q1, and Q2 detect a receiver input from the TR-7400A squelch.†

Voltage-follower U12A input pin 3 is connected to Q13 collector on the receiver board inside the Kenwood. Q13 collector goes to about +1.8 volts when a signal is received; to +0.1 volts with no signal. Q1 and Q2 saturate when the buffered squelch voltage is high. C6 is normally charged to +5 volts through R9 and R10 but is discharged to ground by Q2 through R11. Voltage follower U12B couples the charge of C6 to the AND gate formed by U11. Latch U13 is normally high, so a low at AND gate pin 5 will cause its output pin 10 to be low. D input pin 5 of U10 is now low, so U10 latches and prevents scan pulses from reaching the address counter.

When the squelch voltage input drops to +0.1 volts, Q1 and Q2 are cut off and C6 begins charging through R10 and R9 to +5 volts. After seven seconds (adjustable by R10), C6 will charge high enough to bring the AND gate pins 5 and 10 high. U10 pin 5 is now high, scan pulses from U1 are passed through U10, and scanning resumes.

Latch flip-flop U13 is connected to the receive-transmit relay that supplies +12 volts when the transmitter is keyed. Zener CR1 clamps the transmit level to +5.1 volts at U13 clock pin 3. U13 D input pin 5 is tied high. U13 will toggle on the first transmit key pulse and remain set; Q pin 2 will drop low and inhibit AND gate U11A pin 6. U10 is then latched, and scanning stops.

U13 will lock up and scan until manually reset by the CLEAR switch. A squelch break or transmit-key pulse will stop scanning, but manual reset is required after transmitting to restore a scan.

The voltage regulator for the module is an LM340T-5 rated at 1.5 amperes. An LM7805CT will work equally well. Both have TO-220 cases.

assembly

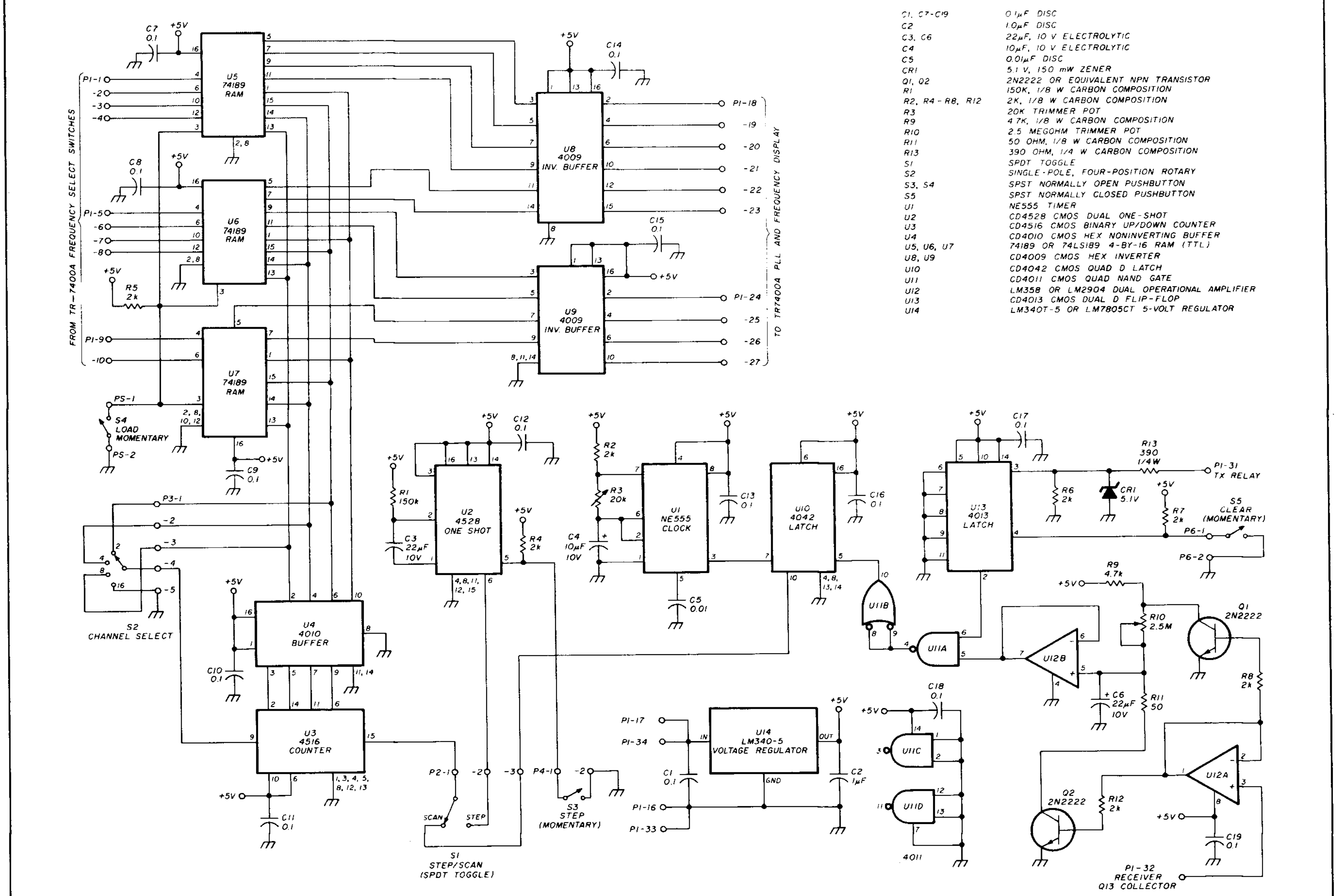
The circuit is fairly dense so I chose wire-wrap assembly. Layout isn't critical but the module must be larger than 12 square inches (80 square cm). Control switches may be mounted off the board. I prefer ribbon cable for both switch wiring and the 34 conductors necessary to interface with the Kenwood transceiver. Harmonica connectors, such as the 3M No. 3414, make neat, low-profile connections for compact places such as inside the transceiver.

Position the wire-wrap sockets and arrange connectors close to the IC locations. Wire wrapping should begin with all power and ground connections. Verify all wiring with an ohmmeter before inserting the ICs. It's much easier to correct wire-wrap mistakes early; the deeper the wiring, the harder it is to trace wires. Regulator U14 can be checked for operation at this point.

---

*Word is a term used to describe several bits in parallel for a specific data or value function.

†The CD4042 latch in U10 is not a conventional D flip-flop. Holding polarity-pin 6 high with clock-pin 5 high will pass any D input at pin 7 to output pin 10. D input will be latched into the output when pin 5 goes low; output is held (latched) until the clock returns to a high state.



C1, C7-C19	0.1μF DISC
C2	1.0μF DISC
C3, C6	22μF, 10 V ELECTROLYTIC
C4	10μF, 10 V ELECTROLYTIC
C5	0.01μF DISC
CR1	5.1 V, 150 mW ZENER
Q1, Q2	2N2222 OR EQUIVALENT NPN TRANSISTOR
R1	150K, 1/8 W CARBON COMPOSITION
R2, R4-R8, R12	2K, 1/8 W CARBON COMPOSITION
R3	20K TRIMMER POT
R9	4.7K, 1/8 W CARBON COMPOSITION
R10	2.5 MEGOHM TRIMMER POT
R11	50 OHM, 1/8 W CARBON COMPOSITION
R13	390 OHM, 1/4 W CARBON COMPOSITION
S1	SPDT TOGGLE
S2	SINGLE-POLE, FOUR-POSITION ROTARY
S3, S4	SPST NORMALLY OPEN PUSHBUTTON
S5	SPST NORMALLY CLOSED PUSHBUTTON
U1	NE555 TIMER
U2	CD4528 CMOS DUAL ONE-SHOT
U3	CD4516 CMOS BINARY UP/DOWN COUNTER
U4	CD4010 CMOS HEX NONINVERTING BUFFER
U5, U6, U7	74189 OR 74LS189 4-BY-16 RAM (TTL)
U8, U9	CD4009 CMOS HEX INVERTER
U10	CD4042 CMOS QUAD D LATCH
U11	CD4011 CMOS QUAD NAND GATE
U12	LM358 OR LM2904 DUAL OPERATIONAL AMPLIFIER
U13	CD4013 CMOS DUAL D FLIP-FLOP
U14	LM340T-5 OR LM7805CT 5-VOLT REGULATOR

fig. 2. Schematic diagram of the Kenscan 74 module. U5-U7 are the RAMs; U8, U9 are CMOS RAM output buffers; U3 is the RAM address counter.

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Super Design. *All Solid-State and Broadbanded*—from the pioneer, Ten-Tec. *Modular* plug-in circuit boards. *Functional Styling* with convenient controls, full shielding, easy-to-use size (5¾"h x 14¼"w x 14"d).

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table 1. Interface connections.

outputs from frequency-select switches (square pins)	scanner module connector P1	input to PLL and display
A3	1	
B3	2	
A2	3	
B2	4	
C2	5	
D2	6	
A1	7	
B1	8	
C1	9	
D1	10	
GND	16,33	
+ 12 V	17,34	
transmit relay	31	
Q13 collector	32	
	18	A3
	19	B3
	20	A2
	21	B2
	22	C2
	23	D2
	24	A1
	25	B1
	26	C1
	27	D1
	28,29,30	no connection
5V		5V
ØK		ØK
5K		5K
CT1		CT1
CT2		CT2

Kenwood TR-7400A modifications

This portion may be the most undesirable part of the project but it isn't difficult. To interface the scanner module, the mating interface connectors must be wired inside the Kenwood transceiver in the following manner.

Find the place on the TR-7400A schematic where the frequency programming information from the front-panel switches goes through a harmonica connector to two other boards. This connector is marked ØK, 5K, A2, C2, and so on. *Stop!* The connector is *not* marked correctly on the schematic. Call the pin marked ØK as 1. There's a space and a 5K marked on the schematic as the next pin. The space in the plug should really be between two pins marked A3 and 5V. The other plug is also misdrawn: the space should be moved from between CT1 and CT2 to between E and B1. Remarking the schematic will eliminate any confusion between the physical wiring and the schematic.

Next remove both top and bottom covers, then remove all knobs. The MHz lever pulls straight off, while all others have setscrews. Remove the four screws holding the front panel. Pull the display modules straight up, which will expose a board that must be loosened and moved aside (this board has two screws on each end). The frequency control connectors will now be visible. Identify each connector with the corrected schematic data.

The Kenwood TR-7400A connectors should now be replaced with the scanner module interface connector. The square pins are mated to the frequency-control switches, while the plugs are inputs to the

table 2. Troubleshooting chart.

problem	possible cause	solution
nothing works	no 5 V	Check U14
	no GND to scanner	check P1-16, P1-33
	no 12 V to regulator	check P1-17, P1-34
will not load frequency	pin 3 of U5, U6, U7 not pulled low	check S4 and wiring to P5
Display incorrect when LOAD is pushed	wiring error A1-B3	check P1-1 through P1-10
	broken wire	check P1-18 through P1-27
will not step	U2 wired incorrectly	check wiring
	S1 wired incorrectly	check wiring
	S3 wired incorrectly	check wiring
will not SCAN	S1 wired wrong	check wiring
	S2 wired wrong	check wiring
	no clock pulse	check U1-3, U10-10, U3-15
	U10 latched	push CLEAR
will not stop at busy channel	no stop signal	check squelch input from Q13
	scan rate too fast	set R3 for 4-Hz rate

display and PLL circuitry. With the aid of **table 1**, reconnect each wire to the scanner module interface connector. Note that five wires marked 5V, ØK, 5K, CT1, CT2 are jumpered *inside* the transceiver and do not go to the scanner module.

Once the interface connector is installed, reassembly is in the reverse order of assembly. Make certain that all knob indices are correct. Proceed with the scanner module and switch cabinet.

final test

Be sure that the regulator on the scanner module has an output of +5 volts before any ICs are installed: overvoltage will damage the RAMs. Eliminate any static electricity possibilities when handling the CMOS ICs or when handling the unconnected module.

With all assemblies and connectors ready, connect the scanner module to the TR-7400A. If any problems occur with the Kenscan 74, refer to the troubleshooting chart in **table 2**.

programming

Place the CHANNEL SELECT switch in the 2 position and the STEP/SCAN switch in STEP. Dial up the first frequency and press the LOAD switch. Push the STEP switch to change memory address. Dial the second frequency and push LOAD. Set the STEP/SCAN switch to SCAN and push CLEAR. The two preset frequencies should now be scanning. Step mode can select either frequency at a single touch of the STEP switch.

The other fourteen memories can be loaded and accessed in the same manner. The CHANNEL SELECT switch will allow you to preset two, four, eight, or sixteen combinations.

etched-circuit boards

At this wiring, etched circuit boards are being designed. For further information on these boards or kits, send a self-addressed, stamped envelope to the author. If you have any problems with the circuit, send an SASE with a detailed description of the problem and I'll try to help.

Author WB7QYB has a versatile adapter with a noncritical interface. While bells and whistles have been kept to a minimum, enough unused IC functions are available to experiment with your own additions. For example, the preset input pins of U3 could be used with a separate switch bank to add individual selection. You could scan eight frequencies with the CHANNEL SELECT switch set at 8 then use the other eight memory locations for fixed frequencies. A bit of study of the CMOS data sheets will show how this could be done. *It has lots of possibilities.* **Editor.**

ham radio

KENWOOD
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TS-830S

"Top-notch"... VBT, notch, IF shift, wide dynamic range

The TS-830S has every conceivable operating feature built-in for 160-10 meters (including the three new bands). It combines a high dynamic range with variable bandwidth tuning (VBT), IF shift, and an IF notch filter, as well as very sharp filters in the 455-kHz second IF. Its optional VFO-230 remote digital VFO provides five memories.

TS-830S FEATURES:
- LSB, USB, and CW on 160-10 meters, including the new 10, 18, and 24-MHz bands. Receives WWV.
- Wide receiver dynamic range. Junction FETs in the balanced mixer, MOSFET RF amplifier at low level, and dual resonator for each band.
- Variable bandwidth tuning (VBT). Varies IF filter passband width.
- Notch filter (high-Q active circuit in 455-kHz second IF.
- IF shift (passband tuning).
- Built-in digital display (six digits, fluorescent tubes), analog subdial, and display hold (DH) switch.
- Noise-blanker threshold level control.
- 6146B final with RF negative feedback. Runs 220 W PEP (SSB)/180 W DC (CW) input on all bands.
- Built-in RF speech processor.
- Narrow/wide filter selection on CW.
- SSB monitor circuit to check transmitted audio quality.
- RIT (receiver incremental tuning) and XIT (transmitter incremental tuning).

OPTIONAL ACCESSORIES:
- SP-230 external speaker with selectable audio filters.
- VFO-230 external digital VFO with 20-Hz steps, five memories, digital display.
- AT-230 antenna tuner/SWR and power meter/antenna switch; 160-10 meters, including three new bands.
- YG-455C (500-Hz) and YG-455CN (250-Hz) CW filters for 455-kHz IF.
- YK-88C (500-Hz) and YK-88CN (270-Hz) CW filters for 8.83-MHz IF.

(VFOs for TS-830S, TS-130 Series, and TS-120S are compatible with all three series of transceivers.)

SP-230 TS-830S VFO-230 AT-230

TS-130S/V

"Small wonder"... processor, N/W switch, IF shift, DFC option

The compact, all solid-state HF SSB/CW mobile or fixed station TS-130 Series transceiver covers 3.5 to 29.7 MHz, including the three new bands.

TS-130 SERIES FEATURES:
- 80-10 meters, including the new 10, 18, and 24-MHz bands. Receives WWV.
- TS-130S runs 200 W PEP/160 W DC input on 80-15 meters and 160 W PEP/140 W DC on 12 and 10 meters. TS-130V runs 25 W PEP/20 W DC input on all bands.
- Built-in speech processor.
- Narrow/wide filter selection on both CW (500 Hz or 270 Hz) and SSB (1.8 kHz) with optional filters.
- Automatic selection of sideband mode (LSB on 40 meters and below, and USB on 30 meters and above). SSB REVERSE switch provided.
- Built-in digital display.
- Built-in RF attenuator.
- IF shift (passband tuning).
- Effective noise blanker.

OPTIONAL ACCESSORIES:
- PS-30 base-station power supply.
- YK-88C (500 Hz) and YK-88CN (270 Hz) CW filters.
- YK-88SN (1.8 kHz) narrow SSB filter.
- AT-130 compact antenna tuner (80-10 meters, including three new bands).
- SP-120 external speaker.
- VFO-120 remote VFO.
- MB-100 mobile mounting bracket.
- PS-20 base-station power supply for TS-130V.

Optional DFC-230 Digital Frequency Controller
Frequency control in 20-Hz steps with UP/DOWN microphone (supplied with DFC-230). Four memories and digital display. (Also operates with TS-120 and TS-830S.)

PS-30 SP-120 TS-130S VFO-120

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R-1000

"Hear there and everywhere"... easy tuning, digital display

The R-1000 is an amazingly easy-to-operate, high-performance, communications receiver, covering 200 kHz to 30 MHz in 30 bands. This PLL synthesized receiver features a digital frequency display and analog dial, plus a quartz digital clock and timer.

R-1000 FEATURES:

- Covers 200 kHz to 30 MHz continuously.
- 30 bands, each 1 MHz wide.
- Five-digit frequency display with 1-kHz resolution and analog dial with precise gear dial mechanism.
- Built-in 12-hour quartz digital clock with timer to turn on radio for scheduled listening or control a recorder through remote terminal.
- Step attenuator to prevent overload.
- Three IF filters for optimum AM, SSB, CW. 12-kHz and 6-kHz (adaptable to 6-kHz and 2.7-kHz) for AM wide and narrow, and 2.7-kHz filter for high-quality SSB (USB and LSB) and CW reception.
- Effective noise blanker.
- Terminal for external tape recorder.
- Tone control.
- Built-in 4-inch speaker.
- Dimmer switch to control intensity of S-meter and other panel lights and digital display.
- Wire antenna terminals for 200 kHz to 2 MHz and 2 MHz to 30 MHz. Coax terminal for 2 MHz to 30 MHz.
- Voltage selector for 100, 120, 220, and 240 VAC. Also adaptable to operate on 13.8 VDC with optional DCK-1 kit.

OPTIONAL ACCESSORIES:

- SP-100 matching external speaker.
- HS-5 and HS-4 headphones.
- DCK-1 modification kit for 12-VDC operation.

SP-100 R-1000 HS-5

HC-10

Digital world clock with two 24-hour displays, quartz time base

The HC-10 digital world clock with dual 24-hour display shows local time and the time in 10 preprogrammed plus two programmable time zones.

HC-10 FEATURES:

- Two 24-hour displays with quartz time base. Right display shows local (or UTC) hour, minute, second, day. Left display shows month, date, world time in various cities, memory time (QSO starting time), and time difference (in hours from UTC).
- Preprogrammed time in 10 cities around the world, plus two programmable time zones.
- "TOMORROW" and "YESTERDAY" indicators.
- Memorization of present time. Can be recalled later, for logging purposes.
- High accuracy (±10 seconds/month).

DM-81

Dip meter performs many RF measurements

The DM-81 dip meter is highly accurate and features, in addition to the traditional inductive-coupling technique, capacitive coupling for measuring metal-enclosed coils and toroidal coils.

DM-81 FEATURES:

- Measuring range of 700 kHz-250 MHz in seven bands.
- Built-in storage compartment for all seven coils, capacitive probe, earphone, and ground clip lead.
- All solid-state and built-in battery.
- HC-25U and FT-243 sockets for checking crystals and marker-generator function.
- Amplitude modulation.
- FET for good sensitivity.
- Absorption frequency meter function.
- Earphone for monitoring transmitted signals.
- Capacitance probe for measuring resonant frequencies without removing coil shields, and also for measuring resonant frequencies of toroidal coils.

quick and simple antenna match

A candidate for monoband mobile work

One component and a length of transmission line will make a quick and simple antenna match. Sound too good to be true? The method borrows an old microwave technique and is limited to narrow bandwidths. It's useful for restricted monoband operation, especially for mobile work.

Familiarity with the Smith chart is required.[1] You must have a good measurement of the antenna impedance at the transmitter end.[2,3] The component value and extra line length can be found by graphic or numerical methods. This matching scheme is best explained by a quick review of certain parts of the Smith chart.

intercepting the R_O circle

A key element of the Smith chart is the resistive circle lying on the bisecting line and passing through chart center and infinite resistance. This circle is marked as R_O equal to Z_0 for a specific impedance or unity on a "normalized" chart.*

This simple matching method requires adding a line to move the measured impedance in a clockwise direction until the rotated impedance point intersects the $R_O = Z_0$ circle at either of two places. A clockwise rotation is marked "toward generator." Once at the R_O circle intersection, the new reactance may be read directly.

*Normalized charts have all impedances divided by characteristic impedance of a line. These charts can be used with any impedance.

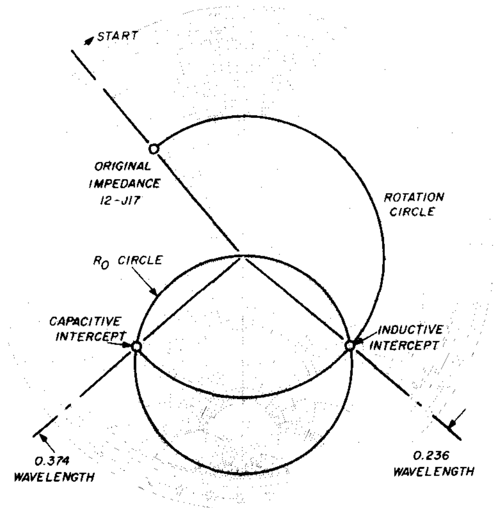

fig. 1. Graphic method of finding a match.

Matching requires adding an *opposite* reactance in series with the new end impedance. It is simply series resonating, so that only the resistive part remains. You have reached the perfect match!

a graphical method

Fig. 1 shows the matching scheme. Suppose you measure an impedance of 12-j17 ohms at the existing-line transmitter end. Using a drafting compass

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pivoting on chart center, draw a circle through this impedance point until the $R_O = Z_0$ circle is intercepted. Next, draw radial lines from chart center through the original impedance and each R_O circle intercept. Radial lines should go through the circular wavelength scales.

Note the reactance component at the R_O circle intercept. In this example it's about 84.5 ohms. Calculate a capacitor and inductor value for this reactance

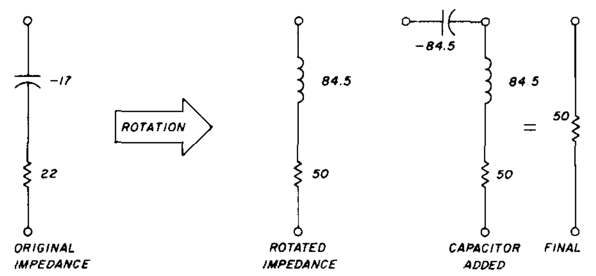


fig. 2. Schematic representation of the Smith chart operations.

at the original impedance-measurement frequency; choice of capacitor or inductor depends on the added line length.

Total the fractional wavelength, in a clockwise direction, from original impedance radial to each of the R_O circle intercept radials. Inductive-reactance side intercept should total 0.236 wavelength; series capacitance is used at this intercept. Capacitive reactance intercept occurs at 0.374 wavelength, and a series inductor would be required here.

The sequence of events is shown in **fig. 2**. Rotating the original $12\text{-}j17$ ohm impedance by 0.236 wavelength (adding a line) will make it $50 + j84.5$ ohms. Adding a series capacitor of $-j84.5$ ohms results in a final impedance of $50 + j0$ ohms.

The antenna hasn't changed impedance, but the load presented to the transmitter is now purely resistive; a better power transfer occurs. The original impedance has a 4.67:1 VSWR and may be out of the range of a tuner or may upset an ALC circuit.

line lengths

This quantity is easily determined by

$$\begin{aligned} &Length = k \times wavelength \\ &\times velocity\ of\ propagation/frequency\ (MHz) \end{aligned} \quad \textbf{(1)}$$

where: $K = 984$ for length in feet,
300 for length in meters

Propagation velocity depends on the added transmission line and would be 0.659 for common polyethylene dielectric coaxial cable. The example in feet at 3.9 MHz with RG-58A cable would be:

$$\begin{aligned} &984 \times 0.236 \times 0.659/3.9 \\ &= 161.86/3.9 = 39.24\,feet \end{aligned}$$

This length of line might shock 80-meter fans at today's prices. An advantage is that the added line is *indoors*; some previously used outdoor line may still be good.

numerical methods

Previously published formulas can be rearranged for those lacking Smith charts.[4] Assigning R_O as the added line characteristic impedance and $R_L + jX_L$ as the original transmitter-end measurement, series reactance is:

$$X_s = \sqrt{[R_O(R_O^2 + X_L^2)]/R_L + R_O(R_L - 2R_O)} \quad \textbf{(2)}$$

This reactance is both capacitive and inductive. It represents the inductive-component-side intercept and the added capacitor reactance. Calculation from the example would be:

$$\begin{aligned} X &= \sqrt{[50(2500 + 289)]/12 + 50(12 - 100)} \\ &= \sqrt{7220.83} = 84.98\,ohms \end{aligned}$$

Line-length formula results in a phase angle, β, that must be divided by 360 degrees to obtain fractional wavelength in **eq. 1**.

$$\beta = arctan\ \frac{R_O(R_O - R_O)}{R_O X_L + R_L X} \quad \textbf{(3)}$$

The X term with no subscript is the value from **eq. 2**. Example calculations are:

$$\begin{aligned} \beta &= tan^{-1}\left[\frac{50(50-12)}{50(-17) + 12(84.98)}\right] \\ &= tan^{-1}\left[\frac{1900}{169.76}\right] = 84.89\ degrees \end{aligned}$$

Dividing 84.98 by 360 gives the wavelength of 0.236, and **eq. 1** is used for physical length.

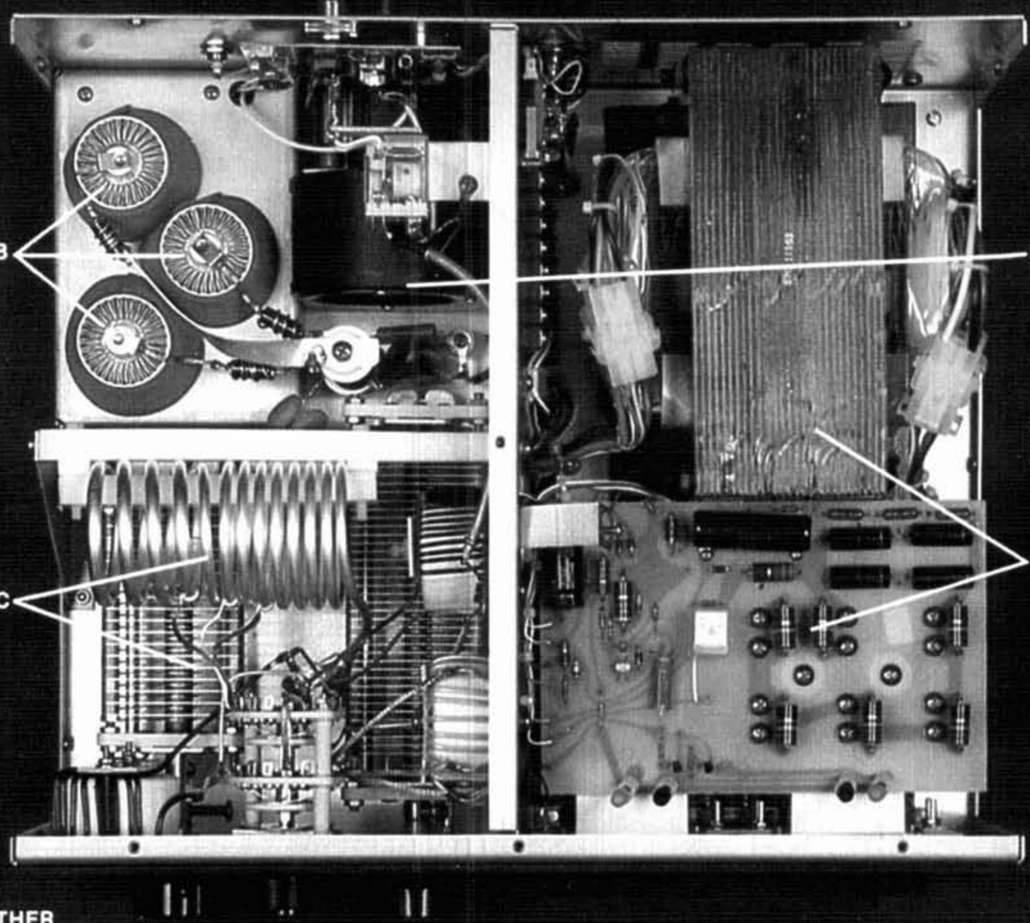
A rotation greater than 0.25 wavelength gives a negative angle for β. Simply add 180 degrees for the correct angle. The reason for a negative result is that an arctangent result can only be within ± 90 degrees. The negative angle represents a counterclockwise rotation. Since the Smith chart is only 180 degrees in circumference, adding 180 degrees to a negative result will give a correct clockwise rotation angle.

Suppose another measurement was $22.5\text{-}j50$ ohms. Numeric equations would yield X of 85.07 ohms and β of -66.92 degrees. Adding 180 degrees to β would give the correct 113.08 degrees or 0.314 wavelength rotation.

case history of a quick match

A newly moved ham wanted to get on 20 meters in

It couldn't be anything but...



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ETO

table 1. Calculated data of the 20-meter dipole matching example. Columns marked "original" and "measured" are from RX noise bridge readings, others are ideal conditions. Capacitive reactance for 200 pF. Target frequency for optimum match was 14.15 MHz.

	original		new Z at 12-foot		ideal new $Z + X_C$		measured	
F (MHz)	impedance	VSWR	extension	X_C	impedance	VSWR	impedance	VSWR
14.00	39.3 − j50.8	3.04	25.7 + j34.2	−56.8	25.7 − j22.6	2.45	26.6 − j22.1	2.35
14.05	32.6 − j45.1	3.11	29.0 + j40.5	−56.6	29.0 − j16.1	1.98	30.3 − j15.3	1.88
14.10	27.8 − j41.3	3.27	31.7 + j46.6	−56.4	31.7 − j9.8	1.67	32.7 − j8.6	1.60
14.15	25.1 − j36.1	3.22	37.7 + j52.3	−56.2	37.7 − j3.9	1.35	39.7 − j3.5	1.28
14.20	23.3 − j31.6	3.15	45.2 + j57.4	−56.0	45.2 + j1.4	1.11	47.0 + j1.0	1.07
14.25	23.2 − j24.9	2.80	60.7 + j58.2	−55.8	60.7 + j2.4	1.22	62.0 + j1.0	1.24
14.30	24.5 − j17.5	2.36	80.7 + j47.0	−55.6	80.7 − j8.6	1.64	78.8 − j9.6	1.52
14.35	25.8 − j10.5	2.05	93.7 + j25.0	−55.5	93.7 − j30.5	2.14	90.9 − j28.7	2.07
14.40	27.7 − j2.8	1.81	90.5 − j2.7	−55.3	90.5 − j58.0	2.74	90.1 − j56.1	2.68

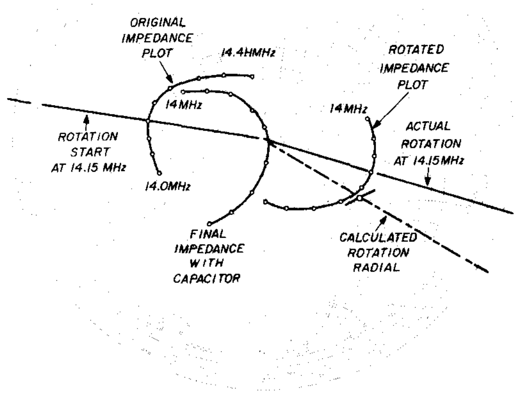


fig. 3. Impedance plots of 20-meter dipole matching.

a hurry with a temporary dipole. The antenna tuner had been mislaid and the transmitter did not seem to load correctly. RX noise bridge readings and calculations showed a reasonable VSWR condition given in the left columns of **table 1**. A better match was desired at 14.15 MHz.

Numerical calculations resulted in 61.90 ohms for X, 101.42 degrees or 0.282 wavelength added line. This translated to 181.7 pF series capacitance and 12.92 feet (3.94 meters) of line. Smith chart plots are given in **fig. 3** with the dash-dot line marking the 50-ohm intersection at 14.15 MHz.

Two six-foot (1.83-meter) lengths of RG-58C were available and a junkbox search came up with two 100-pF transmitting mica capacitors along with a small box having appropriate cable connectors (to house the capacitors). HP-67 calculator work indicated this combination should do well.[5] The ideal rotation plus series capacitance is plotted in **fig. 3**; ideal and actual results are in **table 1**. Total time for measurement and matching took half of a Sunday afternoon.

This type of match is no substitute for a good antenna tuner but shows what can be done quickly and with fixed components. The rotated impedance in **fig. 3** is slightly more than 90 degrees away from the original. Plot points of the ideal case are on each resistive-part locus, differing only by the series reactance.

applications

Monoband mobile work seems the best application for this simple matching scheme. Restricting the frequency to phone bands permits matching to within 2:1 VSWR. A fixed series reactance should survive vibration better than a variable tuner; it should be in a sturdy metallic box since both ends are above ground. Added cable can be tucked away in the trunk or behind seat backs. Do *not* put any cable under adjustable seats; slashed cable is the invariable result.

Although a simple technique, the method requires accurate measurement and calculation.[6,7]

references

1. James Fisk, W1DTY, "How to Use the Smith Chart," *ham radio*, November, 1970, page 16 (reprinted March, 1978, page 92).
2. Robert A. Hubbs, W6BXI, and A. Frank Doting, W6NKU, "Improvements to the RX Noise Bridge," *ham radio*, February, 1977, page 10.
3. Paul White, W6BKX, 'BKX Bridge," *73*, November, 1969, page 32.
4. Leonard H. Anderson, "Antenna Bridge Calculations," *ham radio*, May, 1978, page 34.
5. *Program 02603D*, HP Users' Library, 1000 N. E. Circle Blvd., Corvallis, Oregon, 97330 (a similar HP-97 program is *Program 02258D*; each is $6.00 and includes a recorded card).
6. T.J. Anderson, WD4GRI, "Noise Bridge Calculations," *ham radio*, May, 1978, page 45 (TI-58, 59 programs).
7. C.R. MacCluer, W8MQW, "Numerical Smith Chart," *ham radio*, March, 1978, page 104 (HP-25 programs).

ham radio

transmission-line circuit design

Using distributed resonant circuits for uhf/vhf transmission lines

In part 1 of this article (*ham radio*, November, 1980) I addressed the governing expressions for calculating resonant transmission-line parameters. Included were data on design relationships such as efficiency, coupling, and resonating capacitance. Programs for the Hewlett-Packard HP-67/97 calculator were detailed using the HP-97 printer capability.

This part of the article examines the geometry of the first four configurations of twelve different transmission lines in common use at 50 MHz and above (50 MHz is not a lower limit, however). The parameters for resonant-circuit design are described for:

1. Coaxial lines.

2. Parallel plates.

3. Parallel wires in air.

4. Single wire over a plane.

Subsequent issues of *ham radio* will present similar data on the remaining transmission-line configurations.

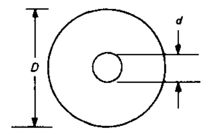
line configurations, graphs, and calculator programs

The critical parameters of the twelve transmission-line configurations previously identified in **fig. 1** (*ham radio*, November, 1980) are described. Graphic solution of characteristic impedance, Z_0, as a function of the physical geometry of each line is given where practical. Also HP-67/97 calculator programs are provided for each line configuration, permitting rapid solution for the selected variables or variables. For most cases the programs are written to permit the reverse solution to be calculated, thereby checking both accuracy of data entry and program.

The programs are not optimized to minimize steps. In all cases the 224 steps available in the HP-67/97 are sufficient to calculate the desired results. In some cases, more than one line configuration can be programmed on a magnetic card. The final choice rests with each user as to exactly what line configurations are grouped together to satisfy design requirements.

Following each program is a table that shows which registers are used in the program and what they contain. Tables indicating how the program is controlled are included and, where useful, a sample problem is given. References give the source data from which the transmission-line parameters were calculated.

coaxial line



This is the basic transmission-line configuration used in resonant circuits. The governing equation for transmission-line parameters (reference 4) is:

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \frac{D}{d} \qquad \textbf{(19)}$$

where Z_0 = transmission line impedance (ohms)
ϵ_r = dielectric constant (air = 1)*
D = inside major diameter
d = outside minor diameter

Fig. 8 shows the graphical relationship of D/d versus Z_0 for Z_0 between zero and 300 ohms. These values were calculated for the HP-67/97 program shown in **table 8**. The storage register contents are in **table 9**; program control is in **table 10**. The program assumes that the dielectic constant, ϵ_r, is 1.0; that is, ϵ_r is equal to air if no value is entered.

By H.M. Meyer, Jr., W6GGV, 29330 Whitley Collins Drive, Rancho Palos Verdes, California 90274

Table 8. HP-67/97 program for calculating D/d and Z_0 for coaxial lines.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	026	x	– 35
002	$\sqrt{X}$	54	027	1	01
003	STO0	35 00	028	3	03
004	RTN	24	029	8	08
005	*LBLB	21 12	030	÷	– 24
006	÷	– 24	031	10^x	16 33
007	*LBL1	21 01	032	STO1	35 01
008	STO1	35 01	033	RTN	24
009	LOG	16 32	034	*LBLD	21 14
010	1	01	035	STO1	35 01
011	3	03	036	GTO1	22 01
012	8	08	037	*LBL2	21 03
013	x	– 35	038	R↓	– 31
014	PCL0	36 00	039	1	01
015	X=0?	16-43	040	STO0	35 00
016	GSB2	23 02	041	÷	– 24
017	÷	– 24	042	STO2	35 02
018	STO2	35 02	043	R/S	51
019	RTN	24	044	*LBL3	21 03
020	*LBLC	21 13	045	1	01
021	STO2	35 02	046	STO0	35 00
022	*LBL4	21 04	047	R↓	– 31
023	RCL0	36 00	048	GTO4	22 04
024	X=0?	16-43	049	R/S	51
025	GSB3	23 03			

table 9. Register contents for HP-67/97 program for calculating D/d and Z_0 for coaxial lines.

STO 0	$\sqrt{\epsilon_r}$
STO 1	D/d
STO 2	Z_0

table 10. HP-67/97 program control for calculating D/d and Z_0 for coaxial lines.

enter	ϵ_r	press A
calculates	Z_0	
enter	D	press ENTER (↑)
enter	d	press B
calculates	D/d	
enter	Z_0	press C
calculates	Z_0	
enter	D/d	press D

Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.

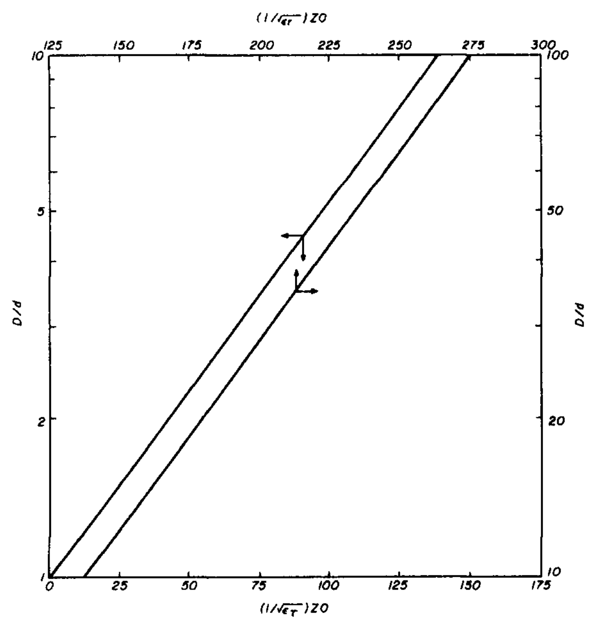
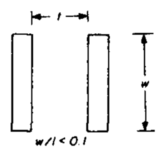


fig. 8. Characteristic impedance of coaxial lines.

parallel plates



This baseline configuration is often used with vacuum tubes. It is also used as a baseline geometry in calculating striplines between and over a ground plane. It is important to note that fringe and distributed capacitance effects, which can introduce substantial errors, are not considered. In addition, the thickness of the lines is assumed to be zero.

The governing equation (reference 4) as shown for this configuration is:

$$Z_0 = \frac{377}{\sqrt{\epsilon_r}} \log_{10} \frac{w}{t} \qquad \textbf{(20)}$$

where Z_0 = line impedance (ohms)
w = line width
t = minimum distance between lines
$w/t > 1.0$

Table 11 is the program for the HP-67/97, **table 12** shows the storage registers used, and **table 13** shows program control. **Fig. 9** shows the relationship of w/t with respect to Z_0. The program assumes $\epsilon_r = 1$ if no value is entered.

Table 11. HP-67/97 program for calculating Z_0 and w/t for parallel plates.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	026	÷	– 24
002	STO0	35 00	027	STO3	35 03
003	√X	54	028	*LBL4	21 04
004	STO1	35 01	029	LOG	16 32
005	RTN	24	030	3	03
006	*LBLB	21 12	031	7	07
007	STO2	35 02	032	7	07
008	*LBL1	21 01	033	x	– 35
009	RCL1	36 01	034	STO4	35 04
010	X = 0?	16-43	035	*LBL2	21 02
011	GSB9	23 09	036	RCL1	36 01
012	x	– 35	037	X = 0?	16-43
013	3	03	038	GSB8	23 08
014	7	07	039	÷	– 24
015	7	07	040	STO2	35 02
016	÷	– 24	041	R/S	51
017	10^x	16 33	042	*LBL8	21 08
018	STO3	35 03	043	1	01
019	R/S	51	044	STO1	35 01
020	*LBL9	21 09	045	RCL4	36 04
021	1	01	046	GTO2	22 02
022	STO1	35 01	047	*LBLD	21 14
023	RCL2	36 02	048	STO3	35 03
024	GTO1	22 01	049	STO4	22 04
025	*LBLC	21 13	050	R/S	51

table 12. Register contents for HP-67/97 program for calculating Z_0 and w/t for parallel plates.

STO 0	ϵ_r
STO 1	$\sqrt{\epsilon_r}$
STO 2	Z_0
STO 3	w/t
STO 4	INTERIM

table 13. HP-67/97 program control for calculating Z_0 and w/t for parallel plates.

enter	ϵ_r	press A
calculates	w/t	
enter	Z_0	press B
calculates	Z_0	
enter	W/t	press C
calculates	Z_0	
enter	w/t	press D

Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.

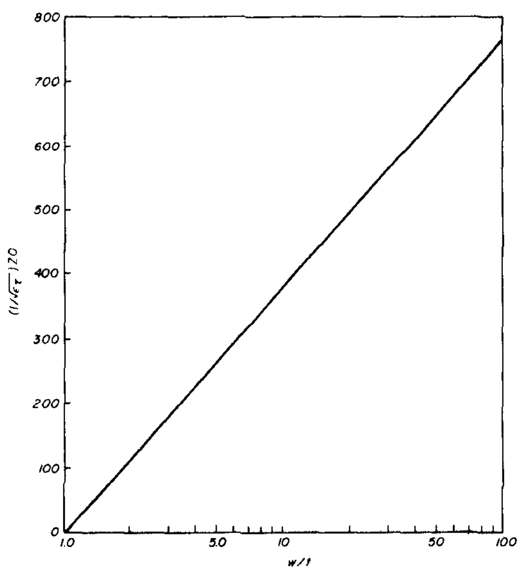
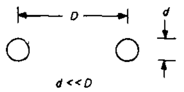


fig. 9. Parallel plate w/t versus Z_0.

parallel wires in air



This is the classic balanced transmission line. It is included here for completeness. The exact equation (reference 4) is:

$$Z_0 = 120 \cosh^{-1} \frac{D}{d} \quad \textbf{(21)}$$

where D = center-to-center spacing between lines
d = outside diameter of each line
Z_0 = line impedance (ohms)

Fig. 10 is a plot of line impedance versus ratio D/d. **Table 14** shows the HP-67/97 program for calculating either the Z_0 or D/d values depending on the available data. **Table 15** shows the storage register content used for the program; **table 16** shows program control.

To calculate $\cosh^{-1} D/d$, the following was used:

$$\cosh^{-1} \frac{D}{d} = \ln \left\{ \frac{D}{d} + \left[\left(\frac{D}{d} \right)^2 - 1 \right]^{1/2} \right\} \quad \textbf{(22)}$$

In calculating the converse, assume $\cosh^{-1} = a$, then:

$$\frac{D}{d} = \frac{(e^a)^2 + 1}{2e^a} \quad \textbf{(23)}$$

Table 14. HP-67/97 program for calculating Z_0 and D/d for parallel lines in air.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	020	*LBL1	21 01
002	STO0	35 00	021	X^2	53
003	1	01	022	1	01
004	2	02	023	−	−45
005	0	00	024	$\sqrt{X}$	54
006	÷	−24	025	RCL3	36 03
007	e^x	33	026	+	−55
008	STO2	35 02	027	LN	32
009	X^2	53	028	1	01
010	1	01	029	2	02
011	+	−55	030	0	00
012	RCL2	36 02	031	x	−35
013	2	02	032	STO0	35 00
014	x	−35	033	R/S	51
015	÷	−24	034	*LBLC	21 13
016	STO3	35 03	035	÷	−24
017	R/S	51	036	STO3	35 03
018	*LBLB	21 12	037	GTO1	22 01
019	STO3	35 03	038	R/S	51

table 15. Register contents for HP-67/97 program for calculating Z_0 and D/d for parallel lines in air.

STO 0	Z_0
STO 1	INTERIM
STO 3	D/d

table 16. HP-67/97 program control for calculating Z_0 and D/d for parallel lines in air.

calculates	D/d	
enter	Z_0	press A
calculates	Z_0	
enter	D/d	press B
calculates	Z_0	
enter	D	
enter	d	press D

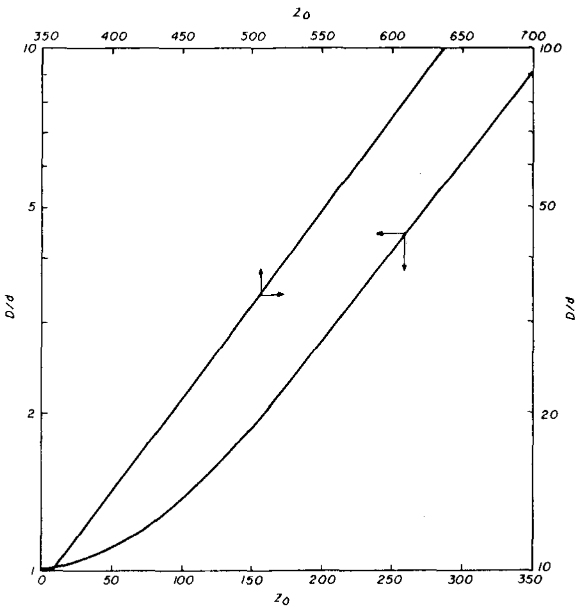
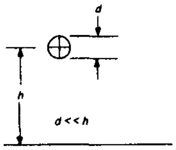


fig. 10. Characteristic impedance of parallel lines.

single wire over a plane



This configuration is often used for resonant circuits to 2 GHz and as the main element of filters in local-oscillator chains. The governing equation (reference 4) is:

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \frac{4h}{d} \qquad d << h \tag{24}$$

where Z_0 = line impedance (ohms)

ϵ_r = dielectric constant

h = centerline height above the plane of the wire

d = wire diameter

Fig. 11 depicts Z_0 versus h/d for a nominal range of values. **Table 17** is the HP-67/97 program for calculating Z_0 or h/d or, given h and d, calculating Z_0. If ϵ_r is not given the program assumes $\epsilon_r = 1$; that is, air. **Table 18** shows the storage registers used for the program; **table 19** indicates program control.

Table 17. HP-67/97 program for calculating Z_0 and h/d for a wire over a plane.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	028	3	03
002	STO0	35 00	029	8	08
003	$\sqrt{X}$	54	030	÷	– 24
004	STO1	35 01	031	STO3	35 03
005	RTN	24	032	*LBL3	21 03
006	*LBLB	21 12	033	RCL1	36 01
007	÷	– 24	034	X = 0?	16-43
008	*LBL1	21 01	035	GSB8	23 08
009	STO2	35 02	036	x	– 35
010	4	04	037	10^x	16 33
011	x	– 35	038	4	04
012	LOG	16 32	039	÷	– 24
013	1	01	040	STO2	35 02
014	3	03	041	R/S	51
015	8	08	042	*LBLD	21 14
016	x	– 35	043	GTO1	22 01
017	STO3	35 03	044	*LBL9	21 09
018	*LBL2	21 02	045	1	01
019	RCL1	36 01	046	STO1	35 01
020	X = 0?	16-43	047	RCL3	36 03
021	GSB9	23 09	048	GTO2	22 02
022	÷	– 24	049	*LBL8	21 08
023	STO4	35 04	050	1	01
024	R/S	51	051	STO1	35 01
025	*LBLC	21 13	052	RCL3	36 03
026	STO4	35 04	053	GTO3	22 03
027	1	01	054	R/S	51

table 18. Register contents for HP-67/97 program for calculating Z_0 and h/d for a wire over a plane.

STO 0	ϵ_r
STO 1	$\sqrt{\epsilon_r}$
STO 2	h/d
STO 3	INTERIM
STO 4	Z_0

table 19. HP-67/97 program control for calculating Z_0 and h/d for a wire over a plane.

enter	ϵ_r	press A
calculates	Z_0	
enter	h	press ENTER
enter	d	press B
calculates	h/d	
enter	Z_0	press C
calculates	Z_0	
enter	h/d	press D

Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.

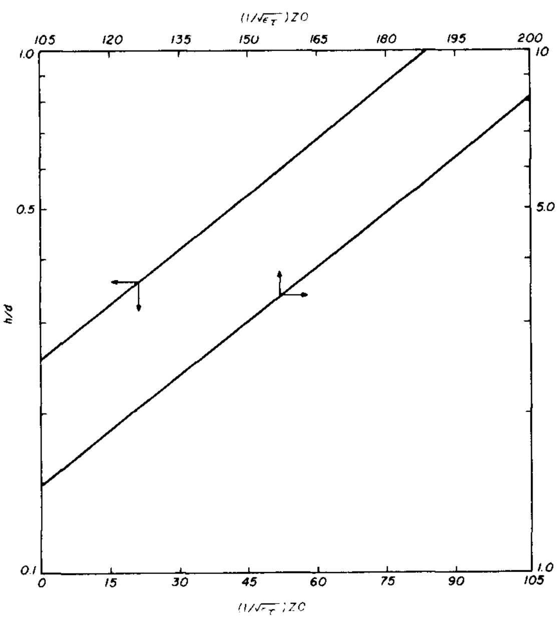


fig. 11. Z_0 versus h/d for a wire over a plane.

In the next series, to be published in a subsequent issue, I shall discuss the geometry and resonant-circuit design of the following line configurations:

1. Circular wire between planes.
2. Parallel wires over a plane.
3. Circular wire in an open trough.
4. Parallel wires between planes/rectangular box.

reference

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Range:	20 Hz to 600 MHz
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Range:	20 Hz to 525 MHz
Sensitivity:	Less than 50 MV to 150 MHz Less than 150 MV to 500 MHz
Resolution:	1.0 Hz (5 MHz range) 10.0 Hz (50 MHz range) 100.0 Hz (500 MHz range)
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Time base:	1.0 ppm TCXO 20-40°C
Power:	12 VAC @ 250 ma

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CT-70 Kit, 90 day parts warranty	84.95
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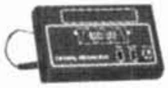
BP-1 Nicad pack + AC adapter/charger	12.95



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PRICES:

MINI-100 wired, 1 year warranty	$79.95
MINI-100 Kit, 90 day part warranty	59.95
AC-Z Ac adapter for MINI-100	3.95
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Here's a handy, general purpose counter that provides most counter functions at an unbelievable price. The MINI-100 doesn't have the full frequency range or input impedance qualities found in higher price units, but for basic RF signal measurements, it can't be beat! Accurate measurements can be made from 1 MHz all the way up to 500 MHz with excellent sensitivity throughout the range, and the two gate times let you select the resolution desired. Add the nicad pack option and the MINI-100 makes an ideal addition to your tool box for "in-the-field" frequency checks and repairs.

SPECIFICATIONS:

Range:	1 MHz to 500 MHz
Sensitivity:	Less than 25 MV
Resolution:	100 Hz (slow gate) 1.0 KHz (fast gate)
Display:	7 digits, 0.4" LED
Time base:	2.0 ppm 20-40°C
Power:	5 VDC @ 200 ma

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Range:	20 Hz to 600 MHz
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Display:	8 digits 0.4" LED
Time base:	2.0 ppm 20-40°C
Power:	110 VAC or 12 VDC

The CT-50 is a versatile lab bench counter that will measure up to 600 MHz with 8 digit precision. And, one of its best features is the Receive Frequency Adapter, which turns the CT-50 into a digital readout for any receiver. The adapter is easily programmed for any receiver and a simple connection to the receiver's VFO is all that is required for use. Adding the receiver adapter in no way limits the operation of the CT-50, the adapter can be conveniently switched on or off. The CT-50, a counter that can work double-duty!

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CT-50 Kit, 90 day parts warranty	119.95
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RA-1 wired and pre-programmed (send copy of receiver schematic)	29.95

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DM-700 Kit, 90 day parts warranty	79.95
AC-1, AC adaptor	3.95
BP-3, Nicad pack +AC adapter/charger	19.95
MP-1, Probe kit	2.95

The DM-700 offers professional quality performance at a hobbyist price. Features include; 26 different ranges and 5 functions, all arranged in a convenient, easy to use format. Measurements are displayed on a large 3½ digit, ½ inch LED readout with automatic decimal placement, automatic polarity, overrange indication and overload protection up to 1250 volts on all ranges, making it virtually goof-proof! The DM-700 looks great, a handsome, jet black, rugged ABS case with convenient retractable tilt bail makes it an ideal addition to any shop.

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Resistance:	0.1 ohms to 20 Megohms, 6 ranges
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Accuracy:	10.1% basic DC volts
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ham radio TECHNIQUES

Bill Orr W6SAI

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BY WILLIAM I. ORR, W6SAI

Happy New Year 1981! What will the New Year bring to Amateur Radio? We have the prospect of the new 10-MHz band in the near future and the probability that the 160-meter band will be expanded as the Loran-A equipment is deactivated. And, the sunspot cycle will continue its inexorable decline.

You probably won't notice much deterioration in DX conditions on the high-frequency bands this spring, but you might find that 10 meters is noticeably less "DX-y" this coming fall. And by spring, 1982, the slump in 10-meter openings should be quite apparent. So enjoy 10 while you have the chance. Pay attention to 6 meters, too; it might be a long time before 6 meters shows the long-distance communications that seem to be almost a daily occurrence these past years.

the popular triband Yagi beam

This column discusses the triband Yagi beam for 20, 15, and 10 meters. This well-known design is used (with impressive results) by many DXers, and it's an inexpensive and effective antenna that's not too big.

The modern triband Yagi was developed from a multifrequency dipole invented and perfected by Howard K. Morgan, Superintendent of Communications, Transcontinental and Western Airline, Inc. The requirement of the airline (the grandfather of the modern TWA) was for a simple, multifrequency antenna that would provide good reception of various aircraft frequencies at ground communication stations. The multifrequency dipole devised by Morgan was described in the August, 1940, issue of *Electronics*, and the original drawing from that article appears in **fig. 1**.

The Morgan antenna consisted of a center-fed dipole with the end insulators replaced by parallel-tuned circuits. Extra wire sections were added beyond the circuits so that the dipole was again resonant at a lower frequency.

For example, if the center dipole section is cut for 21.2 MHz and the parallel-tuned circuits (commonly called *traps*) are tuned to 21.2 MHz, the dipole works in a normal manner; the very high impedance of the resosulator. Wires that have been added after the traps have little, if any, effect on antenna operation at, or near, 21.2 MHz.

If wire sections are added after the traps are cut to the proper length, the overall antenna system will resonate at a lower frequency, say, 14.0 MHz. The presence of the tuned circuits affects the length of the antenna, so resonance is obtained at 14.0 MHz with an overall antenna length somewhat shorter than normal. A typical antenna is shown in **fig. 2C**. The traps act as electrical switches that are either open or closed, depending on the frequency of operation of the antenna.

Morgan's article pointed out that antennas for operation on as many as four different frequencies had been built successfully. Finally, the article provided detailed information concerning adjustment of the traps for proper antenna operation.

resurrection of the multiband dipole

Morgan's multifrequency antenna died a quick death. Here was the perfect antenna for operation on the various Amateur high-frequency bands; the traps were easy to build and adjust, low-impedance transmission line was readily available, yet nobody carried the idea forward. With the coming of World War II and the ban on Amateur Radio, the multiband-antenna principle fell by the wayside.

It was not until after the war that the concept of multiband operation surfaced again, in a design by Chester Buchanan, W3DZZ, described in the December, 1950, issue of *Radio and Television News*. Buchanan described a dual-band beam for 10 and 20 meters using trapped elements. He also provided the first complete description of how the trapped antenna worked. His final beam design, known to many DXers as the "W3DZZ beam," was fully described in *QST*, for March, 1955.

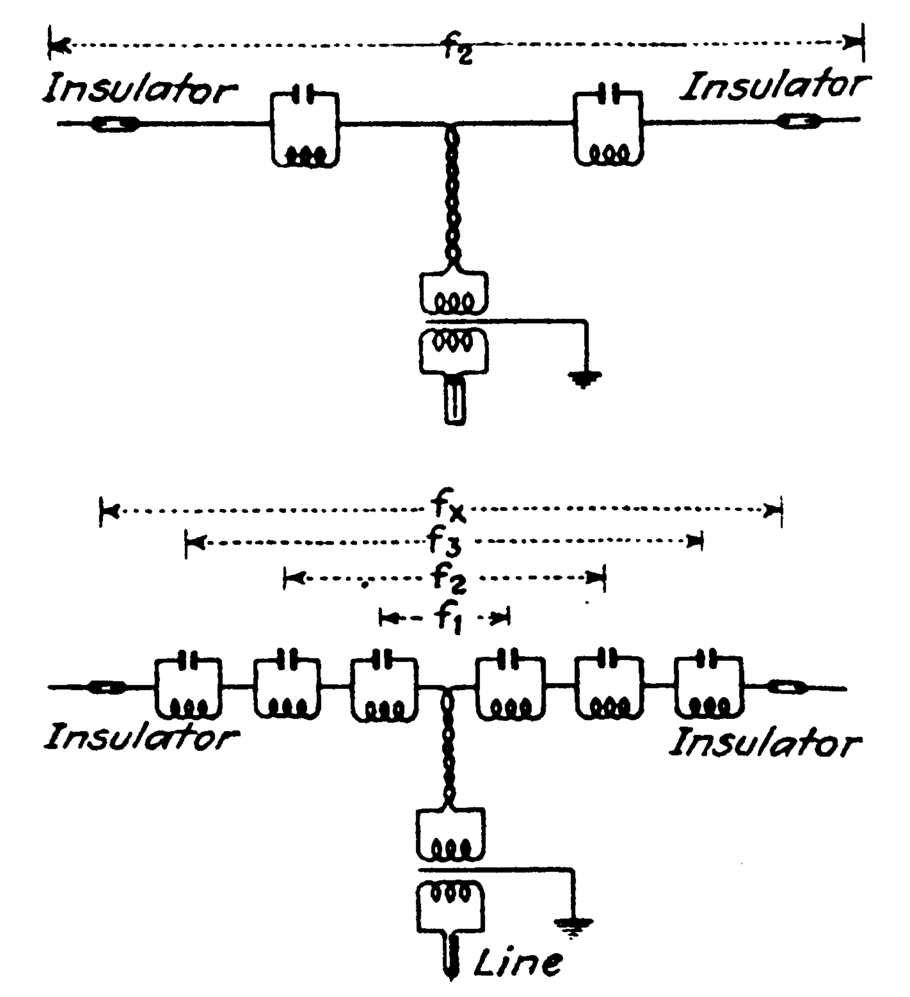

fig. 1. The original illustration in the issue of *Electronics* magazine depicting the trap dipole scheme evolved by Howard K. Morgan. The multifrequency antenna was originally designed for ground-station reception in the aeronautical service. Parallel-tuned trap circuits served as insulators at the resonant frequency of trap and antenna.

how the triband beam operates

Frequency-sensitive "switches" are the operating secret of the triband beam. The switches consist of a capacitor and inductance connected in parallel. This is a simple parallel-tuned circuit, which provides a very high impedance across the terminals at the resonant frequency.

The actual value of the impedance is the reactance of the coil times its Q, ($Q \times Q_L$). If the value of Q is high (Q being the electrical excellence of the coil), the circuit works as a high-impedance insulator at the circuit resonant frequency.

The curve in **fig. 2D** shows that the off-frequency reactance of the circuit is quite small: inductive at frequencies lower than resonance and capacitive at frequencies higher than resonance.

When the trap is placed in an antenna, the equivalent circuit of the antenna above and below resonance is shown in **fig. 2C**.

On 15 meters the center portion of the antenna works as a dipole with trap "insulators" tuned to 15 meters. When the antenna is used on 20 meters, the inductive reactance of the traps is quite low, and they act as loading coils. The wire length between the traps is cut so that the wire, plus the loading coils, is resonant at 20 meters, in conjunction with the center section.

Thus, on 20 meters, the trap dipole

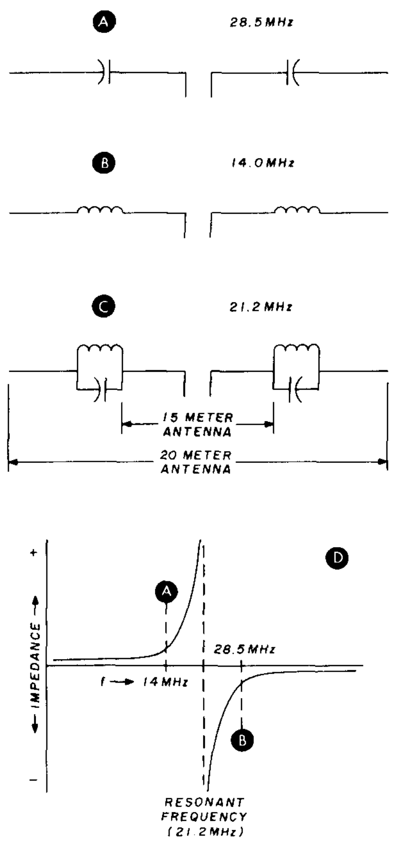

fig. 2. Operation of tuned trap for 20-15 meter, two-band dipole antenna. A: Equivalent circuit of antenna and trap at some frequency above resonant frequency of trap. Trap "looks like" a series capacitance. B: Circuit of antenna and trap at some frequency below resonant frequency trap. Trap "looks like" series inductance. C: At trap resonance, trap acts as a very high-impedance — the equivalent of an insulator. Examples given are for a trap tuned to 21.2 MHz (15 meters). D: Impedance chart of 15-meter trap showing reactance curve through the 14-28 MHz region. At 14 MHz the coil is the dominant component of the trap, and at 28.5 MHz the capacitor is the dominant component. Antenna is resonant at 14 MHz but is not resonant at 28.5 MHz. An extra set of traps is required for 10-meter operation, as shown in fig. 3.

is considerably shorter than normal due to a portion of the antenna being duplicated by the series inductance of the traps — the antenna is *nonresonant* on 10 meters unless extra traps are added.

To put it all together, in a triband element, the inner section and inner traps are resonant at 10 meters, the middle portion of the antenna and associated traps are resonant at 15 meters, and the whole antenna assembly is resonant at 20 meters (**fig. 3**).

trap performance

The trap is the heart of the triband antenna. A good trap will have reasonably high Q and must be waterproof. Many Amateurs make their own traps for triband dipoles[1] from an airwound inductance and a transmitting-type ceramic capacitor. The trap is placed in a waterproof housing.

Commercially made traps for triband beams are more sophisticated and are designed for mass production. The two traps in one section of an element may be combined into one structure, as shown in **fig. 4**. This arrangement provides a compact and rugged assembly.

No reliable information exists, as far as I know, as to the actual gain of a triband beam compared to a full-size antenna. Admittedly, the perfect trap has not yet been built, so some power is lost in each trap. In addition, on the two lowest bands, the antenna elements are not full size, so additional power is lost because of the reduced element length. This power is lost in the trap, which acts as a loading coil.

On the whole, the tribander design is good. The triband Yagi beams on the market *work*, and work well, judging from the number of DXers who use them and the robust signals they put out.

triband-beam bandwidth

One specification in which the triband beam suffers is bandwidth. On 10 meters, where the inner set of traps act as insulators, the bandwidth of the triband Yagi compares favorably with that of a conventional 10-meter Yagi beam. On 15 and 20 meters, operational bandwidth is somewhat restricted, because a portion of the element on each band is made up of the trap (or traps) for the higher-frequency band.

Then, too, some triband beams are built on shorter-than-normal booms to conserve space. This compromise further reduces operating bandwidth (and gain) — especially on 20 meters.

A set of representative SWR curves for a triband Yagi and a full size 20-meter Yagi is shown in **fig. 5**. Both beams are built on 20-foot (6-meter) booms. Observe that the 20-meter bandwidth of the tribander suffers in comparison with the full-size 20-meter beam, but bandwidth improves on 15 meters, and is equal for both antenna designs on 10 meters. This is of little consequence to the Amateur having a tube-type final-amplifier stage with an adjustable output network, but it poses a problem to those who have a solid-state output stage that requires an antenna with a low standing wave ratio.

One way around this problem is to build an SWR "flattener" that will reduce the SWR on the line at the transmitter end of the line (**fig. 6**). This simple matching network is placed between the coaxial line to the antenna and the station SWR meter. The capacitors and number of coil turns are adjusted for lowest SWR on the operating band. It can be easily adjusted for near-zero SWR at any point in the 10, 15, or 20-meter bands by tuning the controls for minimum SWR as observed on the meter. The settings can be logged for future use.

is a triband Yagi beam practical?

Based on personal observations over the years, the answer to this question is *yes*. If you have a well-made tribander, you have the tremendous advantage of three-band operation with one relatively small antenna. I've used a triband Yagi for years, alternating with a full-size 20-meter beam, on occasion. As far as working DX goes, I can do equally well with either antenna, and I notice no difference between the three-band design and the single-band beam.

Common sense and measurements made on the triband Yagi tell me that it isn't as efficient as the full-size beam. The bandwidth is somewhat restricted, and the front-to-back ratio isn't quite up to snuff, particularly on 15 and 10 meters. But these complaints fade away when I consider the convenience of working *three bands*, and the fact that I can compete in DX work and get reports that are just about equal to those of others in the area.

Tests. Before I installed the tribander, I wouldn't have believed this, as I took the traps into the company laboratory and measured the Q on a precision meter. A trap that I made out of the best materials available (air-wound, silver-plated, copper coil and a transmitting-type ceramic capacitor) provided a Q of over 300 at 30 MHz.

The Q of the commercial trap, measured at 30 MHz, was only about 180. This so discouraged me that the triband beam sat in my garage in the original box for about a year. Finally, deciding to see for myself how the antenna performed, the 20-meter Yagi came down, and the tribander went up in its place. Despite my misgivings, the triband worked, and worked well.

Some of my engineering colleagues sniffed in disdain at my unscientific test and were unmoved when I beat them out in a DX pileup. "Pure luck," was their conclusion.

Well, I don't know about that. Luck and operating skill surely are factors to be reckoned with. But if the antenna doesn't work, all the luck and operating skill in the world are to no avail.

Power transfer. It's true that some transmitting power is lost in the traps. I have a telescoping tower and can reach the traps in my antenna from the garage roof when the tower is retracted. Running a kilowatt input for 10 minutes, key down (when the band is dead!) results in the traps being slightly warm to the touch. Obviously, some rf power is being converted to heat in the traps. Other hams (having more ego than common sense) have attempted running excess power to a trap Yagi beam and have damaged the traps.

In conclusion, then, a good trap Yagi beam is an acceptable compromise for the Amateur who wants

table 1. Data for the G3LDO wire-beam antenna. All dimensions are in inches and are for insulated wire; multiply by 1.04 for uninsulated wire. Dimension C is approximate.

band	reflector	driven element	A	B	C	D
20	452	417	245	263	180	33
15	302	279	154	166	113	22
10	225	208	114	122	85	15

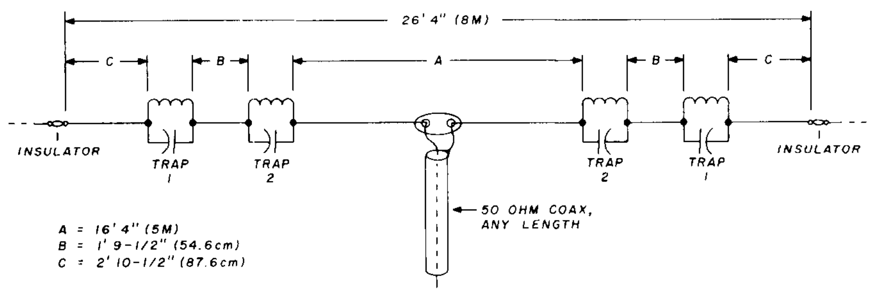

fig. 3. A triband element for 20, 15, and 10-meters. Dimensions are given for no. 16 (1.3mm) antenna wire size (not critical). Each trap is resonated to the design frequency before it is installed in the antenna. Length of each trap is about 2 inches (51 mm). Small ceramic or mica capacitors, rated at 3 kV, are used, which should have zero temperature coefficient.

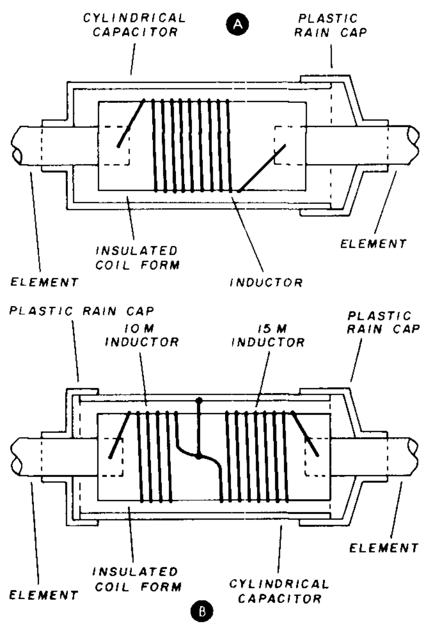

fig. 4. Typical trap construction. A: Single trap composed of inductor connected in parallel with cylindrical capacitor. The capacitor serves as an outer shield for the inductor and provides capacitance between cylinder and coil. End of assembly is sealed against weather with a plastic rain cap. B: Dual, two-band trap composed of two coils mounted within a single cylindrical capacitor. Number of turns on coils and placement within cylinder determines effective capacitance. Ends of assembly are sealed with plastic rain caps. Connection between coils and cylinder is made at center junction of inductors.

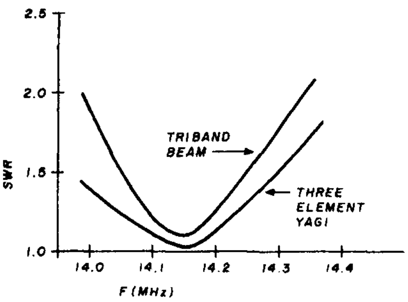

fig. 5. SWR bandwidth curve of typical triband Yagi beam is quite sharp on 20 meters, approaching 2 at band edges. A three-element, full-size Yagi exhibits a more moderate SWR curve for the same frequency span. On 15 meters, tribander bandwidth is somewhat improved and is essentially equal to full-size Yagi on the 10-meter band.

three-band operation. If a solid-state transmitter is used, an SWR "flattener" will prove helpful in making the transmitter perform at top efficiency.

the G3LDO wire

While on the subject of Yagi antennas, the interesting design by G3LDO is worth considering.[2] Experiments were run on 144 MHz with wire beams, and G3LDO came up with the interesting observation that the resonant length of a wire element depended upon the insulation on the wire. Uninsulated copper wire and enamel-coated copper wire provided "normal" dimensions, whereas insulated copper wire (hookup wire) had a velocity factor of about 0.965. The insulation on the tested wire was PVC (polyvinyl chloride).

Based on this information, G3LDO built a test beam on 2 meters, and then a larger model for 10 meters (**fig. 7**). He found that bending the elements back in the plane of the antenna caused an increase in the resonant frequency of the bent element and also resulted in a drop in gain. The solution was to fold the elements back in umbrella fashion, with the ends of the elements forming guys for the bamboo or fiberglass support structure. Dimensions for the beam are given in **table 1**. The performance of this simple and inexpensive wire beam was equal in every way to a standard equivalent design using full-size elements. This looks like a good antenna design for the ham who has a problem locating aluminum tubing.

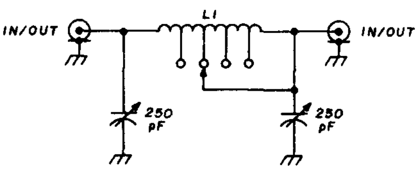

fig. 6. Simple SWR "flattener" for coaxial line. Capacitors are single-spaced receiving types for powers to 250 watts. Mica compression units can be used for low power. Inductor consists of 15 turns, 1-inch (25.4mm) diameter and 2 inches (51mm) long. Tap to coil is through a ceramic, single-pole rotary switch, such as Centralab 2501 (two to six positions, nonshorting). The coil is tapped about every other turn. Wire tap can easily be soldered to coil by depressing turn on either side of tap with screwdriver to allow tap wire to pass around a turn of the coil (coil may be a B&W miniductor or equivalent). Network is symmetrical; either terminal may be used for input or output.

a TVI filter for the 6-meter operator

Do you have a problem with 6-meter operation? It's a tough deal, what with TV channel 2 only a few megahertz away. Filters that can protect channels 2 and 3 (and provide attenuation of TV garbage in the 6-meter receiver) are hard to find.

My attention has just been directed to a new filter that will be of interest to all 6-meter operators. It is the *Unadilla/Reyco* Interfilter, specifically designed for 6 meters. It's rated for full Amateur power over the range 50-52 MHz and provides over 63 dB attenuation at TV channel 3. (Attenuation at channel 2 is somewhat less.) For full information on this interesting filter, write Unadilla/Reyco, 6743 Kinne St., East Syracuse, New York 13057.

references

1. Design and construction of trap antennas is covered in detail in *Simple Low-cost Wire Antennas*, by the author of this column. Available from Ham Radio's Bookstore for $6.95.
2. This material is abstracted from *Radio Communications*, a publication of the Radio Society of Great Britain, 35 Doughty Street, London WC1N 2AE, England.

ham radio

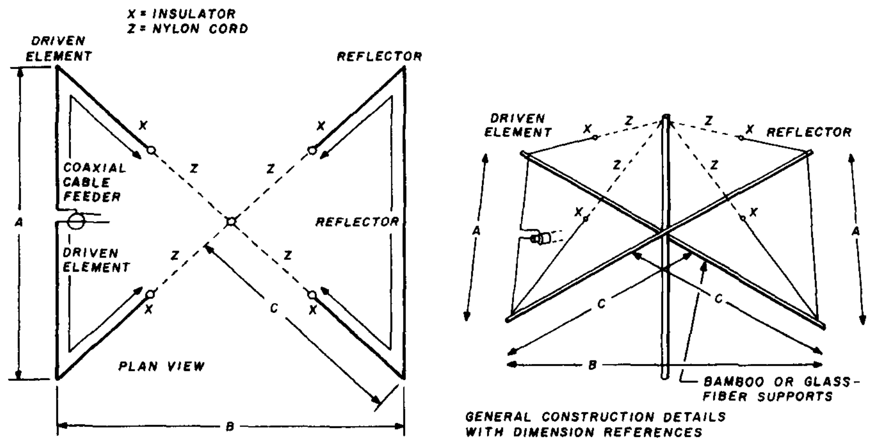

fig. 7. The G3LDO wire beam antenna for 10, 15, or 20 meters. Framework consists only of vertical mast and four bamboo or glass fiber support rods. Antenna is fed with coaxial line at midsection of driven element. Dimensions are given in table 1.

More Details? CHECK—OFF Page 110

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427	1.10
429	7.00
877	1.00
959	2.00
976	15.00
070	8.00
071	15.00
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589	6.00
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591	11.00
635	5.44
636	11.60
637	20.00
641	5.00
643	14.00
645	10.00
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More Details? CHECK—OFF Page 110

DX FORECASTER

Introducing Garth Stonehocker, KØRYW, who will present a series of columns on propagation and DX forecasting. Garth is a physicist and electronics engineer with extensive experience in propagation forecasting. He was associated with the Central Radio Propagation Laboratory and also did short-term radio propagation forecasting for the North Atlantic Warning Service in Boulder, Colorado.

First licensed in 1950 as W4RCF, then as W7ROY, Garth has been a member of several DXpeditions in conjunction with propagation forecast studies. Among them were an expedition to Antarctica (KC4USK) in 1956-58 and to the Cook Islands (KH6MG/ZK1) in 1958. Garth is a DXer and enjoys recording DX signals from beacons for propagation studies. Editor

New DXers

Are you a new ham who has come onto the thrill of your first DX with that new rig for Christmas, or an old one that has taken time out from years of rag chewing for a renewed go at DX chasing? If either, you may be interested in some propagation fundamentals and a look at which frequencies to use for putting the signal where you want it. To get your signal there the loudest of the bunch and to receive his probably weaker signal (other countries usually have lower power limits) is the goal. This can be done by fine-tuning your system for the particular location of the DX of your interest. That is, the antenna and ionosphere can be coupled together to increase the chance of their hearing you, and you hearing them, by the signal going to the right azimuth and out to the right distance.

The height of "reflection" in the ionosphere depends on the frequency band used. The angle of incidence into the ionospneric layer also affects what depth in the ionosphere the radio wave penetrates to be "reflected." Your antenna system needs to feed the maximum power at the correct vertical take-off angle to be maximum at the correct distance away and the correct azimuth for the correct direction (great circle bearing, usually). This is the best coupling to the ionosphere. The distances mentioned in the band-by-band summary show how the frequency penetration into the ionosphere varies the signal distance from band to band. A different antenna should be used for best results in short skip and in long skip; a high take-off angle for short skip and a lower angle for long skip. Articles on antennas often mention the take-off angles, or you can consult the antenna handbooks.

January is very similar to December in propagation effects. The ionosphere is a balanced energy system that changes slowly in its seasonal change from month to month. There is an intense but short meteor shower over a few hours sometime between January 2nd through 4th. It is known as the Quadrantid shower.

Band-by-band summary

Six meters will open occasionally for F2 long skip propagation with hops 1,000-2,500 miles long and with many hops usable. The openings will follow the sun during the day and early evening.

Ten and Fifteen meters will have openings similar to 6 meters, but more often and longer. Worldwide DX is usual from after sunrise until well after sunset during peirods of 27-day solar flux maxima. Short skip of 1,200 miles maximum distance is also possible and will also be following the sun across the earth.

Twenty meters will be open most all days and nights to some area of the globe, with long skip and plenty of short skip with similar distances and number of hops as the 15, 10, and 6 meter bands. This is the workhorse of the bands. Tall towers and high beams or quads are a natural for long skip for this band.

Forty and Eighty meters will be the most usable night-hour bands for DX. Most areas of the world can be worked between darkness until just before sunrise. Hops shorten on these bands to about 2,000 miles for 40 and 1,500 miles for 80 meters, but the number of hops can increase, since signal absorption in the ionospheric D region is low during the night. The path direction follows the darkness across the earth, similar to the higher bands following the sun. Daytime short skip can be used during the day and night, particularly if low-height horizontal antennas (high take-off angle) are used. Vertical antennas over good ground systems give the lowest take-off angles for long skip on these bands.

One-Sixty meters will be about like eighty meters and provide good stuff for the enthusiastic DXer who can come up with low-take-off angle antennas with good efficiency at the long wavelength of this band.

ham radio

JANUARY

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	10	—	15*	10	15	10	10	10
0100	5:00	10	—	20	10	15	10	10	10
0200	6:00	10	—	20	10	20*	10	10	10
0300	7:00	10	—	20	15	20*	10	10	15*
0400	8:00	15	40	20	15	20*	10	10	20*
0500	9:00	15	40	20	15	20	15	15*	20
0600	10:00	20	40	20	20*	20	15	15	20
0700	11:00	—	40	—	20	20	15	15	—
0800	12:00	—	20	—	20	20	20	20*	—
0900	1:00	—	20	—	20	20	20	—	—
1000	2:00	—	20	—	40	20	20	—	40
1100	3:00	—	—	—	40	20	20	—	40
1200	4:00	—	—	20	—	40	40*	—	40
1300	5:00	—	20	15	20	20	40	20	40
1400	6:00	—	20	15	20	—	40	40*	40
1500	7:00	—	15	15*	20*	—	40*	40*	40
1600	8:00	20	15*	10	10	20	20	20	40*
1700	9:00	20	15	10	10	—	15	15	40
1800	10:00	20	15	10	10	—	15*	15	40
1900	11:00	20	20	10	10	—	10	15	40
2000	12:00	40	20	10	10	—	10	15*	20
2100	[illegible]	40*	20	15	10	—	10	10	15
2200	[illegible]	20	20	15	10	—	10	10	10
2300	[illegible]	15	—	15	10	—	10	10	10

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	10	40	15	10	15	10	10	10	6:00
0100	6:00	10	40	20*	15	15	10	10	10	7:00
0200	7:00	15	20	20	15	15	10	15*	15*	8:00
0300	8:00	15	20	20	15	20*	10	15	20*	9:00
0400	9:00	20	40*	20	15	20	15*	—	20	10:00
0500	10:00	—	40*	20	15	20	15	—	20	11:00
0600	11:00	—	40	20	20*	20	15	—	20	12:00
0700	12:00	—	40	—	20	20	15	—	20	1:00
0800	1:00	—	40*	—	40	20	20*	—	20	2:00
0900	2:00	—	40	—	40	20	20	—	—	3:00
1000	3:00	—	—	—	40	20	20	—	—	4:00
1100	4:00	—	—	—	20	20	20	—	—	5:00
1200	5:00	—	—	20*	20*	20	20	—	—	6:00
1300	6:00	—	20	15	15	20	40*	—	—	7:00
1400	7:00	20	15	15*	10	—	20	20	40	8:00
1500	8:00	20	15*	10	10	—	20	20*	40	9:00
1600	9:00	20	10	10	10	—	15	15	20	10:00
1700	10:00	20	10	10	10	—	15	10	20	11:00
1800	11:00	20	15*	10	10	—	15	10	—	12:00
1900	12:00	—	15	10	10	—	10	10	—	[illegible]
2000	[illegible]	—	20	10	10	—	10	10	—	[illegible]
2100	[illegible]	—	20	15*	10	—	10	10	20	[illegible]
2200	[illegible]	20	20	15	10	—	10	10	15	[illegible]
2300	[illegible]	15	40*	15	15	—	10	10	10	[illegible]

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	15*	20	15	15*	20*	10	10	10
0100	8:00	15	20	20*	15	20*	10	15*	15
0200	9:00	15	40*	20	15	20*	15*	15	15
0300	10:00	15	40	20	15	20	20*	15	15
0400	11:00	20	40	20	15	20	20	15	20*
0500	12:00	20	40	20	15	20	20	20*	20
0600	1:00	—	40	20	15	20	20	20	20
0700	2:00	—	40	—	20*	20	20	—	—
0800	3:00	—	—	—	20	20	20	—	20
0900	4:00	—	—	—	20	20	20	20	—
1000	5:00	—	40	—	40	20	20	20	—
1100	6:00	—	20*	—	40	20	40	—	—
1200	7:00	20	15	15*	40	20	40	40	—
1300	8:00	20	10	10	15	—	40	20*	20
1400	9:00	20	10	10	10	—	80*	15	—
1500	10:00	20	10	10	10	—	40	15	—
1600	11:00	20	10	10	10	—	40	10	—
1700	12:00	20	15*	10	10	—	20	10	—
1800	[illegible]	—	15*	10	10	—	15	—	—
1900	[illegible]	—	20	10	10	—	10	—	—
2000	[illegible]	—	20	15*	10	—	10	10	—
2100	[illegible]	—	20	15	10	—	10	10	20
2200	[illegible]	20	20	15	10	20	10	10	15
2300	[illegible]	15	20	15	10	20*	10	10	10

the ham notebook

how accurate is that SWR meter?

A friend recently contracted a radio operator's most dreaded ailment: reflectophobia, the unwarranted fear of reflected waves on a transmission line. Fortunately, this was not a terminal case. I still remember the early morning phone call and the troubled voice saying, "Something is wrong with my antenna. The SWR won't go below 1.8." To make a long story as short as possible, there was nothing wrong with the antenna. The SWR meter he was using wouldn't read below 1.8. This made me think about SWR meters and their accuracy. Is there a way to check the accuracy of an SWR meter without using expensive or complicated test equipment? There is.

Basic transmission line theory tells us the SWR on the line will be the ratio of the load resistance to the line impedance.[1] If the load resistance equals the *line impedance*, the SWR will be 1:1. However, if the load resistance is either double or half the line impedance, the SWR will be 2:1. This is the key to a simple way to check the accuracy of an SWR meter. While it would be possible to use a 100-ohm resistor as a dummy load and measure the SWR, this method has several limitations. If the resistor has a 10 per cent tolerance, its resistance could be anywhere between 90 and 110 ohms; and it might not provide an SWR of 2:1. Additionally, carbon composition resistors can't dissipate more than a few watts, and anything but a QRP rig would overheat the 100-ohm dummy load. Using a wirewound resistor to increase the amount of power the dummy load can handle won't work because at higher frequencies the inductance of the wire in the resistor will have enough inductive reactance that the SWR won't be 2:1. Instead of using a 100-ohm dummy load with a 50-ohm transmission line to obtain an SWR of 2:1, we can use a 25-ohm dummy load and obtain the same results.

But where can I get a 25-ohm dummy load? It's easy. Just connect two 50-ohm dummy loads in parallel. Several commercially available dummy loads consist of a 50-ohm noninductive resistor in a bucket of transformer oil. The resistor itself can dissipate some power and the oil absorbs the heat, effectively increasing the amount of power the dummy load can safely handle.

Basic theory tells us the total resistance of two resistors in parallel[2] is $R = \frac{R1 \times R2}{R1 + R2}$. In this case, both resistors are 50 ohms; thus $R = 25$ *ohms*. Since $SWR = \frac{Z\,line}{Z\,load}$, then $\frac{50}{25} = 2.1$. The easiest way to connect two dummy loads in parallel is to use a coaxial T fitting (Amphenol 83-1T or equivalent) with short patch cords as shown in **fig. 1**. With two dummy loads in parallel, the load will

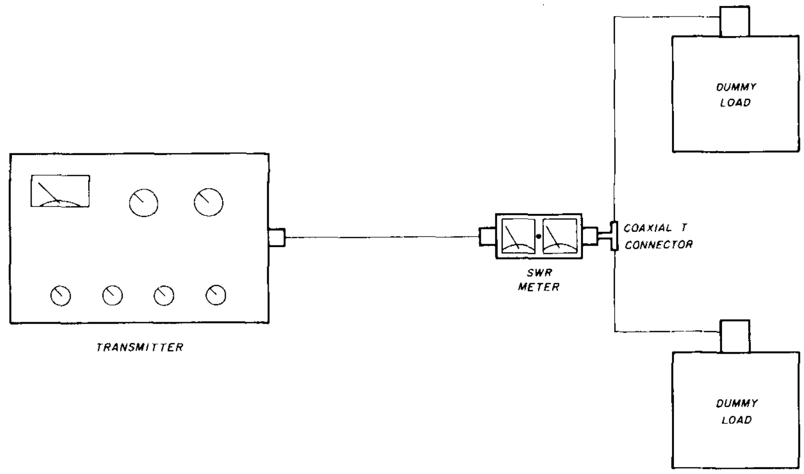

fig. 1

be 25 ohms and the SWR will be 2:1. If you are using a bidirectional wattmeter, the reflected power should be 11.11 per cent of the forward power. That's really all there is to it. This method of checking the accuracy of the meter can be used for hf, vhf, QRP, or even CB. In fact, any frequency and power level within the meter's capabilities can be used.

Now for the bad news. If your meter doesn't indicate an SWR of 2:1 when connected between a 50-ohm transmitter and a 25-ohm load, there might not be much you can do about it. Many lower priced SWR meters don't have any internal adjustments for calibration. Check the schematic for your meter to see whether it has a calibration adjustment. Even if your meter doesn't have a calibration adjustment, at least you'll know whether it reads high or low.

references

1. *Understanding Amateur Radio*, ARRL, 1977, page 96.
2. *Understanding Amateur Radio*, ARRL, 1977, page 9.

John W. Frank, WB9TQG

aligning Yagi beam elements

Assembling beam elements so that they are both in line with each other and parallel to the ground can be a frustrating experience when the eyeball or ground-assembly methods are used. The scheme shown in **fig. 2** is effective in providing a neat installation in a few minutes, and doesn't require any special tools other than a small level.

Roy Lehner, WA2SON

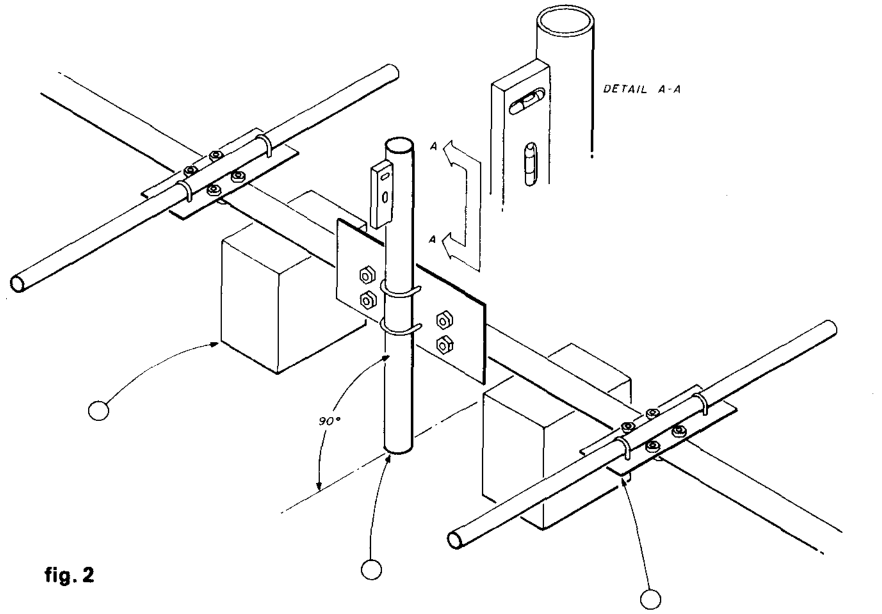

fig. 2

Drake R-4C receiver audio improvement

In the December, 1977, issue of *ham radio*, the article "Present-Day Receivers — Some Problems and Cures" suggests a 0.0015-μF capacitor be placed across R83 to correct a phase error in the Drake R-4C receiver.

I made this change and noted an improvement in audio quality. However, I also found another problem. With a headset connected to the earphone jack and the receiver in a standby position, an annoying shotgun noise was heard while transmitting. This noise was not noticed when the earphones were removed and the speaker alone was working.

A quick check of the R-4C circuit indicated that a 0.01-μF bypass capacitor was connected from the speaker terminal to ground. This effectively bypassed any rf picked up by the speaker leads. However, when the headset is plugged in, this bypass capacitor is disconnected from the circuit. The addition of a 0.01-μF capacitor from earphone jack to ground bypassed any rf picked up by the headset leads, and the problem was solved. Another solution is to remove the present 0.01-μF capacitor and reconnect it so it appears across both the earphone and speaker leads to the output transformer.

Bernard White, W3CVS

75S-3 alignment

During a recent rf circuit alignment procedure on my Collins 75S-3 receiver, I found a slight alteration that proved to be very beneficial to the reception of 10-meter signals. The modified procedure has been repeated a number of times with equal results.

The rf alignment procedure in the instruction manual is followed from steps **a** through **f**. After step **f**, repeak the OSC, RF, and ANT slug adjustments on the moveable platform; only a minor readjustment will be necessary. Once this is accomplished, continue with steps **g** through **o**. Then recheck the RF and ANT trimmer capacitor adjustments at 3.7 MHz. Again, a slight readjustment will suffice.

The foregoing should result in a marked improvement in the reception of 10-meter signals.

Paul K. Pagel, N1FB

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VIRGINIA: Frostfest 81 on January 11 (8AM - 4PM) at the Chesterfield County Fairgrounds, Chesterfield, Virginia. Prizes, new and larger facilities, exhibits, flea market, auction, contests and much more. Admission: $3.00 at gate. Talk-in on 146.34/.94 and 146.28/.88. Richmond Amateur Telecommunications Society, P.O. Box 1070, Richmond, Virginia 23208.

MICHIGAN: Oak Park Amateur Radio Club's Swap and Shop, Sunday, January 11, 1981 at Oak Park High School, Oak Park Blvd., Oak Park, Michigan. 8 AM to 3 PM. Admission: $2.00 with under twelve free. League table, door prizes, YL raffle and table. Refreshments. Talk-in on 146.04/64 and 146.52 simplex. For info and reservations, SASE to: Rob Numerick, WB8ZPN, 23737 Couzens, Hazel Park, MI 48030 or (313) 398-3189.

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JANUARY 17th and 18th: The Ford Tin Lizzy Club's North Metro Chapter's 3rd annual "Freeze Your Arctic Off" expedition from 2000Z the 17th to 1500Z the 18th. Handsome certificate available. More info: Box 545, Sterling Heights, Michigan 48078.

JANUARY 17th and 18th: Sponsored by West Virginia State Amateur Radio Council from 1700Z the 17th to 1700Z the 18th. Single operator only. Operation may be on all bands and repeater contacts are allowed. More info: N8AH, 933 Glen Way, South Charleston, West Virginia 25309.

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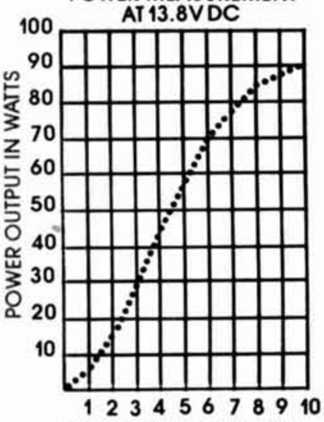

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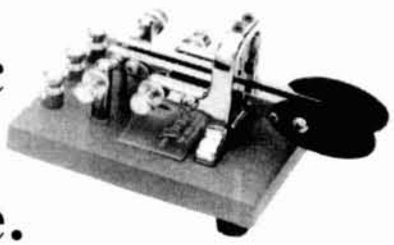

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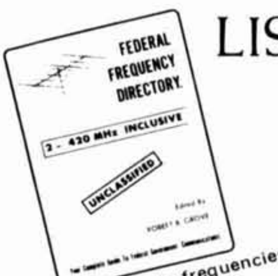

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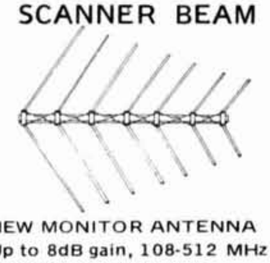

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MHz electronics

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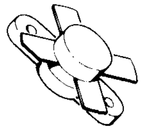
NPN SILICON RF POWER TRANSISTORS

. . . designed for power amplifier applications in industrial, commercial and amateur radio equipment to 30 MHz.



- Specified 12.5 Volt, 30 MHz Characteristics –
  Output Power = 80 Watts
  Minimum Gain = 12 dB
  Efficiency = 50%

MRF472

$2.50

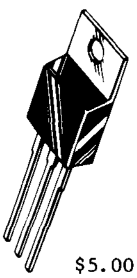
NPN SILICON RF POWER TRANSISTOR

. . . designed primarily for use in large-signal output amplifier stages. Intended for use in Citizen-Band communications equipment operating at 27 MHz. High breakdown voltages allow a high percentage of up-modulation in AM circuits.

- Specified 12.5 V, 27 MHz Characteristics –
  Power Output = 4.0 Watts
  Power Gain = 10 dB Minimum
  Efficiency = 65% Typical

MRF475

NPN SILICON RF POWER TRANSISTOR



$5.00

. . . designed primarily for use in single sideband linear amplifier output applications in citizens band and other communications equipment operating to 30 MHz.

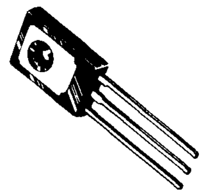
- Characterized for Single Sideband and Large-Signal Amplifier Applications Utilizing Low-Level Modulation.
- Specified 13.6 V, 30 MHz Characteristics –
  Output Power = 12 W (PEP)
  Minimum Efficiency = 40% (SSB)
  Output Power = 4.0 W (CW)
  Minimum Efficiency = 50% (CW)
  Minimum Power Gain = 10 dB (PEP & CW)
- Common Collector Characterization

MRF458 $20.68

NPN SILICON RF POWER TRANSISTOR

. . . designed for power amplifier applications in industrial, commerical and amateur radio equipment to 30 MHz.

- Specified 12.5 Volt, 30 MHz Characteristics –
  Output Power = 80 Watts
  Minimum Gain = 12 dB
  Efficiency = 50%
- Capable of Withstanding 30:1 Load VSWR @ Rated P_{out} and V_{CC}



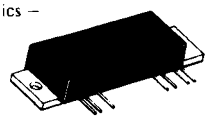
MHW710 - 2

$46.45
440 to 470MC

UHF POWER AMPLIFIER MODULE

. . . designed for 12.5 volt UHF power amplifier applications in industrial and commercial FM equipment operating from 400 to 512 MHz.

- Specified 12.5 Volt, UHF Characteristics –
  Output Power = 13 Watts
  Minimum Gain = 19.4 dB
  Harmonics = 40 dB



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MHz electronics

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390A-3	Microwave Diode Test Set	25.00
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ALTO		
34A1	FM TV Receiver 220MHZ	200.00
B&K		
161	Transistor Tester	75.00
607	Tube Tester	100.00
BALLANTINE		
303-06	AC Voltmeter	100.00
355	Digital Voltmeter AC/DC	75.00
6130A	Time Mark Generator	350.00
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2970	Color Video TV Camera with Monitor	300.00
BIRD		
67C	RF Wattmeter 0 to 2.5KW at 30 to 500MHZ	200.00
TS-118A/AP	RF Wattmeter 2 to 500W at 20 to 1400MHZ	150.00
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845AB	DC Null Detector/High Impedance Voltmeter	250.00
853A	Differential Meter	250.00
871A	DC Differential Voltmeter	250.00
GENERAL MICROWAVE		
454AR	Microwave Power Meter	100.00
GENERAL RADIO		
546C	Audio Frequency Microvolter	25.00
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740B	Capacitance Test Bridge	75.00
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1211C	Oscillator .5 to 50MHZ	100.00
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431C	Microwave Power Meter 10MHZ to 40GHZ 10μW to 10mW	350.00
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543A	Oscilloscope with Dual Trace Plug In 33MHZ	400.00
545A	Oscilloscope with Dual Trace Plug In 30MHZ	450.00
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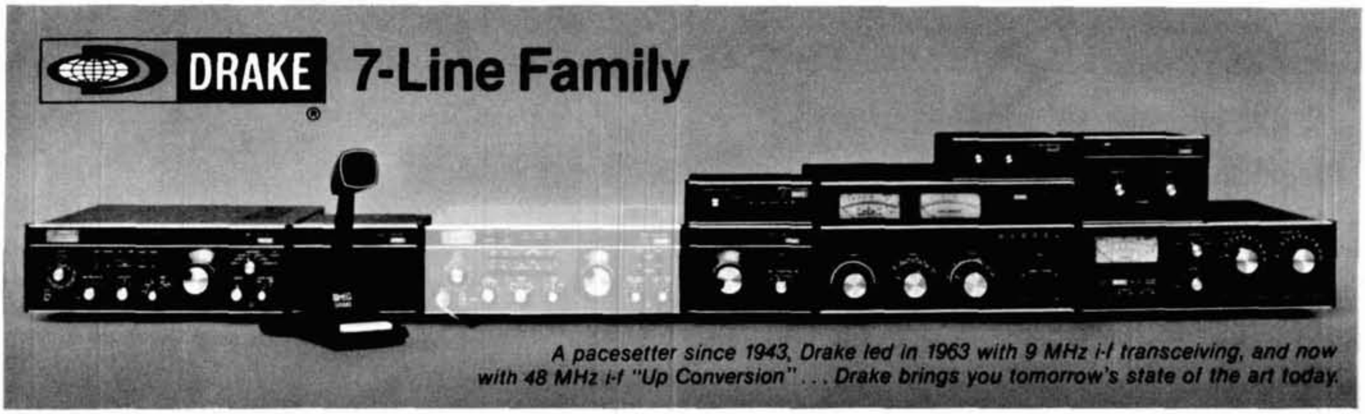

Model **1336**

TR7

solid state continuous coverage synthesized hf system

Continuous Frequency Coverage—The TR7 provides continuous coverage in receive from 1.5 to 30 MHz. Transmit coverage is provided for all amateur bands from 160 through 10 meters. The optional AUX7 Range Program Board allows out-of-band transmit coverage for MARS, Embassy, Government and Commercial services as well as future band expansions in the 1.8 through 30 MHz range.* The AUX7 Board also provides 0 through 1.5 MHz receive coverage and crystal-controlled fixed-channel operation for Government, Amateur or Commercial applications anywhere in the 1.8 to 30 MHz range.

Synthesized/PTO Frequency Control—A Drake exclusive: carefully engineered high-performance synthesizer, combined with the famous Drake PTO, provides smooth, linear tuning with 1 kHz dial and 100 Hz digital readout resolution. 500 kHz up/down range switching is pushbutton controlled.

Advanced, High-Performance Receiver Design—The receiver section of the Drake TR7 is an advanced, up-conversion design. The first intermediate frequency of 48.05 MHz places the image frequency well outside the receiver input passband, and provides for true general coverage operation without i-f gaps or crossovers. In addition, the receiver section features a high-level double balanced mixer in the front end for superior spurious and dynamic range performance.

True Passband Tuning—The TR7 employs the famous Drake full passband tuning instead of the limited range "i-f shift" found in some other units. The Drake system allows the receiver passband to be varied from the top edge of one sideband, through center, to the bottom edge of the opposite sideband. In fact, the range is even wider to accommodate RTTY. This system greatly improves receiving performance in heavy QRM by allowing the operator to move interfering signals out of the passband, and it is so flexible that you can even transmit on one sideband and listen on the other.

Unique Independent Receiver Selectivity—Space is provided in the TR7 for up to 3 optional crystal filters. These filters are selected, along with the standard 2.3 kHz filter, by front panel pushbutton control, independent of the mode control. This permits the receive response to be optimized for various operating conditions in any operational situation. Optional filter bandwidths include 6 kHz for a-m, 1.8 kHz for narrow ssb or RTTY, and 500 Hz and 300 Hz for cw.

Broadband, Solid State Design—100% solid state throughout. All circuits are broadbanded, eliminating the need for tuning adjustments of any kind. Merely select the correct band, dial up the desired frequency, and you're ready to operate.

Rugged, Solid State Power Amplifier—The power amplifier is internally mounted, with nothing outboard subject to physical damage. A Drake designed custom heat sink makes this possible. The unique air ducting design of this heat sink allows an optional rear-mounted fan, the FA7, to provide continuous, full power transmit on SSTV/RTTY. The fan is not required for ssb/cw operation, since normal convection cooling allows continuous transmit in these modes.

Effective Noise Blanker—The optional NB7 Noise Blanker plugs into the TR7 to provide true impulse-type noise blanking performance. This unit is carefully designed to maximize both blanking and dynamic range in order to preserve the excellent strong-signal handling characteristics of the TR7.

* NOTE: Transmitter coverage for MARS, Government, and future WARC bands is available only in ranges authorized by the FCC, Military, or other government agency for a specific service. Proof of license for that service must be submitted to the R. L. Drake Company, including the 500 kHz range to be covered. Upon approval, and at the discretion of the R. L. Drake Company, a special range IC will be supplied for use with the Aux7 Range Program Board. Prices quoted from the factory. See Operator's Manual for details. (Not available for services requiring type acceptance.)

Specifications, availability and prices subject to change without notice or obligation.

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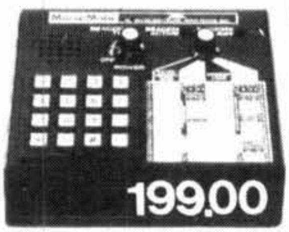

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		100	1.8	5.9
		200	2.6	8.5
		300	3.3	10.8
		400	3.8	12.5
RG8/u Regular .66VF	8237 28¢/ft.	100	2.0	6.6
		200	3.0	9.8
		400	4.7	15.4
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RG8/u Non-contaminating	8267 36¢/ft.	100	2.0	6.6
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