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ham radio

magazine

JULY 1980

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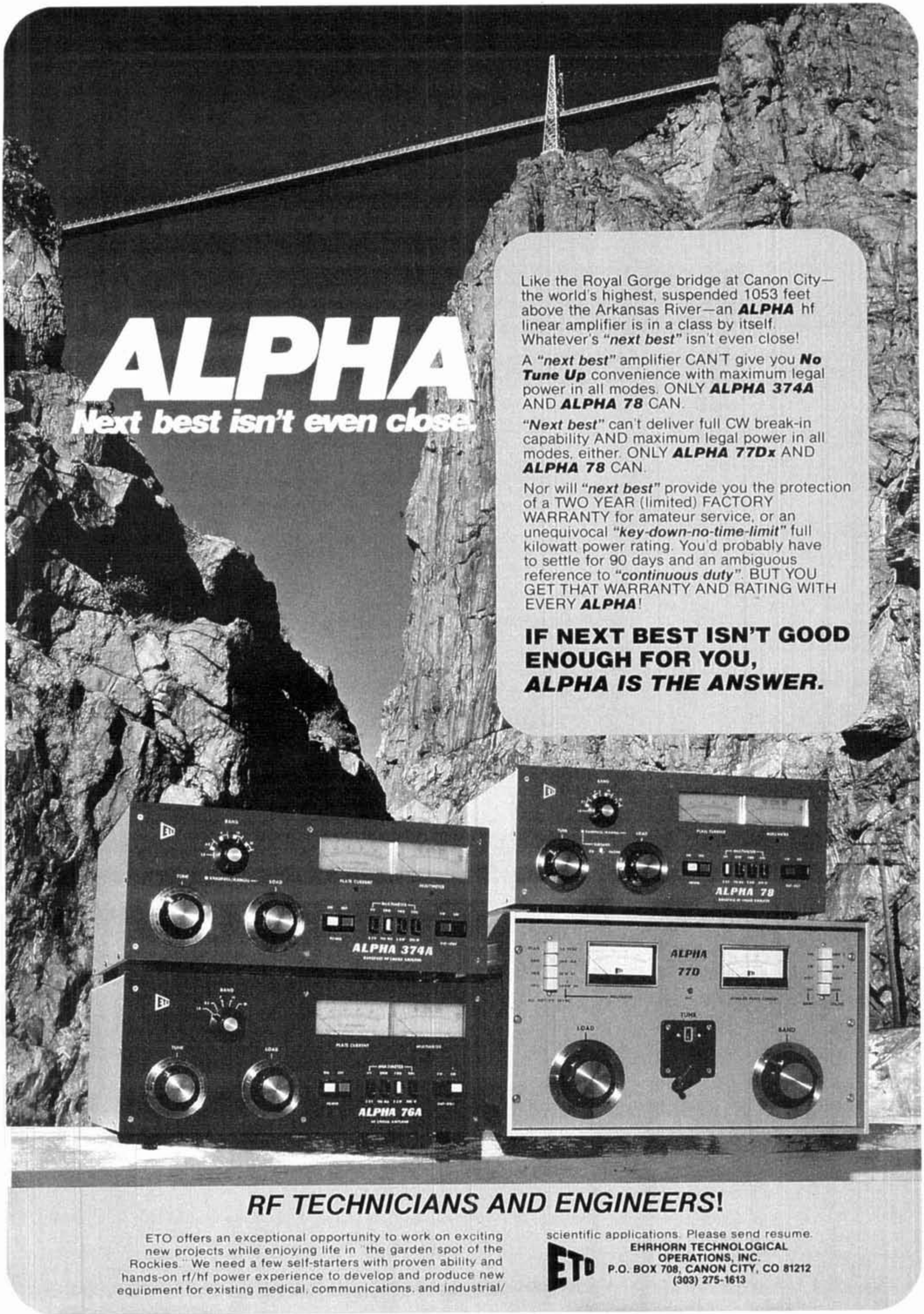
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More Details? CHECK—OFF Page 94

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JULY 1980

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contents

on top of the news

by Joe Schroeder

It is sometimes fashionable to say that the spirit of Amateur Radio has gone, and today's Amateur is an uninformed uninvolved appliance operator. What took place Friday evening, May 23rd, on 75 meters was a stinging rebuttal to that gloomy assessment. Every night that week AMSAT members and other interested Amateurs had been meeting on 3850 at 0200Z for a progress report on the launch of AMSAT's Phase-III satellite. After one of the French Ariane launch vehicle's four rocket motors failed during launch that morning and put Phase III (with a lot of other expensive space projects) into the Atlantic Ocean near Devil's Island, the group met once again to commiserate with each other over the disaster.

What began as a wake quickly turned into an almost unprecedented outpouring of support. With AMSAT's President, Tom Clark, W3IWI, as net control, most of the active North American participants in Phase III's development were joined by perhaps one hundred other check-ins from across the United States, Canada, and even Cuba. There were eulogies for the Phase-III bird and the loss of its unique capabilities, to be sure, but the predominant message over the following several hours was, "Let's keep moving ahead!" And this message came, significantly, not only from the already involved AMSAT membership but from bystanders — many of whom checked in to say: "I haven't been on OSCAR yet but always admired what you guys were doing. With your loss today it's time I became involved, so my check for membership plus a contribution is in the mail. How else can I help?" Needless to say, such support provided a priceless boost for those who'd heard their efforts of the past several years splash down in the ocean just twelve hours earlier . . . and they reflect that real Amateur Radio spirit that has too often been passed off as dead.

What does the loss of Phase III mean to AMSAT, and along with AMSAT, to the Amateur community? It means the loss of years of very hard work by a relatively small group of Amateurs in a half dozen countries. It means the loss of the $150,000 in hard cash that AMSAT invested in Phase III. It means the loss of more than a year, and possibly several years, before a new free-world Amateur satellite can be put up to replace OSCAR 7 (still operational well beyond its designed lifetime but showing its age) and OSCAR 8.

What's needed to keep our space program going? First and foremost, money, and plenty of it. Space efforts cost money, and the kind of sophistication that makes our "amateur" satellites suitable traveling companions for the best efforts of the pros cannot be accomplished on a shoestring. The Phase III investment brought AMSAT's treasury to a dangerously low point, and it's going to need rapid infusions of new money if we are not going to lose momentum. The second need is participation, people to volunteer for all kinds of tasks from bookkeeping and basic administration to state-of-the-art design work and computer programming. Finally, AMSAT needs members, for, by joining, an Amateur becomes not only a contributor but an *involved* contributor.

The response to AMSAT's needs was almost instantaneous after the news of Phase III's loss. Following the pledges on the Friday-night net, AMSAT's mail box has been bulging with new member applications and contributions. Within a few days, Amateur Electronic Supply in Milwaukee, Ham Radio Center in St. Louis, and the Ham Radio Publishing Group had all pledged $1,000 each to AMSAT, and many more industry contributions are expected.

What can you do? Join AMSAT. Annual dues are only $10 a year before July 1, $20 thereafter; life memberships are now $100, going up to $200 after July 1. Contributions to AMSAT are tax deductible. AMSAT, Box 27, Washington, D.C. 20044 is the place to send your check. Do it today and help demonstrate that Amateur Radio spirit is as real now as it ever was!

Joe Schroeder, W9JUV
associate editor


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comments

factory service

Dear HR:

The editorial in the January, 1980, issue of *ham radio* is interesting, but somewhat misleading. Even though it's stated that some manufacturers require their dealers to provide warranty service, the general impression, after reading the editorial, is that all Amateur dealers apparently do not have the capability to provide adequate after-sales service.

I agree it is a bad situation when a manufacturer requires its dealers to provide the warranty service, and then does not supply the dealer with up-to-date service information or parts.

However, Trio-Kenwood goes to great lengths to avoid this situation. First, every authorized Kenwood dealer is required to have the facilities to perform after-sales service. The customer may send his equipment either to the dealer or to us for warranty service. We are constantly mailing up-to-date service information to our dealers, and they are equipped with all of our service manuals and other technical material.

When you bring a rig to an authorized Kenwood dealer for repair, you are bringing it to a factory-trained technician. We conduct service seminars each year for the dealers, providing them with several days of intensive instruction. Our dealers are reimbursed for parts and labor, and are adequately stocked with the most commonly needed parts. Therefore, in most cases, the local Kenwood dealer can repair a rig as quickly and as thoroughly as we do, saving the customer several days of shipping and avoiding possible shipping damage or loss.

The "Kenwood Users' Report" in the January, 1980, issue of *Ham Radio Horizons* shows that more rigs are serviced by the dealers than by us, and that the majority of owners are *satisfied* with the service; and local service is certainly much more convenient for the customer.

We support the local dealer by training him to perform after-sales service. We *must* support the local dealer, because Amateurs depend on him for advice on the selection and use of equipment, as well as fast and competent service. A *responsible* Amateur Radio manufacturer, besides offering full factory-backed warranties, also keeps his dealers fully prepared to offer after-sales service. I wish your editorial had recommended contacting these dealers first, for service.

Kenneth M. Bourne, W6HK
Manager, Marketing Services
Trio-Kenwood
Communications, Inc.

We did not mean to imply that customers who require service should not seek help from their dealers first; we simply wanted to point out that dealer service is only as good as the support those dealers receive from the manufacturers, and not all manufacturers have the extensive dealer support program sponsored by Trio-Kenwood. Too many Radio Amateurs take good customer service for granted, when it is not always available — regardless of the reputation or integrity of the dealer.

Editor

ni-cad battery charging

Dear HR:

I enjoyed WA6TBC's article on the any-state ni-cad charger in the December issue of *ham radio*. Ni-cad batteries are something of a problem . . . they are, to a point, difficult to maintain. They either run down entirely and reverse polarity in one or more cells, or else are damaged by mere overcharging. My own experience has shown, however, that batteries are better kept on constant charge (a trickle); that minimizes annoyances all around. For example, I have a Vivitar electronic flash unit which has been on charge all the time for more than a year; test shows the unit to be operating very well. I installed a 100-ohm series resistor to reduce the charging current; this was placed right at the charger jack inside the flash unit. My impression is that ni-cad batteries tolerate all-time trickle charging. At any rate, this trickle charging insures that I have an operative unit always ready for use.

I might mention that I run all my calculators on chargers, and to date I have never had any problem with inoperative batteries due to "overcharging." Acquaintances tell me that they run their Hewlett-Packard calculators all the time off their chargers. I even run my Non-Linear Systems Miniscope on its charger all the time; the system has an automatic current limiting system built in.

At any rate, ni-cad batteries and gel batteries are a mixed blessing. Convenience dictates that each calculator, electronic flash unit, or oscilloscope be ready and running on a second's notice, without regard to state of battery charge. Hence the need for constant trickle charging.

Robert H. Weitbrecht, W6NRM
Belmont, California

(Continued on page 12)

presstop

AMSAT'S PHASE-III SATELLITE WAS LOST May 23 when one of the four Viking rocket engines on the French Ariane launch vehicle lost power after liftoff, sending it into the Atlantic. Liftoff was at 1429:42Z, a few seconds before the launch window closed, following a countdown delayed by rain and minor technical difficulties. After the launch, the onboard computer was unable to hold course because of uneven thrust, and only a minute or so into the flight the Ariane exploded from either fuel tank rupture or range officer command. Amateurs throughout the world heard the sad event unfold via the ALINS net on 10, 15, and 20.

About $150,000 And Thousands of hours worth of work by Amateurs in many countries is now on the ocean bottom, along with the "Firewheel" experiment. Phase III was not insured, because companies won't write such insurance until after the launch vehicle has four successful flights. This was Ariane's second flight. Fortunately, a duplicate Phase III structure had been built and, along with a full set of solar panels, circuit board art, software and circuit design, is available for another Amateur satellite. One could probably be assembled in less than a year.

Finding A New Launch opportunity is a major problem facing the Amateur space community. The European Space Agency is one good possibility, though this failure may set back their schedule for regular Ariane launches in 1981. The U.S. Space Shuttle is a possibility, though it's also behind schedule. Other possibilities will surely come along, but when is impossible to predict. Until a commitment is firmed up another Amateur spacecraft can't be completed, because it must be tailored to the launch vehicle's needs.

AMATEUR RADIO COMMUNICATIONS has been deeply involved in the aftermath of the eruption of Washington state's Mt. St. Helens volcano. First word of the disastrous explosion that blew the top off the mountain on May 18 came from an Amateur whose camper has been on the mountain's slope. After describing the beginning eruption, he ended his transmission with, "I'm getting the hell out of here!" He, along with two other Amateurs and many others, have been unaccounted for since.

About 200 Amateurs, half of them in locations near the eruption or working directly with rescue crews as search and welfare observers, and the other half serving at various key locations around the state, worked around the clock after the volcano blew. Many were involved in sample gathering and prediction work, an activity that began with the initial eruption on March 27, as well as handling all kinds of traffic for various local, state and federal agencies. Amateur Radio proved more effective than the telephone for handling much of the emergency traffic.

THE EXODUS OF REFUGEES FROM CUBA has been assisted by Radio Amateurs, both en route and following arrival in Florida. Some of the estimated 2000 small boats ferrying Cubans between Mariel Harbor in Cuba and Key West or other South Florida ports have used Amateur Radio to coordinate their efforts, mostly on 40 meters. WB4RDD and WD4EHO have headed up efforts on the 7175 kHz relief net, assisted by WD4LLT, WD4LAP, KA4JAI, KA4AQQ, and others.

In The Miami Area, 2 Meters has been used to aid the refugees. The SIRA repeater, with WD4PNS NCS, has been used extensively to assist authorities in locating friends and relatives of the arriving refugees. Area Civil Defense stations and other area repeaters are also helping in the relocation efforts and the SIRA club station, WB4ESB, operated on both 40 and 2 meters 18 hours a day for the operation.

BRITAIN WILL HAVE A CB service shortly, as the Home Office announced recently that it will relax its long-standing objections to establishing a Citizens Radio Service. When the new service will begin is uncertain, however. The 27-MHz band is used in Britain for hospital paging, so another frequency range will have to be found.

Though Britain Has Prohibited the use of CB radios, there are an estimated 50-60,000 CBers active in the British Isles, 10,000 of them in London alone. Illegal CB operators have caused real problems to some Amateurs. During his recent visit to London, HR Report Editor W9JUV was told "horror stories" by several stations he worked on 2 meters about Amateurs whose cars had been confiscated — and one Amateur who was jailed by over-zealous bobbies convinced the Amateur was actually an illegal CBer.

HMR COMMUNICATIONS WAS FOUND guilty in a Greensburg, Pennsylvania, Arbitration Board hearing May 20. Henry M. Robbins, Jr., of HMR was ordered by the three-judge panel to pay Gary Bratten $460 (on his $750 claim against HMR) plus $60 court costs, and Bratten was also allowed to keep the HMR equipment he'd received. Many other claims against HMR are still pending.

The U.S. Post Office is also investigating HMR for possible fraud, and wants a complete, detailed rundown on any dealings with HMR. Copies of all correspondence including envelopes, cancelled checks, and the like are requested of any Amateur who's done business with HMR and isn't involved in one of the 24 cases they're already investigating. Contact W.H. Lewis, Jr., U.S.P.S. Postal Inspector, Box 2068, Pittsburgh, Pennsylvania. Refer to case #242-81362-F(2).

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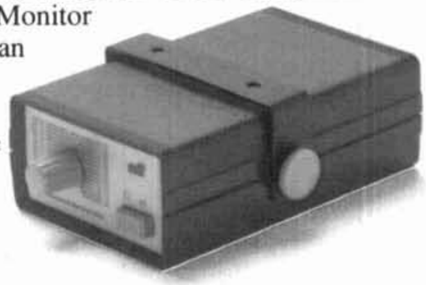

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comments

(Continued from page 6)

frequency synthesizer

Dear HR:

Tom Cornell's article on the CMOS 2-meter synthesizer in the December issue is an example of a well-designed circuit, along with a down-to-earth description that will probably warm the soldering irons of many interested readers. It is for these reasons that I would like to suggest that any would-be builders examine the owner's manual of their radio to determine the method of frequency modulation.

In the article, Tom mentions that he was using the synthesizer with a Regency HR-2B. Undoubtedly, the synthesizer will work fine with that radio because it uses a reactance-type phase modulation technique. It has been my experience, however, that most present day crystal-type rigs and transmitter strips use varactor diode crystal rubbering techniques to deviate the carrier. Thus, using the synthesizer with this type transmitter will produce a clean carrier, but will lack any modulation. Therefore, would-be builders would be well advised to check their rigs before buying parts.

Don Cwynar, WA3AXS
Reading, Pennsylvania

You bring up a good point. Older fixed-frequency and crystal controlled fm rigs such as the Regency HR-2B used by author K9LHA were based on a phased modulator which work fine with the CMOS synthesizer — newer equipment often has a varactor across the crystal to deviate the carrier. For owners of these newer rigs K9LHA is designing a phase modulator which will allow the use of his CMOS synthesizer; it will be published later this year in ham radio.

Editor

EI2W six-meter report

Dear HR:

EI2W commenced operations on the six-meter band on 20 October, 1979, when VE1AVX was worked at 1423 GMT. This report is for the period from 20 October to 20 December, 1979 (inclusive). VE1AVX with 1000 watts of power and an 11-element beam on a 30-foot (10 meter) boom has been the outstanding signal on the band.

EI2W has been using low power; a FT620B transceiver, kindly loaned by South Midlands Communications Limited, of Southampton, England; output about 10 watts. The beam used is a 3-element spaced 0.2 wavelength and a folded dipole driven element. During the two months' operation 1552 QSOs were made with approximately 600 different stations in all W/K call areas plus VE1, VE2, VE3, VE4, XE, KP4, and VP2 (Virgin Island).

Activity has been much greater during this cycle than during the International Geophysical Year (IGY) 1957/58. This has been due in some part to the use of SSB as against a-m during the IGY period.

The best day's operation was on the 18th of November when 106 stations were worked in all W call areas and VE1-2-3-4 and VO; very little fading was noticed on that day. The highest recorded MUF at this station during the period under review was on December 15th when it rose to 62.75.

On December 11th KØSFH was worked in Kansas, the USA station only using 3 watts; his signal was R5, S7/8 in Dublin.

A total of forty-three states have now been worked including California, Nevada, and the Dakotas. Tests with a tilted antenna will be carried out during the middle of January with VE1AVX; this arrangement met with great sccess during the IGY. Information on propagation and F2 Layer working is being collected and a full report on the present cycle will be made later in 1980.

H.L. Wilson, EI2W
9 Haddington Lawn
Glenageary, County Dublin
Republic of Ireland

More Details? CHECK—OFF Page 94

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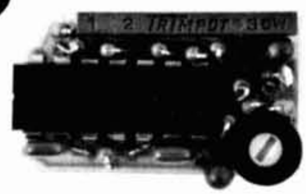

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rotary-dial mechanism

for digitally tuned transceivers

An ingenious application of up/down counters and photo-detectors for continuous tuning using a single knob

More and more ham equipment now uses digital synthesizers for frequency control. The usual method for programming the synthesizer is with rotary or thumbwheel switches. Although this works well for channelized communications, such as that on the 2-meter fm band, it's cumbersome for continuous tuning, such as on the high-frequency bands. For continuous tuning it's necessary to program the synthesizer indirectly, through an up/down counter.

Tuning can then be done, for example, by using two pushbutton-operated pulse generators, one of which sends a slow series of pulses to make the counter count up, and the other to make it count down. Fast and slow pushbuttons can be added, or alternatively a joystick and variable-rate pulse generator can be used to vary the tuning speed. Most of us, though, are used to tuning with a knob, and this is still about the most flexible method. In building a frequency-synthesized high-frequency transceiver, I couldn't find a readily available digital dial knob mechanism suitable for use with an up/down counter, so I built my own.

The mechanism I built is an improved version of that described by Earnshaw.[1] His readout consisted of a metal disk with holes punched in it through which two phototransistors viewed a small light bulb. By suitable placement of the detectors and with the appropriate logic circuitry, it was possible to generate separate pulses as the shaft was rotated: one for each direction of rotation (how this works is discussed in detail later).

Since I needed at least a hundred pulses per shaft revolution, and didn't want to spend time drilling holes, I used photographic methods to generate "holes" in the disk. The new disk was made by contact-printing the computer-generated pattern shown in **fig. 1** onto heavy graphic arts film.* Also, additional CMOS logic circuitry was added to provide twice as many pulses per revolution as there are marks on the disk, as well as circuits to drive several types of counters.

Instead of the phototransistors, photointerruptors specifically designed for this type of application were used. Given the mask, only simple tools and readily available parts are needed to duplicate this unit. It can be wired to give 50, 100, or 200 pulses per revolution.

A single photodetector can be used to indicate the shaft rotation rate, but not the direction. To tell the direction of rotation it's necessary to use a second detector spaced one-fourth the mark-to-mark distance (or an odd multiple thereof) from the first detector. To see how this works, consult **fig. 2**. In the position shown, both detectors sense a dark region (mark), which means that their outputs are at logic 0.

Suppose the disk is moved to the left. As the mark goes past detector B, its output goes to 1. We will indicate this as $\bar{A} \cdot B\uparrow = L1$; that is, if **B** changes from 0 to 1 while **A** is 0, we generate a left-rotation pulse. As the shaft continues to rotate, **A** will go to 1 while **B** remains 1; we denote this $A\downarrow \cdot B = L2$. Continuing, **A** will remain as 1 as **B** goes back to 0, denoted by $A \cdot B\downarrow = L3$; and finally we get $\downarrow A \cdot \bar{B} = L4$. We are now back to a position with both detectors over a mark, the way we started. Now consider moving the mask to the right. The sequence of transitions will be $A\uparrow \cdot \bar{B} = R1$, $A \cdot B\uparrow = R2$, $A\downarrow \cdot B = R3$, and $\bar{A} \cdot B\downarrow = R4$.

Although there are eight possible combinations of transitions from the center of one mark to the center of the next, it's not good practice to use both the turn-on and the turn-off transitions at the same posi-

*Photo disks are available from the author for $1.00 each. Please send a self-addressed, stamped envelope.

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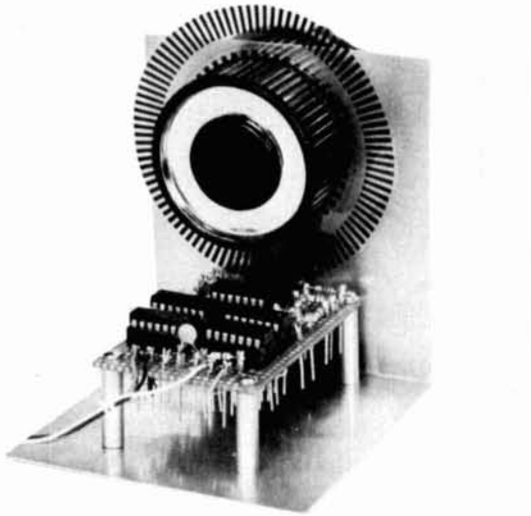

The assembled prototype mounted on an aluminum angle bracket for checkout and alignment.

tion; the encoder might then be sensitive to vibration and electrical noise when the detector is at the very edge of a mark. Consequently, in the circuit described in this article, only four of the eight transitions are used: thus $A\uparrow \cdot \bar{B}$ is used but $A\downarrow \cdot \bar{B}$ is not. As a result, there is a 1/4-mark backlash in the dial, but in practice this is not noticeable.

The schematic of the circuit necessary to process the detector signals is shown in **fig. 3**. I used GE type H21A5 interrupters, but TI type TIXL45 and Monsanto type MCA8 should work as well (all these types use a photo-darlington detector transistor to avoid the need for an output amplifier). Because the opto-couplers have a gain of about 1/20, the load resistor on the output transistor should be about twenty times the value of the current-limiting resistor to the LED emitter. The values shown for these resistors (R1-R4) are for 5-volt operation; they should be increased in proportion if the circuit operating voltage is made higher (the circuit will work over the 3 to 15 volt range).

The outputs of the detectors are squared up by Schmitt triggers U1A and C, and the complements of these signals are generated in U1B and U1D. Thus A, $\bar{A}$, B, and $\bar{B}$ signals are available from this gate. Type D flip-flops (U2 and U3) are used to sense the transitions from light to dark. U2A has $\bar{B}$ applied to its data (D) input and A applied to its clock input; it therefore clocks (in the notation used above) on $A\uparrow \cdot \bar{B} = R1$. (The flip-flop is reset later through the R input when B returns to 1.)

As the flip-flop sets, $\bar{Q}$ goes to 0 and produces a negative-going pulse (with duration determined by C1 and R5) at pin 1 of NAND gate U4A. Similarly, U2B generates pulses at pin 2 of this gate on the $A\downarrow \cdot B = R3$ transition. Normally, both inputs of this gate are at 1, so its output is at 0. When either of its input pulses goes to 0, its output briefly goes to 1. This pulse is buffered and inverted by U3A to provide a negative-going TTL-compatible output pulse (discussed further below). In similar fashion, pulses for the other direction are available at the U5B output. I've labeled these pulses CW and CCW, although depending on the mechanical arrangement, CW may actually produce pulses for counterclockwise rotation and vice versa.

Not all up/down counters use separate up and down clock pulses; some use a "direction" (U/$\bar{\text{D}}$) control signal and a single clock pulse. To generate signals compatible with these requirements, a flip-flop consisting of U4C and U4D is used. A negative-going pulse at pin 8 (the CCW pulse signal) sets this flip-flop, while a CW pulse resets it through pin 13. Both polarities of the direction signal are available at the outputs of this flip-flop ("CCW direction" and "$\overline{\text{CCW}}$ direction"). The two direction pulses are first combined by U5C, delayed by C5 and R9 to allow the direction control signal time to stabilize, and then buffered by U5D to produce a negative-going "$\overline{\text{either}}$" pulse.

construction

A photo of the prototype digital dial is shown; it

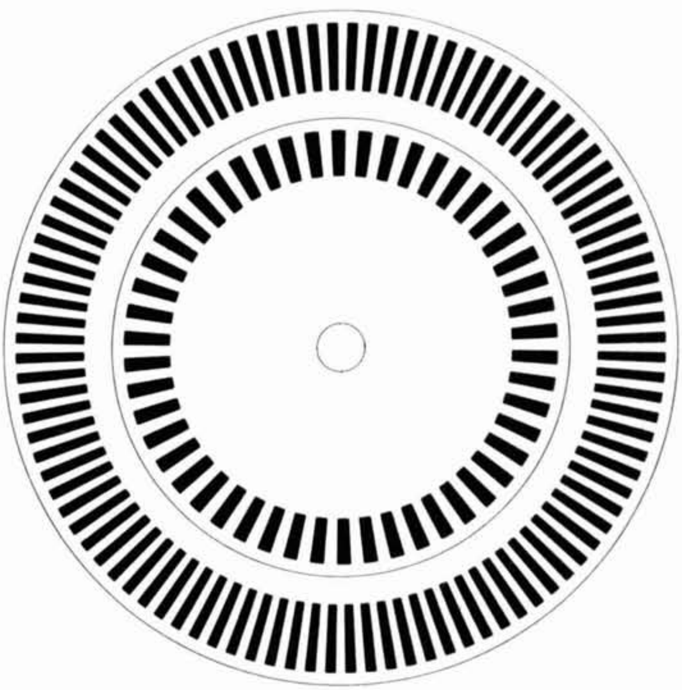

fig. 1. Optical mask used to encode shaft rotation (shown actual size). Inner set of marks generate 50 pulses per revolution; outer set 100 pulses per revolution.

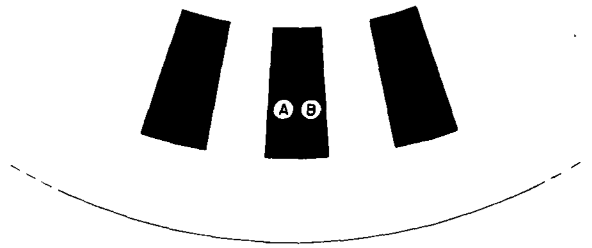

fig. 2. Fields of view of the two optical detectors, A and B, spaced one-quarter the mark period. If the mask is rotated leftward, detector B will sense light before detector A; conversely, if the mask is moved to the right, detector A will change first. The appropriate transitions are electronically processed to generate pulses indicating the extent and direction of rotation.

was assembled on an aluminum angle bracket for checkout and alignment. The whole assembly was then mounted in the final enclosure.

A method must be found to make the disk rigid and to mount it to the shaft. I resolved both problems by cementing the disk with epoxy to a large knob, slightly smaller in diameter than the mark pattern. For good adhesion, it helps to roughen the region on the disk and the knob where the cement will be with fine sandpaper (protect the rest of the disk with adhesive tape). The knob is then attached to a panel bearing, such as the Millen 10061; no doubt a satisfactory bearing could be salvaged from an old tuning drive, ten-turn pot, or something similar.

The electronic circuitry and detectors are mounted on a small section of perforated fiberglass board. The detector mounting holes should be slotted to allow adjustment of their separation. The gap between the two interrupters should be about 1/16 inch (1.6 mm) for the 50-mark pattern and about 1/64 inch (0.4 mm) for the 100-mark pattern. Point-to-point wire-wrap techniques were used for wiring. The board was mounted on the aluminum bracket with 1-inch (2.5-cm) aluminum stand-offs.

alignment

Alignment is not critical. Shaft height and knob position should be adjusted so that the disk always clears the interrupters but so that sector marks pass in front of the detectors at all shaft angles.

Apply power and check with a scope or voltmeter that the **A** and **B** signals oscillate as the shaft is rotated. Check that the "direction" signal changes when the direction of rotation is reversed; if necessary, adjust the interrupter spacing until this signal is not erratic. Final fine tuning (which probably is not really necessary) is done by observing the "either" pulses with an oscilloscope while the knob is spun.

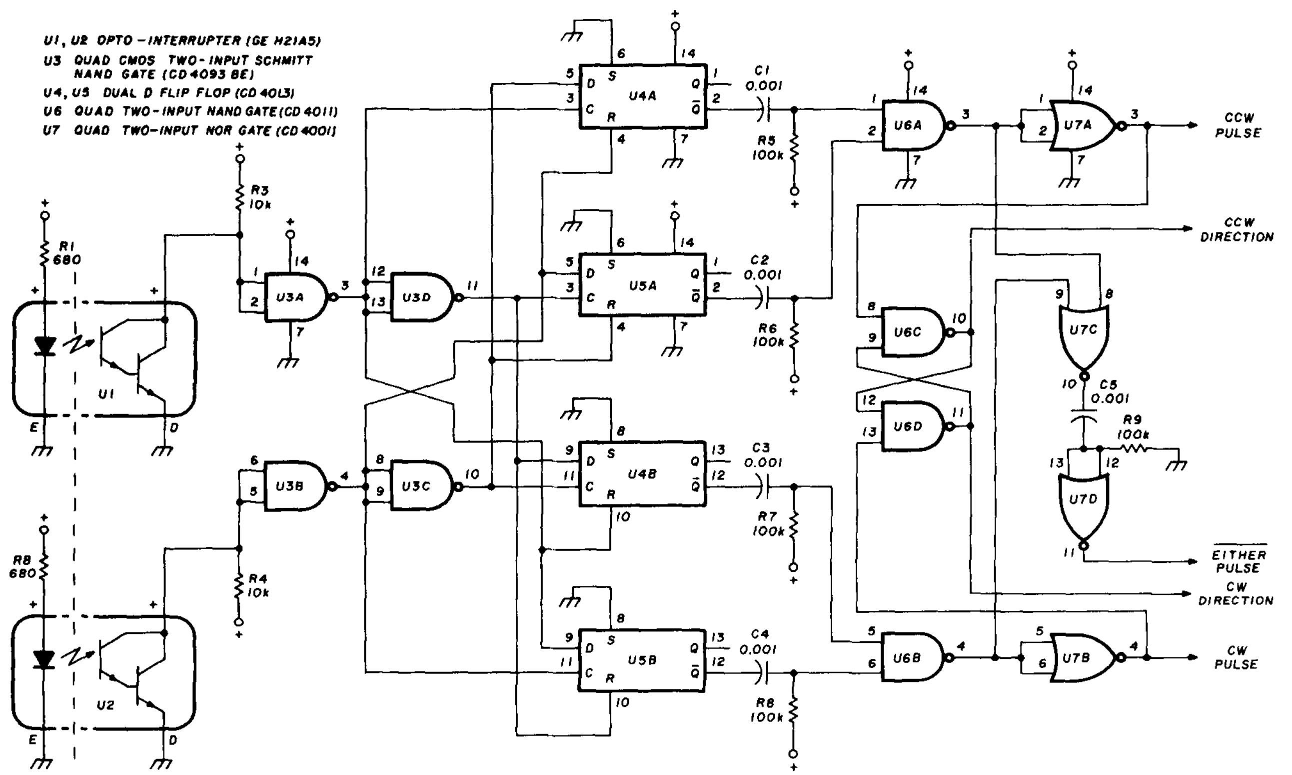

fig. 3. Schematic of the electronic processing circuit. GE H21A5 interrupters are used, but TI type TIXL45 and Monsanto type MCA8 should work as well. Output transistor load resistors, R1-R4, are for 5-volt operation. They should be increased proportionally if the circuit operating voltage is made higher.

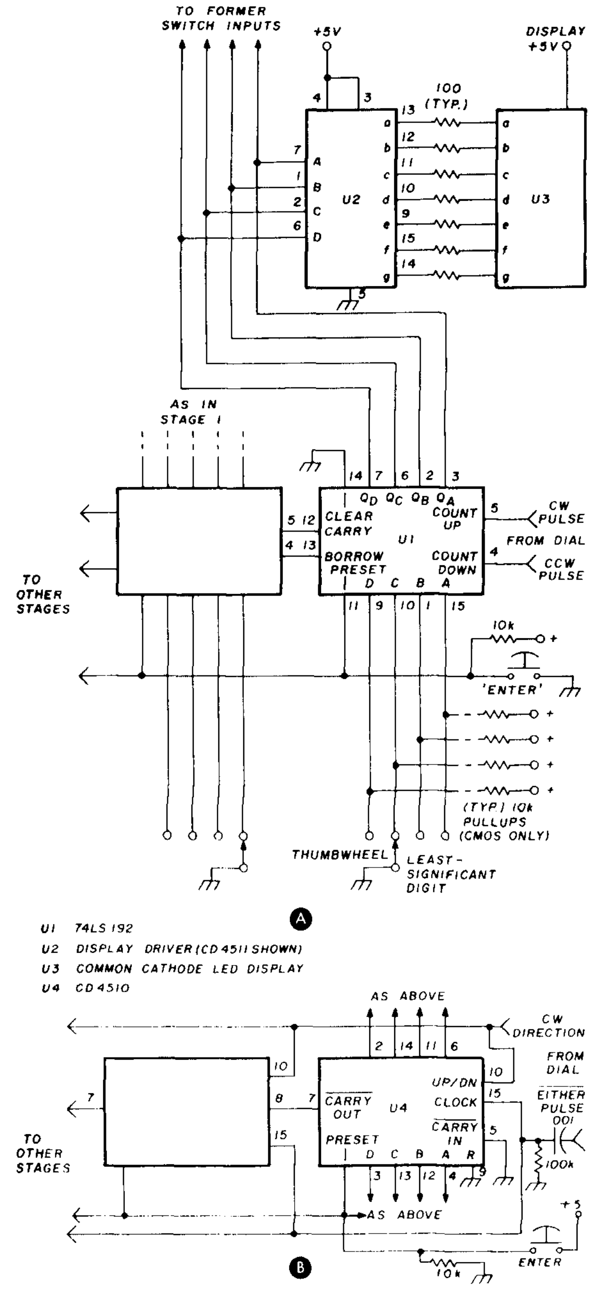

fig. 4. Examples of up/down counters that can be driven by the encoder. (A) shows an up/down counter with preset entry and an optional LED readout. The 74192 devices are highly recommended because of pinout compatibility with TTL and CMOS units. In (B) an up/down counter using a type CD 4510BE is shown, which is also compatible with the encoder. Standard IC power connections not shown.

The interrupter spacing should be tweaked until these pulses are evenly spaced.

applications

The simplest use of this dial system is as a supplement to existing pushbutton-tuned systems that already have an up/down counter. In this case it will be necessary only to merge the signals from the internal pulse generator with those from the dial mechanism. It may be possible to do this by simply adding a couple of diodes, although I have no direct experience with such a modification.

up/down counters

Most applications[2] will require the addition of an up/down counter. The original thumbwheel or rotary switch wires are routed to the counter, and a pushbutton is added to initialize the counter to the settings on the switch dials. After that, tuning is with the shaft encoder.

Fig. 4A shows a suitable up/down counter with preset entry and an optional LED readout, if the transceiver does not have one. The circuit uses 74192 up/down counters, which are highly recommended because they are available as pin-compatible TTL (74192), low-power TTL (74L192) low-power Shottky TTL (74LS192), and CMOS (74C192). The logic type used should be the same as that in the synthesizer. The output of the dial electronics is compatible with all types.

Some other counters, such as the CMOS 4510 and TTL 74168, use the direction and clock control signals as shown in **fig. 4B**. This type of counter often presents multiple control signal and/or clock loads to the encoder, so buffers will be needed when using the encoder with non-CMOS units.

additional remarks

I've been using the prototype encoder to tune a 5-5.5 MHz frequency-lock loop in an experimental 80 and 20 meter receiver. I've tried 100-, 40-, 20-, and 10-Hz tuning increments. To me, 100-Hz steps are intolerably coarse for CW work, 40-Hz steps are noticeable, but 20-Hz steps seem smooth enough. With 10-Hz increments, the 100-mark disk gives a leisurely 2-kHz/revolution tuning rate. Although the knob is easily spun for fast tuning (the only drag is in the panel bearing), pushbuttons for faster tuning would be nice. Those readers who use 100-Hz tuning increments should use the 50-mark disk; if this gives tuning which is too fast, disable half the pulses by disconnecting C2 and C4.

I'd be pleased to correspond with anyone considering use of the encoder and would appreciate reports on successful applications. I'd be most grateful for an SASE with all correspondence.

references

1. Lester Earnshaw, "Basics of the Digital VFO — A Tunable Synthesizer," *ham radio*, November, 1978, page 18.

2. Examples of equipment that could use the encoder are described by Earnshaw (*op. cit.*) and by Raymond C. Pettit, W7GHM, "Frequency Synthesized Local-Oscillator System for the High-Frequency Amateur Bands," *ham radio*, October, 1978, pages 60-65.

ham radio

Yagi antenna design: optimizing performance

Considerations for optimizing element length and position for maximum forward gain and front-to-back ratio

Over the past two months I have explored the properties of *simplistic* Yagi-Uda antennas, i.e., antennas of a given boom length but having uniformly spaced elements one of which is a reflector, one of which a driver, and with director parasites all of uniform length.[1,2] In real life, however, one is not restricted to the simplistic design; it is therefore interesting to examine a number of departures from the simplistic design to see if there are ways to further improve performance; in this and succeeding articles I shall attempt to explore a few ideas in a systematic way. It will soon be apparent that some of the departures from the simplistic design produce only subtle changes in performance, while others are major departures which produce significant changes in performance. It is fortunate that the accuracy inherent in computation can show very subtle changes; these changes, though small, can usually be trusted although the absolute accuracy of the model on which computation is based may not be better than a few per cent.

Departures from the simplistic design may be accomplished in a number of ways, but primarily by allowing the *lengths* of the directors (hence their resonant frequencies) to vary and by allowing the *placement* of the element(s) on the boom to change. Additionally, for a given boom and a given total number of elements, the number of reflectors (and hence directors) can be changed. This is a much more drastic change, and produces more pronounced performance variations. These changes will be analyzed in this and subsequent articles. Only free-space performance will be investigated at this time.

I shall start with a "good" simplistic design (6-element Yagi on a boom $0.75\ \lambda_o$ long), but will first change the lengths of all parasites to bring the center frequency of the desired 4 per cent band of maximum gain to $F = 1.0$. **Fig. 1** shows the main properties of this test antenna and the required element lengths. Note that the position of maximum F/B ratio is somewhat lower ($F = 0.988$) than the gain band center ($F = 1.0$); **fig. 2** shows the pattern of this antenna at band center ($F = 1.0$).

There are two visible nulls, the first one ($K = 1$) occurs at 87° and the second ($K = 2$) at 144°. The second null can be identified with the peak in the F/B ratio occurring at $F = 0.988$; at this lower frequency the null ($K = 2$) moves out to 180°. This antenna can be operated at best gain over the band ($F = 1.0$) and compromise F/B (17 dB at $F = 1.0$), or operated at the frequency of best F/B ($F = 0.988$, $F/B = 38$ dB) and somewhat compromise the gain available. In either case this Yagi-Uda design seems to be a good one and I shall use it as a test case around which certain departures from simplistic design can take place.

Since we are interested in *very* subtle changes I show in **table 1** the detailed performance in the region of chief interest, accurately calculated for this antenna over the frequency range from $F = 0.970$ to $F = 1.030$. The driving point reactance at a given frequency is somewhat arbitrary; it can be easily shifted or offset by changing the length of the driven element. The length of the driven element, however, remains fixed (free space resonance = 1.0) throughout this series of explorations so that reactance changes can be properly sensed.

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table 1. Computed performance characteristics of the 6-element Yagi over the frequency range from $F = 0.970$ to $F = 1.030$.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.058	12.928	24.677	−38.229
0.972	10.145	14.043	24.454	−35.512
0.974	10.225	15.279	24.190	−32.775
0.976	10.299	16.676	23.889	−30.006
0.978	10.369	18.298	23.559	−27.199
0.980	10.434	20.245	23.206	−24.347
0.982	10.495	22.700	22.838	−21.443
0.984	10.554	26.040	22.463	−18.483
0.986	10.608	31.184	22.088	−15.464
0.988	10.659	38.034	21.723	−12.381
0.990	10.706	31.810	21.375	−9.232
0.992	10.748	26.459	21.055	−6.016
0.994	10.785	23.044	20.770	−2.732
0.996	10.816	20.570	20.531	0.623
0.998	10.841	18.634	20.346	4.050
1.000	10.857	17.045	20.227	7.548
1.002	10.864	15.693	20.186	11.131
1.004	10.861	14.520	20.235	14.789
1.006	10.847	13.485	20.390	18.525
1.008	10.821	12.562	20.668	22.339
1.010	10.782	11.731	21.090	26.232
1.012	10.730	10.979	21.679	30.206
1.014	10.665	10.296	22.468	34.262
1.016	10.586	9.675	23.496	38.398
1.018	10.495	9.112	24.814	42.610
1.020	10.393	8.603	26.489	46.889
1.022	10.281	8.148	28.611	51.214
1.024	10.162	7.747	31.301	55.542
1.026	10.038	7.401	34.722	59.793
1.028	9.913	7.113	39.097	63.809
1.030	9.790	6.889	44.709	67.290

If we keep the average value of the length(s) of the directors constant we can explore "linear" length tapers and "parabolic" tapers. I will define a linear taper as a uniform linear free-space resonant frequency progression from $D1$ to $D4$ keeping the *average* resonant frequency constant. In other words, the director resonant frequencies are linearly related to director position measured from the center of the director assembly (to keep the average value constant). Similarly, a "parabolic" taper is one in which the directors' free-space resonant frequencies are proportional to the square of the distance from the center of the director assembly, with the further condition that the average value of director resonance is held constant. If we define the change in resonant frequency of $D1$ as Δ, then with a total of four directors all other director resonances will change:

Element	Linear Taper	Parabolic Taper
Director 1	$+\Delta$	$+\Delta$
Director 2	$+\Delta/3$	$-\Delta$
Director 3	$-\Delta/3$	$-\Delta$
Director 4	$-\Delta$	$+\Delta$

The degree or magnitude of taper is fixed by the size of Δ; moreover, Δ can be chosen either as an increase or decrease in resonant frequency.

I have selected and investigated six taper schedules which are delineated in **table 2**; as indicated, antenna performance in the critical region of interest are shown in **tables 3** to **8**. These results are to be compared with the simplistic design shown in **table 1**.

Table 3 shows the results for a "−2 per cent" linear taper and it is obvious that all performance parameters are virtually unchanged! **Table 4** shows the results for a "−4 per cent" linear taper, and even in this rather extreme case performance is almost totally unchanged in the central frequency region! Remember that for this case the free-space resonant frequency of $D1$ has dropped from *1.06* to *1.02* and it is easy to see that performance deteriorates at higher frequencies ($D1$ no longer behaves like a director). Nevertheless, it is remarkable that the linear taper — even of this magnitude — has virtually *no* effect on the performance of the Yagi-Uda antenna!

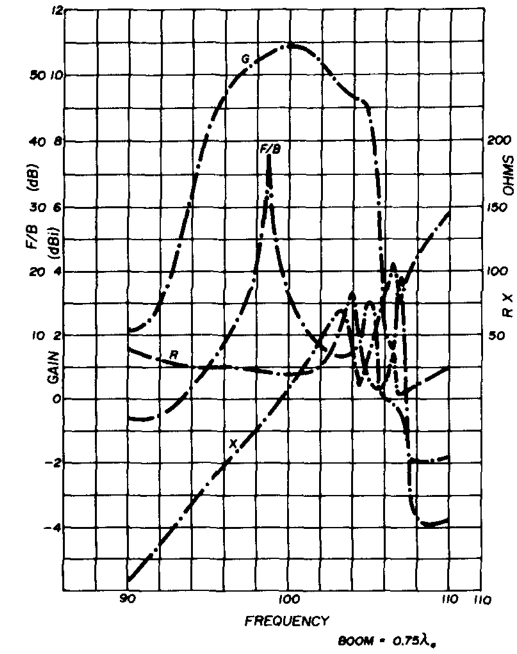
Table 5 shows results for a −2 per cent parabolic taper; **tables 6**, **7**, and **8** show results for parabolic tapers of −1, +1, and +2 per cent, respectively. In comparing these tables with the standard simplistic Yagi-Uda antenna we again see that truly minimal changes are made to the chief performance indices. The variations in maximum F/B ratios are not of great significance; I shall come back to this point later.

Analogous to the *length* (and corresponding resonant frequency) taper variations I have also investigated element *placement* variations along the boom. Again it is possible to vary the space intervals between elements linearly and pseudo parabolically. **Table 9** shows element positions for six schedules I have investigated and the individual results are shown in **tables 10** through **15**.

Table 10 shows the results where elements are crowded towards $D4$. Note that truly large changes in placement have been made! Similarly, **table 13** shows the results where elements are severely crowded towards the reflector. **Tables 11** and **12** are intermediate schedules; **tables 14** and **15** show results where end spaces are (relatively) increased.

table 2. Schedule of director lengths (λ_o) for the six 6-element Yagi-Uda performance characteristics listed in table 3 through table 8.

table	taper type	Dir 1 (λ_o)	Dir 2 (λ_o)	Dir 3 (λ_o)	Dir 4 (λ_o)	
3	Linear	0.46341	0.45719	0.45263	0.44524	−2%
4	Linear	0.47306	0.46028	0.44817	0.43667	−4%
5	Parabolic	0.46341	0.44524	0.44524	0.46341	−2%
6	Parabolic	0.45873	0.44964	0.44964	0.45873	−1%
7	Parabolic	0.44964	0.45873	0.45873	0.44964	+1%
8	Parabolic	0.44524	0.46341	0.46341	0.44524	+2%



element	length (λ_o)	free space resonance
Reflector	0.50195	0.96
Driven Element	0.48167	1.00
Director 1	0.45414	1.06
Director 2	0.45414	1.06
Director 3	0.45414	1.06
Director 4	0.45414	1.06

fig. 1. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$. Note that the position of maximum F/B ratio ($F = 0.988$) is somewhat lower than the gain band center ($F = 1.0$). The pattern of this antenna at the band center is plotted in fig. 2.

These tables all show that these placement variations have only a *very* minor effect on directivity or gain, and while the maximum F/B ratio is somewhat affected (generally adversely), we shall soon see that it may not be very significant.

Up to this point we have looked at taper schedules which are *linear* and *parabolic* and which also involve director *length*, or resonant frequency, and element *placement* along the boom. It is truly remarkable that *all* of these schedules produce minimal changes in antenna performance; it is therefore plausible that *combinations* of these schedules will also produce minimal performance variations. This leads to the conclusion that the original simplistic design (dimensions listed in **fig. 1**) is just about as good as any. No real improvement on gain can be expected by any new tricky design; as far as F/B ratio is concerned, it will soon be apparent that you can "tune up" the maximum F/B ratio starting with almost any of these schedules.

A summary of raw performance of all of these cases is shown in **table 16**, where, in addition to data already shown in the previous tables, information on pattern (not explicitly shown here) has been added. This table shows that all cases produce about the same gain; the very small variations are due to the effective "illumination" pattern of the boom aperture.[2] The F/B ratio at central gain frequency ($F \approx 1.0$) varies somewhat, but a very slight change in operating frequency would easily make them all comparable. The frequency position of maximum F/B ratio and the angle of the second null at $F = 1.0$ are related. *Lower* frequencies of maximum F/B *should* correspond (at central frequency) to *longer* effective boom illuminated apertures, thus corresponding to *lower* null angles at $F = 1.0$ and somewhat *higher* gain. An examination of **table 16** shows all of these quantities to be well correlated; it appears therefore that all results are understood and self-consistent.

From all of this information it is reasonable to draw a general conclusion that the simplistic Yagi design gives about as much gain as any other design off the

table 3. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director lengths tapered linearly at −2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.074	12.997	23.856	−39.110
0.972	10.160	14.083	23.636	−36.425
0.974	10.238	15.280	23.379	−33.725
0.976	10.309	16.622	23.089	−31.003
0.978	10.375	18.164	22.771	−28.250
0.980	10.435	19.993	22.433	−25.462
0.982	10.492	22.262	22.082	−22.634
0.984	10.544	25.284	21.723	−19.762
0.986	10.592	29.878	21.365	−16.843
0.988	10.635	39.987	21.016	−13.876
0.990	10.674	38.240	20.684	−10.859
0.992	10.708	29.358	20.375	−7.794
0.994	10.736	25.064	20.099	−4.679
0.996	10.758	22.209	19.862	−1.516
0.998	10.774	20.068	19.673	1.694
1.000	10.781	18.356	19.542	4.948
1.002	10.780	16.927	19.478	8.251
1.004	10.770	15.706	19.491	11.594
1.006	10.750	14.643	19.593	14.972
1.008	10.720	13.705	19.796	18.381
1.010	10.680	12.872	20.115	21.812
1.012	10.629	12.128	20.567	25.259
1.014	10.568	11.462	21.173	28.707
1.016	10.497	10.869	21.960	32.141
1.018	10.418	10.343	22.957	35.535
1.020	10.331	9.884	24.203	38.851
1.022	10.240	9.492	25.742	42.033
1.024	10.145	9.171	27.623	44.991
1.026	10.049	8.925	29.894	47.585
1.028	9.954	8.764	32.572	49.596
1.030	9.861	8.701	35.588	50.696

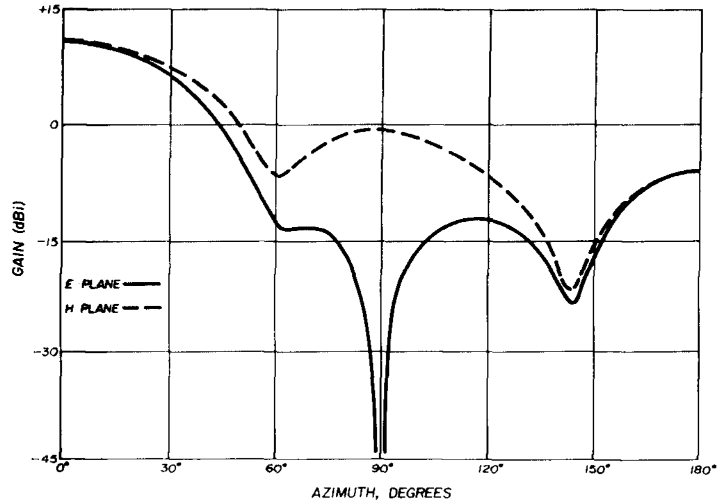


fig. 2. Radiation pattern of a 6-element Yagi beam on a $0.75\,\lambda_o$ boom (only 180 degrees are shown). The pattern nulls occur at 87 and 144 degrees; the second null is correlated with the peak of F/B ratio at $F = 0.988$.

same boom length. Tapering element lengths or element position intervals along the boom is of no apparent value. The characteristic of the directors which *is* important is the *average* length (or average free-space resonant frequency). But this conclusion has been demonstrated only for a boom length of $0.75\ \lambda_o$; we must be careful not to generalize too much. Recall that the NBS data (see **fig. 7** of the NBS report[3]) suggested that for booms longer than one wavelength some improvement in gain over simplistic Yagi-Uda performance could be obtained with particular director length schedules. My calculations support the NBS result; nevertheless, for boom lengths shorter than one wavelength the simplistic design is as good as any!

It can be seen from **table 16** that the best "null" ($K = 2$) positions give quite different values of maximum F/B. Indeed, *good* nulls or correspondingly *high* values of F/B must be viewed as accidental vectorial cancellations in the reverse or back direction; such good cancellations will not generally occur with any arbitrary boom illumination. Note that the various cases shown in **table 16** display maximum F/B ratios ranging from 22 to 40 dB. It is an interesting exercise to see if there is some way to significantly enhance the maximum F/B ratio by some variational procedure.

Let us start with the simplistic Yagi-Uda design (**fig. 1**) and vary the *position* of, say $D3$, along the boom. We now know that small variations in position will not significantly affect gain, but vectorial cancellation effects in the back direction can be expected to be significant. If we can find a position for $D3$ which maximizes the F/B ratio, its vectorial contribution in the back direction should be approximately out of phase with the residue from all other elements. At this point some other element (say $D1$) can be positioned for a new (still higher) maximum F/B ratio; after this is done $D3$ can be readjusted again for a new maximum F/B ratio, etc. By iterating the two adjustments it should be possible to continuously improve F/B ratios, presumably to as high a value as desired! With such an iteration procedure it is desirable to start with a fairly good value of F/B so that only small variations in element position can have a significant effect.

I have carried out such an iteration (using $D3$ and $D1$) for the simplistic Yagi-Uda design and have arrived at the following positions:

element	$X(\lambda_o)$
Reflector	0.000
Driven Element	0.150
Director 1	0.28967
Director 2	0.450
Director 3	0.58945
Director 4	0.750

table 4. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director lengths tapered linearly at −4 per cent (director lengths shown in *table 2*).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.064	13.741	20.930	−40.263
0.972	10.152	14.915	20.598	−37.496
0.974	10.232	16.225	20.240	−34.709
0.976	10.305	17.715	19.860	−31.898
0.978	10.371	19.458	19.464	−29.058
0.980	10.432	21.568	19.058	−26.185
0.982	10.486	24.235	18.648	−23.276
0.984	10.535	27.758	18.241	−20.331
0.986	10.577	31.908	17.841	−17.348
0.988	10.613	31.851	17.456	−14.328
0.990	10.642	27.723	17.092	−11.271
0.992	10.663	24.260	16.755	−8.181
0.994	10.676	21.650	16.450	−5.058
0.996	10.679	19.600	16.183	−1.907
0.998	10.672	17.921	15.959	1.268
1.000	10.654	16.504	15.785	4.463
1.002	10.624	15.276	15.666	7.677
1.004	10.580	14.197	15.607	10.897
1.006	10.523	13.237	15.612	14.115
1.008	10.452	12.374	15.687	17.320
1.010	10.365	11.594	15.835	20.498
1.012	10.264	10.886	16.058	23.631
1.014	10.147	10.242	16.356	26.698
1.016	10.013	9.653	16.721	29.669
1.018	9.861	9.115	17.140	32.511
1.020	9.689	8.621	17.578	35.181
1.022	9.489	8.159	17.977	37.637
1.024	9.252	7.713	18.235	39.849
1.026	8.955	7.252	18.205	41.835
1.028	8.562	6.722	17.715	43.715
1.030	8.012	6.025	16.654	45.750

table 5. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_0$, director lengths tapered parabolically at −2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.013	16.381	19.732	−42.299
0.972	10.112	18.110	19.184	−39.363
0.974	10.207	20.196	18.625	−36.379
0.976	10.298	22.808	18.063	−33.345
0.978	10.386	26.155	17.504	−30.258
0.980	10.471	29.690	16.954	−27.117
0.982	10.553	29.326	16.421	−23.921
0.984	10.632	25.681	15.910	−20.669
0.986	10.706	22.446	15.428	−17.361
0.988	10.776	19.930	14.980	−13.998
0.990	10.840	17.920	14.572	−10.578
0.992	10.897	16.257	14.211	−7.103
0.994	10.946	14.841	13.903	−3.573
0.996	10.986	13.610	13.653	0.013
0.998	11.014	12.522	13.469	3.655
1.000	11.030	11.547	13.359	7.354
1.002	11.033	10.662	13.332	11.124
1.004	11.021	9.856	13.398	14.956
1.006	10.993	9.120	13.570	18.854
1.008	10.950	8.444	13.864	22.822
1.010	10.892	7.824	14.299	26.865
1.012	10.820	7.256	14.903	30.991
1.014	10.735	6.737	15.709	35.208
1.016	10.641	6.267	16.768	39.527
1.018	10.540	5.846	18.148	43.959
1.020	10.436	5.475	19.953	48.511
1.022	10.336	5.157	22.337	53.182
1.024	10.243	4.897	25.544	57.936
1.026	10.163	4.702	29.962	62.653
1.028	10.104	4.583	36.221	66.971
1.030	10.070	4.555	45.262	69.868

This design, optimized for maximum F/B ratio (at $F = 0.990$) produces the performance displayed in **table 17**. A careful comparison of this table with the original simplistic model shows virtually identical performance in all respects *except* that the F/B maximum has gone up from an excellent 38 dB to an astounding 98 dB! Even this high value is not a real limit; it is limited only by the number of iterations which were made.

Notice that these astronomical values of F/B are of *no practical significance*. It occurs at essentially a single frequency and its effective bandwidth becomes vanishingly small. Moreover, extremely small variations in the Yagi-Uda dimensions will upset the cancellation; in practice you could not likely construct a mechanically satisfactory, fully optimized Yagi-Uda antenna. Nevertheless, the mathematical iteration shows that it is possible, in principle, to obtain (at a single frequency) an arbitrarily high F/B ratio. It is likely that there are a large number of potential solutions involving iterations with other elements. Furthermore, we now know that the variations in F/B maxima shown in **table 16** result from the particular illumination chosen, and it is very likely that minor element placement variations could make an arbitrarily high F/B ratio design starting from *any* of the cases shown.

To understand this iteration procedure it is helpful to show the vectorial contributions of each element to the forward and back waves. The (current) contribution from a given element will be a vector whose magnitude is the magnitude of element current and whose phase consists of two parts. The first part is the actual (time) phase of the element current referred to some time origin (say the driver current) and the second is the (space) phase change due to the element position referred to some space origin (say at $X = 0$ along the boom). Note that this second part changes sign in going from a forward wave to a reverse wave! **Fig. 3** shows these (current) vectorial contributions at $F = 0.988$ for the original simplistic Yagi-Uda design (**fig. 1**) to both forward and reverse waves.

table 6. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_0$, director lengths tapered parabolically at −1 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.040	14.338	22.533	−40.493
0.972	10.132	15.667	22.133	−37.696
0.974	10.219	17.186	21.704	−34.863
0.976	10.300	18.973	21.253	−31.986
0.978	10.378	21.151	20.787	−29.060
0.980	10.453	23.931	20.313	−26.082
0.982	10.524	27.615	19.839	−23.048
0.984	10.592	31.590	19.372	−19.956
0.986	10.656	30.418	18.920	−16.804
0.988	10.717	26.254	18.490	−13.590
0.990	10.773	22.942	18.089	−10.314
0.992	10.824	20.432	17.725	−6.975
0.994	10.870	18.445	17.406	−3.572
0.996	10.908	16.807	17.138	−0.105
0.998	10.938	15.417	16.932	3.427
1.000	10.959	14.209	16.794	7.024
1.002	10.970	13.140	16.736	10.700
1.004	10.969	12.183	16.770	14.446
1.006	10.956	11.321	16.909	18.265
1.008	10.930	10.538	17.168	22.160
1.010	10.890	9.825	17.568	26.133
1.012	10.838	9.173	18.134	30.191
1.014	10.772	8.578	18.896	34.338
1.016	10.694	8.035	19.899	38.580
1.018	10.607	7.543	21.198	42.922
1.020	10.512	7.101	22.875	47.367
1.022	10.413	6.710	25.041	51.906
1.024	10.312	6.372	27.863	56.516
1.026	10.215	6.089	31.585	61.122
1.028	10.126	5.869	36.575	65.547
1.030	10.051	5.720	43.373	69.362

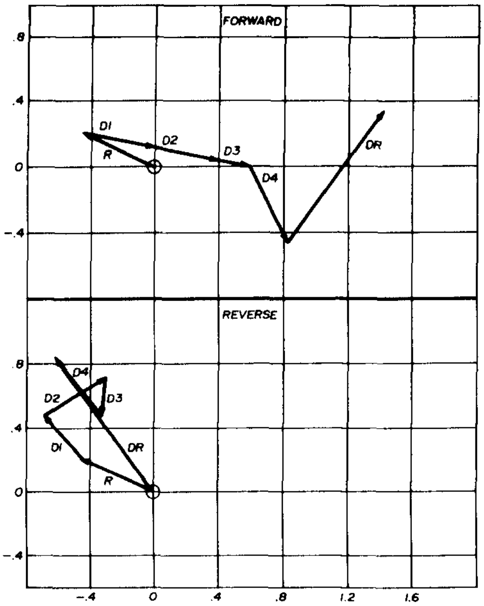


fig. 3. Current vectorial contributions at $F = 0.988$ for 6-element Yagi ($boom = 0.75\lambda_o$), forward and reverse waves.

Note that for the forward wave the individual element contributions do not all fully reinforce the forward wave; in fact, the contribution from the reflector is even *negative* with respect to the final total (current) vector! This curious result is typical of all Yagi-Uda arrays. Note that the contributions to the back or reverse wave, in total, nearly cancel out, leaving only a small residue which accounts for the 38 dB F/B ratio. Now it is easy to see conceptually what happens in the iterative procedure to reduce the backwave residual.

If you look at the reverse wave vector plot, it is easy to imagine that as $D3$ is moved along the boom, the $D3$ vector rotates around *its* origin. The backwave residual will then be changed along an axis at right angles to the $D3$ vector and can be minimized by the $D3$ position. After this is done *another* element, say $D1$, can be moved along the boom; its vector contribution is at a different angle and can therefore reduce the residual still further. Thus, in principle, iterative motions of two elements whose backwave vectors contribute at different angles can ultimately reduce the backwave residual to as low a value as desired.

The iterative convergence will be most rapid if the two element vectors are orthogonal; nevertheless, it can converge adequately for many element combinations. Of course, this conceptual picture is oversimplified; as any element is moved on the boom not only does its vector rotate, but *all* element currents and phases readjust somewhat. However, these readjustments are usually minor and in practice cause little difficulty as long as you start with a reasonably small residual as shown in **fig. 3**.

Fig. 4 shows the vectorial contributions for the optimized Yagi-Uda beam. Note that the element contributions are only slightly modified in the optimization procedure.

At this point I must issue a warning. Recall that the mathematical model being used in these computations involves certain approximations. These approximations make relatively little difference in the calculations for forward gain, but they become crucial in calculations involving vectorial cancellation or closure for back radiation. Thus the explicitly calculated *positions* for and *magnitude* of a very high F/B ratio are not to be trusted. Nevertheless, the *general* behavior is still valid. The real Yagi-Uda can still be made to have a high F/B ratio, just as our mathematical model shows, but it may occur at a slightly different frequency and it may require slightly different positions for $D3$ and $D1$ in the final optimization.

table 7. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director lengths tapered parabolically at +1 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.066	11.872	26.168	−35.821
0.972	10.150	12.855	26.121	−33.131
0.974	10.225	13.922	26.025	−30.437
0.976	10.293	15.096	25.884	−27.728
0.978	10.355	16.414	25.702	−24.993
0.980	10.412	17.927	25.485	−22.222
0.982	10.465	19.720	25.240	−19.407
0.984	10.514	21.939	24.975	−16.542
0.986	10.558	24.879	24.699	−13.619
0.988	10.599	29.294	24.419	−10.635
0.990	10.635	38.438	24.147	−7.584
0.992	10.667	38.763	23.891	−4.463
0.994	10.693	29.441	23.663	−1.269
0.996	10.714	25.008	23.472	2.000
0.998	10.727	22.079	23.330	5.345
1.000	10.732	19.886	23.248	8.768
1.002	10.728	18.127	23.244	12.280
1.004	10.714	16.660	23.328	15.873
1.006	10.688	15.401	23.519	19.547
1.008	10.650	14.301	23.833	23.302
1.010	10.597	13.325	24.293	27.136
1.012	10.530	12.450	24.923	31.048
1.014	10.448	11.661	25.755	35.033
1.016	10.350	10.946	26.823	39.083
1.018	10.237	10.296	28.174	43.188
1.020	10.108	9.708	29.862	47.325
1.022	9.965	9.177	31.956	51.463
1.024	9.808	8.701	34.542	55.548
1.026	9.640	8.280	37.723	59.490
1.028	9.462	7.916	41.618	63.148
1.030	9.277	7.613	46.355	66.289

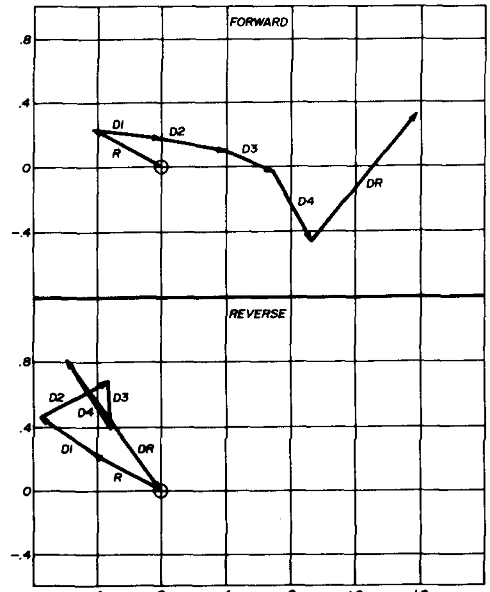


fig. 4. Current vectorial contributions for the optimized 6-element Yagi at $F = 0.990$.

optimum design

We now have the necessary tools with which to design truly excellent Yagi-Uda antennas. We start first from a knowledge that the boom length should be approximately an odd number of quarter wavelengths for an initial simplistic design; we have seen that such a boom length promotes an inherently high F/B ratio at a frequency near the center of the best gain band; boom length also determines ultimate gain. After the boom length is chosen a resonant frequency schedule is chosen (see appropriate figures from the simplistic Yagi-Uda articles[1,2]) for reflector and director(s); a preliminary calculation is then made to accurately determine the frequency of maximum F/B ratio which will not necessarily correspond with the frequency at the center of the best gain portion. The *useful* band, however, is now to be centered around the F/B point; it is necessary to insure that there is enough gain bandwidth left for the intended purpose.

Remember that the overall gain bandwidth is basically controlled by the resonant frequency schedule of the parasites. This bandwidth should not be larger than necessary, because gain is compromised somewhat as the bandwidth increases.

Now translate the frequency ($F1$) of best F/B to $F = 1.0$ by multiplying all parasite lengths by $F1$; a new preliminary calculation, possibly iterated once more, will insure that the best F/B ratio is exactly $F = 1.0$. Next, alternately vary the X position of $D3$

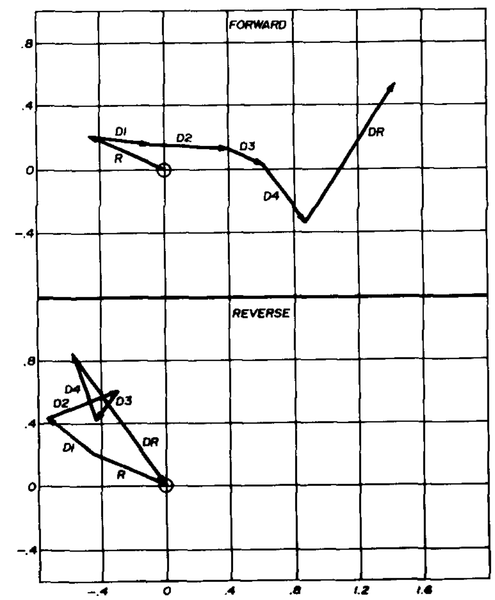


fig. 5. Current vectorial contributions of the $D3$-$D1$ optimized Yagi; forward wave, above, and reverse wave, below.

table 8. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director lengths tapered parabolically at +2 per cent (director lengths shown in table 2).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.065	11.048	27.104	−33.539
0.972	10.145	11.945	27.202	−30.843
0.974	10.215	12.904	27.249	−28.160
0.976	10.277	13.942	27.243	−25.474
0.978	10.332	15.081	27.187	−22.775
0.980	10.382	16.353	27.086	−20.051
0.982	10.426	17.805	26.944	−17.289
0.984	10.465	19.507	26.770	−14.482
0.986	10.500	21.575	26.572	−11.621
0.988	10.530	24.216	26.358	−8.697
0.990	10.555	27.812	26.138	−5.704
0.992	10.575	32.685	25.924	−2.637
0.994	10.588	33.646	25.725	0.508
0.996	10.594	28.730	25.555	3.736
0.998	10.592	24.844	25.424	7.050
1.000	10.581	22.026	25.348	10.453
1.002	10.559	19.844	25.342	13.954
1.004	10.525	18.071	25.421	17.548
1.006	10.476	16.576	25.602	21.236
1.008	10.412	15.281	25.902	25.017
1.010	10.330	14.139	26.345	28.891
1.012	10.228	13.116	26.954	32.855
1.014	10.106	12.189	27.756	36.904
1.016	9.962	11.341	28.786	41.032
1.018	9.794	10.559	30.080	45.228
1.020	9.602	9.835	31.684	49.473
1.022	9.384	9.161	33.649	53.741
1.024	9.140	8.530	36.034	57.993
1.026	8.870	7.938	38.907	62.169
1.028	8.570	7.378	42.337	66.185
1.030	8.241	6.846	46.394	69.921

table 9. Schedule of director placement on boom (λ_o) for the six 6-element Yagi-Uda performance characteristics listed in table 10 through table 15.

table	taper type	Refl	DR	D1	D2	D3	D4
		element position on boom (λ_o)					
10	Linear	0	0.200	0.3750	0.5250	0.650	0.750
11	Linear	0	0.175	0.3375	0.4875	0.625	0.750
12	Linear	0	0.125	0.2625	0.4125	0.575	0.750
13	Linear	0	0.100	0.2250	0.3750	0.500	0.750
14	Parabolic	0	0.200	0.3167	0.4333	0.550	0.750
15	Parabolic	0	0.175	0.3083	0.4417	0.575	0.750

and then $D1$ to get larger and larger values of F/B at $F = 1.0$ until the value is sufficiently high.

design example

An example will illustrate this design procedure. For example, let's choose a boom length of $0.780\ \lambda_o$. From an inspection of the results of our test Yagi-Uda simplistic design (**fig. 1**) we can probably use the same parasite resonant frequency schedule and still obtain an adequate ultimate gain bandwidth performance. Listed below are the initial element positions along the boom:

element	X (λ_o)	initial length (λ_o)	intermediate length (λ_o)
Reflector	0.000	0.50195	0.49343
Driven El	0.156	0.48167	0.48167
Director 1	0.312	0.45414	0.44643
Director 2	0.468	0.45414	0.44643
Director 3	0.624	0.45414	0.44643
Director 4	0.780	0.45414	0.44643

table 10. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director spacing tapered linearly, large positive interval, directors crowded toward $D4$.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.872	11.767	43.699	−38.627
0.972	9.912	12.501	42.699	−36.912
0.974	9.952	13.285	41.580	−35.082
0.976	9.992	14.127	40.362	−33.122
0.978	10.034	15.036	39.064	−31.025
0.980	10.078	16.025	37.706	−28.783
0.982	10.123	17.099	36.306	−26.392
0.984	10.171	18.259	34.883	−23.851
0.986	10.221	19.481	33.453	−21.161
0.988	10.273	20.686	32.032	−18.323
0.990	10.326	21.689	30.634	−15.341
0.992	10.381	22.191	29.272	−12.219
0.994	10.435	21.951	27.960	−8.960
0.996	10.489	21.034	26.709	−5.571
0.998	10.540	19.734	25.528	−2.055
1.000	10.587	18.319	24.429	1.581
1.002	10.628	16.928	23.427	5.334
1.004	10.661	15.619	22.524	9.196
1.006	10.683	14.407	21.732	13.164
1.008	10.691	13.289	21.060	17.233
1.010	10.682	12.256	20.518	21.398
1.012	10.654	11.300	20.116	25.656
1.014	10.604	10.412	19.866	30.001
1.016	10.529	9.586	19.780	34.432
1.018	10.427	8.816	19.875	38.944
1.020	10.299	8.098	20.168	43.535
1.022	10.143	7.428	20.680	48.203
1.024	9.961	6.805	21.438	52.946
1.026	9.756	6.226	22.478	57.762
1.028	9.530	5.693	23.843	62.649
1.030	9.289	5.205	25.595	67.602

table 11. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director spacing tapered linearly as shown in table 9 (mild positive linear interval).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.039	12.781	33.045	−37.524
0.972	10.096	13.710	32.479	−35.253
0.974	10.151	14.723	31.839	−32.918
0.976	10.204	15.841	31.136	−30.510
0.978	10.256	17.094	30.381	−28.020
0.980	10.308	18.519	29.588	−25.443
0.982	10.359	20.164	28.767	−22.772
0.984	10.409	22.078	27.931	−20.005
0.986	10.459	24.247	27.093	−17.140
0.988	10.509	26.340	26.262	−14.177
0.990	10.557	27.212	25.450	−11.116
0.992	10.603	25.986	24.667	−7.956
0.994	10.646	23.746	23.923	−4.700
0.996	10.685	21.521	23.228	−1.350
0.998	10.719	19.567	22.592	2.094
1.000	10.747	17.882	22.023	5.629
1.002	10.767	16.414	21.536	9.264
1.004	10.777	15.123	21.136	12.988
1.006	10.775	13.973	20.836	16.799
1.008	10.760	12.938	20.648	20.697
1.010	10.729	11.998	20.584	24.681
1.012	10.682	11.140	20.659	28.749
1.014	10.617	10.353	20.893	32.903
1.016	10.534	9.629	21.306	37.142
1.018	10.432	8.961	21.925	41.466
1.020	10.312	8.347	22.785	45.875
1.022	10.174	7.783	23.929	50.368
1.024	10.022	7.268	25.415	54.941
1.026	9.857	6.802	27.321	59.584
1.028	9.684	6.386	29.750	64.273
1.030	9.505	6.023	32.855	68.965

Initial performance of this Yagi-Uda model is shown in **table 18**; the frequency for maximum *F/B* is $F = 0.984$. Shortening all elements by approximately this frequency factor yields the intermediate design also shown above. Performance for this intermediate design is shown in **table 19**. Note that since all lengths were not scaled (boom not scaled), this intermediate Yagi-Uda is not really quite the same as our starting model; the maximum *F/B* ratio has, in fact, fallen to 27 dB. However, this is of no concern; it is now time to iteratively vary *D3* and *D1* positions to "tune up" the *F/B* ratio. Alternatively, if our concept of optimization is correct, iterative variations of *D3* and *DR* could also tune up the *F/B* ratio. I have carried out both iterations and the resulting optimized Yagi parameters are as shown:

element	length (λ_o)	*D3-D1* opt X (λ_o)	*D3-DR* opt X(λ_o)
Reflector	0.49343	0.000	0.000
Driven El	0.48167	0.156	0.175595
Director 1	0.44643	0.291564	0.312
Director 2	0.44643	0.468	0.468
Director 3	0.44643	0.64075	0.6328873
Director 4	0.44643	0.780	0.780

Performance of this *D3-D1* optimized antenna is shown in **table 20**; it is nearly the same as that of the intermediate design (**table 19**) except that the *F/B* ratio at $F = 1.0$ has gone up from 27 dB to an astronomical 120 dB! Similarly, the performance for the *D3-DR* optimized antenna is shown in **table 21**. Again an astounding *F/B* ratio figure is achieved; moreover, the newer optimized beam performance is essentially identical with that of the first optimized model!

It is instructive to examine the final vector contributions to forward and reverse waves; **fig. 5** shows such a plot for the *D3-D1* optimized Yagi-Uda and **fig. 6** a similar plot for the *D3-DR* optimized model. Note that they look similar, differing only in minute details. Incidentally, it is noteworthy that the reverse plots show vectorial contributions going around the

table 12. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director spacing tapered linearly as shown in table 9 (mild negative linear interval).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.877	11.719	17.840	−41.204
0.972	10.024	12.962	17.848	−38.079
0.974	10.155	14.333	17.848	−34.966
0.976	10.271	15.879	17.838	−31.856
0.978	10.375	17.669	17.821	−28.741
0.980	10.468	19.823	17.800	−25.615
0.982	10.552	22.556	17.777	−22.470
0.984	10.627	26.310	17.758	−19.300
0.986	10.694	31.942	17.750	−16.100
0.988	10.753	34.618	17.757	−12.865
0.990	10.805	28.634	17.788	−9.590
0.992	10.849	24.386	17.851	−6.272
0.994	10.885	21.483	17.956	−2.905
0.996	10.913	19.314	18.111	0.512
0.998	10.932	17.592	18.328	3.984
1.000	10.942	16.170	18.622	7.513
1.002	10.943	14.958	19.011	11.116
1.004	10.934	13.909	19.511	14.783
1.006	10.914	12.989	20.145	18.516
1.008	10.884	12.173	20.939	22.317
1.010	10.843	11.445	21.930	26.185
1.012	10.793	10.795	23.158	30.117
1.014	10.733	10.213	24.682	34.104
1.016	10.664	9.694	26.572	38.128
1.018	10.588	9.236	28.926	42.152
1.020	10.507	8.836	31.868	46.112
1.022	10.422	8.495	35.562	49.886
1.024	10.336	8.216	40.211	53.254
1.026	10.252	8.001	46.034	55.818
1.028	10.171	7.857	53.178	56.886
1.030	10.098	7.795	61.476	55.329

table 13. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$, director spacing tapered linearly with the elements crowded toward the reflector (large negative interval taper).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.271	8.830	11.990	−47.108
0.972	9.569	10.190	12.078	−43.541
0.974	9.823	11.648	12.199	−40.009
0.976	10.040	13.240	12.349	−36.506
0.978	10.225	15.016	12.523	−33.027
0.980	10.383	17.045	12.720	−29.565
0.982	10.517	19.420	12.940	−26.112
0.984	10.631	22.216	13.185	−22.662
0.986	10.727	25.165	13.458	−19.209
0.988	10.807	26.553	13.764	−15.746
0.990	10.873	24.901	14.109	−12.267
0.992	10.925	22.296	14.501	−8.776
0.994	10.964	19.998	14.950	−5.237
0.996	10.992	18.123	15.468	−1.675
0.998	11.009	16.585	16.071	1.924
1.000	11.014	15.299	16.776	5.563
1.002	11.010	14.210	17.609	9.241
1.004	10.995	13.272	18.595	12.963
1.006	10.971	12.455	19.768	16.728
1.008	10.939	11.741	21.173	20.528
1.010	10.899	11.114	22.864	24.352
1.012	10.852	10.564	24.916	28.172
1.014	10.800	10.085	27.420	31.940
1.016	10.744	9.674	30.495	35.570
1.018	10.686	9.326	34.284	38.911
1.020	10.627	9.044	38.945	41.704
1.022	10.571	8.829	44.607	43.505
1.024	10.518	8.685	51.240	43.611
1.026	10.472	8.620	58.368	41.024
1.028	10.435	8.643	64.591	34.749
1.030	10.408	8.770	67.340	24.792

clock *twice* corresponding to the $K = 2$ null which we have constructed.

There is one final point worth mentioning. An examination of **tables 1, 20,** and **21** reveals that the frequency for the best F/B ratio is not generally quite the same as the frequency center of the gain bandwidth. It is offset by an amount which depends only on the boom length. This offset is of small importance as long as the gain bandwidth is large enough; it is nevertheless possible to empirically measure the offset frequency as a function of boom length. Let us fix the frequency of best F/B ratio as $F = 1.0$, and designate the frequency of (central) best gain (4 per cent BW) as F_G *(Offset frequency* $= F_G - 1.0)$; empirical results are shown in **fig. 7**.

Note that if the boom length is $0.63\ \lambda_o$ the offset disappears. For booms shorter than this value the offset is negative, and for booms longer than $0.63\ \lambda_o$ the offset is positive. But it is clearly possible to design a satisfactory Yagi over a considerable range of boom lengths without incurring an offset which is comparable to the bandwidth itself; it is only neces-

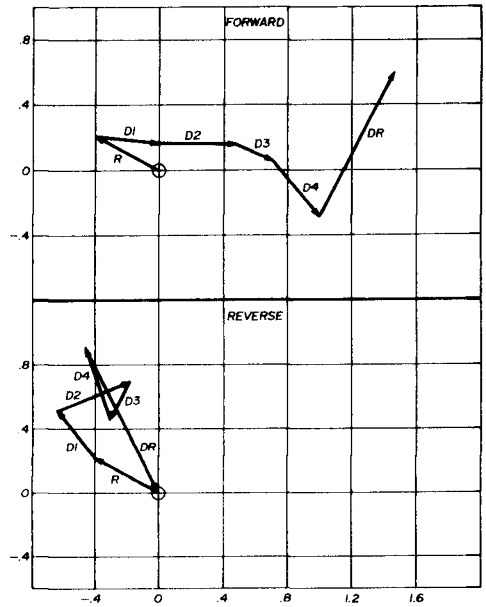


fig. 6. Current vector contributions for the $D3$-DR **optimized Yagi beam.**

table 14. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$**, director spacing tapered pseudo parabolically according to the schedule of table 9 (large positive interval).**

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.248	15.471	33.711	−34.346
0.972	10.301	16.499	33.381	−31.789
0.974	10.350	17.639	33.016	−29.188
0.976	10.398	18.923	32.623	−26.535
0.978	10.442	20.389	32.209	−23.924
0.980	10.485	22.087	31.780	−21.049
0.982	10.526	24.059	31.344	−18.204
0.984	10.564	26.253	30.909	−15.284
0.986	10.600	28.196	30.485	−12.285
0.988	10.634	28.611	30.079	−9.203
0.990	10.664	27.043	29.703	−6.034
0.992	10.691	24.764	29.366	−2.774
0.994	10.714	22.595	29.079	0.580
0.996	10.731	20.705	28.855	4.033
0.998	10.743	19.071	28.707	7.588
1.000	10.749	17.645	28.651	11.250
1.002	10.747	16.388	28.706	15.021
1.004	10.736	15.262	28.891	18.909
1.006	10.716	14.243	29.228	22.918
1.008	10.685	13.312	29.746	27.056
1.010	10.643	12.457	30.482	31.330
1.012	10.588	11.667	31.479	35.747
1.014	10.522	10.935	32.794	40.313
1.016	10.442	10.254	34.504	45.035
1.018	10.350	9.621	36.705	49.912
1.020	10.247	9.033	39.535	54.935
1.022	10.132	8.488	43.177	60.072
1.024	10.009	7.986	47.893	65.246
1.026	9.879	7.526	54.045	70.288
1.028	9.745	7.109	62.141	74.841
1.030	9.611	6.739	72.838	78.159

table 15. Performance characteristics of a 6-element Yagi beam with a boom length of $0.75\ \lambda_o$**, director spacing tapered pseudo parabolically according to the schedule of table 9 (mild positive interval).**

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.195	13.965	30.953	−35.247
0.972	10.257	14.993	30.705	−32.716
0.974	10.315	16.131	30.412	−30.152
0.976	10.369	17.412	30.079	−27.546
0.978	10.420	18.888	29.713	−24.890
0.980	10.469	20.636	29.322	−22.177
0.982	10.515	22.792	28.915	−19.400
0.984	10.558	25.606	28.500	−16.556
0.986	10.599	29.598	28.086	−13.639
0.988	10.637	35.381	27.683	−10.646
0.990	10.673	34.817	27.302	−7.573
0.992	10.704	29.112	26.952	−4.419
0.994	10.732	25.234	26.645	−1.182
0.996	10.755	22.471	26.391	2.142
0.998	10.772	20.342	26.203	5.553
1.000	10.782	18.610	26.096	9.053
1.002	10.785	17.148	26.086	12.658
1.004	10.780	15.884	26.190	16.356
1.006	10.765	14.770	26.428	20.151
1.008	10.740	13.777	26.823	24.044
1.010	10.704	12.882	27.405	28.036
1.012	10.655	12.069	28.209	32.129
1.014	10.595	11.328	29.282	36.321
1.016	10.522	10.652	30.680	40.609
1.018	10.437	10.033	32.481	44.981
1.020	10.341	9.470	34.781	49.413
1.022	10.236	8.959	37.715	53.861
1.024	10.123	8.501	41.468	58.239
1.026	10.004	8.096	46.283	62.379
1.028	9.882	7.747	52.474	65.967
1.030	9.761	7.458	60.408	68.405

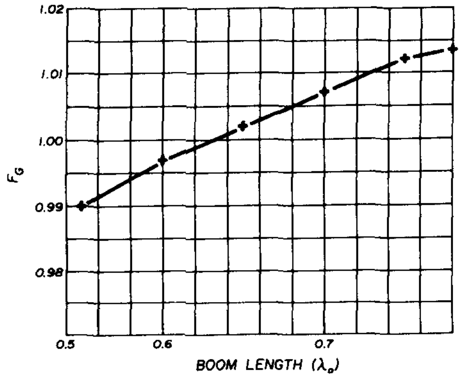


fig. 7. Plot illustrating frequency ratio for best central gain to best *F/B*. Note that frequency offset disappears for a boom length of $0.63\ \lambda_o$.

sary to take this offset into account in fixing the original bandwidth over which gain must be high. From a *gain* consideration alone the longer booms are best; that is why the example I used for illustrative purposes had a boom of $0.78\ \lambda_o$.

number of reflectors

It is interesting to consider a major change in possible Yagi-Uda antenna design: to explore the effect of changing the number of reflectors in a Yagi-Uda array. Up to this point we have assumed only a single reflector with a variable number of directors. It is tempting to consider increasing the number of reflectors in the hope of a significant improvement in the average *F/B* ratio over the entire bandwidth to be used. This question is now easily explored. I shall assume that the simplistic test Yagi-Uda of **fig. 1** will be our standard. To keep conditions other than the number of reflectors as constant as possible I shall keep the total boom length constant at $0.75\ \lambda_o$, and the total number of parasites constant at five. We shall compare the cases where the number of reflectors is zero, on (our test *standard*), two, and three. **Fig. 8** shows frequency-swept gain curves for all four cases; the curves are keyed to the legend on the diagram. Severe resonance effects are noticed near the free-space resonances of the reflector ($FR = 0.96$) and the directors ($FR = 1.06$); these resonances, however, were purposely spread far enough to allow the 4 per cent band of interest to display a good gain figure.

The highest curve (curve 1) displays gain for the *standard* simplistic Yagi-Uda (same as **fig. 1**) and it is clearly the best performer. The zero reflector case (curve 0) yields substantially less gain in the region of interest; it also contains no resonance effect at the reflector frequency, because there is no reflector. The two- and three-reflector cases (curves 2 and 3) show progressive loss of gain over the original standard; the reason is to be found in the much lower currents induced in the additional reflectors.

Shown below are the reflector currents when the driver is excited by one ampere at the central frequency ($F = 1.0$):

number of reflectors	magnitude of reflector current (amps) 1	2	3
1	0.477		
2	0.043	0.538	
3	0.031	0.046	0.626

Note that the reflector next to the driver has substantial current while all other drivers are hardly excited at all. Thus where there are multiple reflectors, the ef-

table 16. Performance comparison of 6-element Yagi beams with varying director lengths and element positions along the boom shows little gain variation.

from table	gain (dBi)	F=1.0 F/B (dB)	max F/B	at freq	resist (ohms)	angle K=2 null
1	10.857	17.04	38.03	0.988	20.23	144
3	10.781	18.36	39.99	0.988	19.54	144
4	10.654	16.50	31.91	0.986	15.79	138
5	11.030	11.55	29.69	0.980	13.36	138
6	10.959	14.21	31.59	0.984	16.79	141
7	10.732	19.89	38.70	0.992	23.25	150
8	10.581	22.03	33.64	0.994	25.35	153
10	10.587	18.32	22.19	0.992	24.43	150
11	10.747	17.88	27.21	0.990	22.02	147
12	10.942	16.17	34.62	0.988	18.62	141
13	11.014	15.30	26.55	0.988	16.78	138
14	10.749	17.65	28.61	0.988	28.65	144
15	10.782	18.61	35.38	0.988	26.10	147

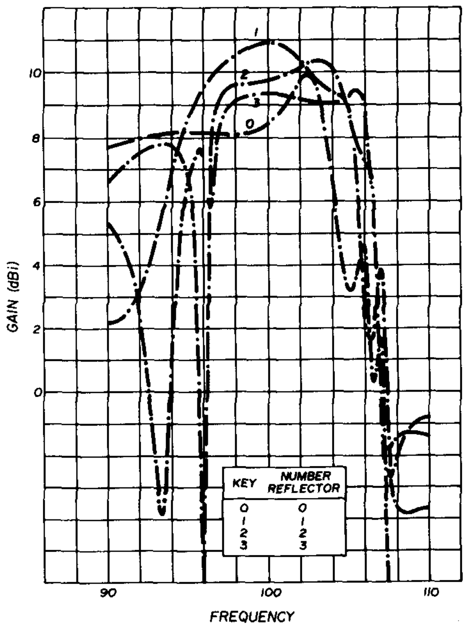


fig. 8. Gain of a 6-element Yagi beam vs. number of reflectors, overall number of elements held constant.

table 17. Performance vs frequency characteristics of a 6-element Yagi beam with a boom length of $0.75\lambda_o$, element positions optimized for maximum F/B ratio (at $F = 0.990$).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.960	9.482	8.066	26.753	−50.678
0.962	9.631	8.848	26.922	−47.892
0.964	9.762	9.661	27.057	−45.159
0.966	9.877	10.512	27.150	−42.467
0.968	9.980	11.410	27.195	−39.803
0.970	10.072	12.366	27.189	−37.154
0.972	10.154	13.397	27.133	−34.507
0.974	10.229	14.523	27.029	−31.849
0.976	10.298	15.773	26.881	−29.169
0.978	10.361	17.191	26.695	−26.458
0.980	10.420	18.843	26.476	−23.705
0.982	10.476	20.839	26.235	−20.903
0.984	10.528	23.386	25.980	−18.044
0.986	10.576	26.947	25.720	−15.124
0.988	10.621	33.001	25.467	−12.137
0.990	10.663	98.800	25.230	−9.078
0.992	10.701	33.035	25.021	−5.946
0.994	10.734	27.029	24.852	−2.737
0.996	10.762	23.517	24.735	0.551
0.998	10.785	21.024	24.685	3.920
1.000	10.801	19.090	24.716	7.372
1.002	10.810	17.505	24.845	10.919
1.004	10.810	16.165	25.094	14.554
1.006	10.800	15.008	25.486	18.276
1.008	10.781	13.990	26.049	22.087
1.010	10.750	13.086	26.817	25.985
1.012	10.707	12.277	27.833	29.965
1.014	10.653	11.549	29.153	34.021
1.016	10.587	10.893	30.847	38.136
1.018	10.509	10.303	33.009	42.280
1.020	10.421	9.775	35.761	46.397
1.022	10.324	9.307	39.263	50.385
1.024	10.219	8.900	43.721	54.062
1.026	10.109	8.555	49.385	57.097
1.028	9.996	8.276	56.488	58.895
1.030	9.884	8.069	65.091	58.439
1.032	9.776	7.943	74.616	54.170
1.034	9.675	7.912	82.965	44.366
1.036	9.585	7.994	85.869	28.942
1.038	9.507	8.216	79.043	12.212
1.040	9.442	8.618	63.459	1.471

fective boom length is shortened and we therefore should expect the gain to fall appreciably. **Fig. 9** shows the F/B ratio for these same four cases. Clearly the standard Yagi-Uda antenna (curve 1) is superior to the zero reflector case (curve 0). In the two-reflector case (curve 2) the peak of maximum F/B (corresponding to the $K = 2$ null) has moved significantly higher in frequency. We have already learned that this occurs when the effective boom length is reduced (in this case by the relatively ineffective first reflector). This effect is exaggerated in the three-reflector case (curve 3) where the effective boom is still shorter due to the first two relatively ineffective reflectors.

Thus we now see that there is a very good reason why a Yagi-Uda should contain one and only one reflector in the linear boom array; one is definitely needed to improve the gain and F/B. More than one reflector reduces the effective boom length and therefore gain; also, because of the relatively small currents induced in the extra reflectors, they do very little to the basic Yagi-Uda F/B ratio potential.

missing parasites

A common observation among Amateurs who have had large Yagi-Uda antennas in operation over a period of time is that when a parasitic element is broken or even entirely missing the Yagi continues to perform surprisingly well. We may now examine quantitatively just what occurs; for comparison I shall use the same 6-element simplistic Yagi-Uda design of **fig. 1**.

When a parasite is missing, the individual element currents all readjust to new values; such a readjustment changes the effective boom illumination function and therefore must cause a change in Yagi-Uda antenna performance. Starting with the standard 6-

table 18. Initial performance characteristics of the 6-element Yagi discussed in the text (boom length $= 0.78\lambda_o$).

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	10.198	14.828	23.551	−35.790
0.972	10.292	16.185	23.337	−32.860
0.974	10.379	17.740	23.107	−29.906
0.976	10.460	19.570	22.866	−26.920
0.978	10.535	21.802	22.618	−23.898
0.980	10.606	24.640	22.369	−20.835
0.982	10.671	28.327	22.126	−17.725
0.984	10.732	31.831	21.895	−14.566
0.986	10.787	30.198	21.685	−11.353
0.988	10.837	26.307	21.501	−8.085
0.990	10.881	23.196	21.354	−4.758
0.992	10.918	20.814	21.252	−1.371
0.994	10.947	18.917	21.204	2.078
0.996	10.969	17.352	21.221	5.591
0.998	10.981	16.024	21.314	9.170
1.000	10.983	14.873	21.498	12.815
1.002	10.975	13.860	21.786	16.528
1.004	10.955	12.959	22.197	20.309
1.006	10.924	12.149	22.751	24.159
1.008	10.881	11.419	23.473	28.076
1.010	10.825	10.756	24.395	32.058
1.012	10.757	10.156	25.553	36.100
1.014	10.679	9.612	26.997	40.193
1.016	10.590	9.121	28.787	44.320
1.018	10.492	8.681	31.003	48.451
1.020	10.387	8.293	33.747	52.534
1.022	10.278	7.956	37.152	56.479
1.024	10.167	7.674	41.383	60.131
1.026	10.056	7.449	46.637	63.216
1.028	9.949	7.288	53.114	65.264
1.030	9.850	7.200	60.879	65.480

table 19. Performance characteristics of the intermediate design 6-element Yagi described in the text; maximum *F/B* at *F = 1.0*.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.077	7.614	24.286	-40.497
0.972	9.271	8.371	24.167	-37.504
0.974	9.447	9.162	24.048	-34.535
0.976	9.609	9.994	23.925	-31.583
0.978	9.757	10.874	23.794	-28.645
0.980	9.892	11.812	23.652	-25.715
0.982	10.016	12.821	23.498	-22.786
0.984	10.130	13.916	23.332	-19.852
0.986	10.236	15.120	23.154	-16.907
0.988	10.333	16.461	22.966	-13.946
0.990	10.424	17.978	22.772	-10.962
0.992	10.509	19.718	22.574	-7.950
0.994	10.587	21.730	22.377	-4.906
0.996	10.660	24.001	22.187	-1.825
0.998	10.727	26.186	22.007	1.297
1.000	10.789	27.120	21.846	4.464
1.002	10.845	25.947	21.709	7.677
1.004	10.894	23.789	21.603	10.941
1.006	10.938	21.658	21.537	14.257
1.008	10.974	19.805	21.518	17.627
1.010	11.002	18.221	21.556	21.054
1.012	11.022	16.859	21.662	24.539
1.014	11.033	15.672	21.846	28.083
1.016	11.034	14.627	22.123	31.687
1.018	11.025	13.695	22.506	35.352
1.020	11.005	12.860	23.015	39.077
1.022	10.974	12.106	23.669	42.863
1.024	10.931	11.422	24.493	46.706
1.026	10.877	10.800	25.518	50.604
1.028	10.813	10.234	26.784	54.552
1.030	10.738	9.721	28.337	58.533

element simplistic design, I have made calculations of performance when one parasite is missing. Frequency swept plots of gain and *F/B* ratio are shown in **figs. 10** and **11**; the individual curves are keyed to the legend in the diagram. Note that there is still significant gain displayed for any of the cases.

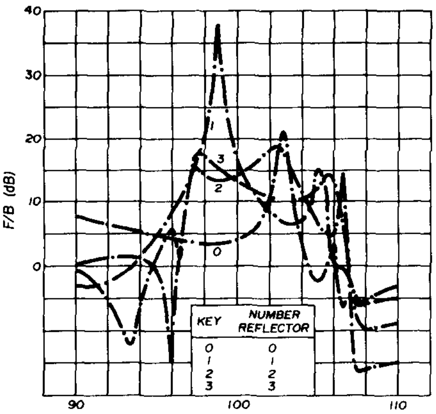


fig. 9. Front-to-back ratio vs. the number of reflector elements for a 6-element Yagi beam.

The greatest loss in performance occurs when the reflector, *R*, is missing; this, of course, is analogous to the previously discussed zero reflector case but now with a shorter (residual) effective boom. The *most surprising aspect of* **fig. 10** is the small but real *increase* in gain occasioned by the loss of *D3*. This can only be understood if the readjustment element currents constitute an effective boom illumination function slightly *longer* than that for the fully populated beam; in this event we would expect the frequency for maximum *F/B* to be *lower* than that for the standard case. **Fig. 11** shows this to be true. For all other cases of missing parasites the frequency of maximum *F/B* is *increased*, indicating a *shortened* effective boom length and hence lowered gain. The lowered gain is verified in **fig. 10**.

Thus a missing parasite is not always disastrous. However, if you look at the performance at the frequency of best *F/B*, the original fully populated Yagi-Uda is best.

table 20. Performance of the 6-element Yagi where the positions of directors *D1* and *D3* have been varied to "tune up" the *F/B* ratio.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.064	7.989	27.181	-43.099
0.972	9.235	8.747	27.090	-40.279
0.974	9.390	9.541	26.981	-37.483
0.976	9.532	10.378	26.850	-34.704
0.978	9.662	11.266	26.694	-31.935
0.980	9.781	12.217	26.511	-29.167
0.982	9.890	13.246	26.302	-26.391
0.984	9.992	14.373	26.069	-23.601
0.986	10.087	15.629	25.813	-20.788
0.988	10.177	17.054	25.539	-17.947
0.990	10.261	18.715	25.252	-15.070
0.992	10.342	20.723	24.957	-12.153
0.994	10.418	23.284	24.659	-9.190
0.996	10.491	26.860	24.366	-6.178
0.998	10.561	32.929	24.083	-3.112
1.000	10.627	119.848	23.819	0.010
1.002	10.690	33.007	23.580	3.192
1.004	10.749	27.018	23.375	6.437
1.006	10.804	23.524	23.212	9.746
1.008	10.853	21.049	23.100	13.123
1.010	10.898	19.133	23.050	16.570
1.012	10.935	17.569	23.072	20.089
1.014	10.966	16.249	23.181	23.684
1.016	10.989	15.109	23.386	27.358
1.018	11.003	14.107	23.707	31.113
1.020	11.008	13.214	24.166	34.952
1.022	11.002	12.412	24.787	38.876
1.024	10.986	11.686	25.602	42.889
1.026	10.958	11.026	26.645	46.990
1.028	10.919	10.425	27.969	51.180
1.030	10.868	9.876	29.634	55.447

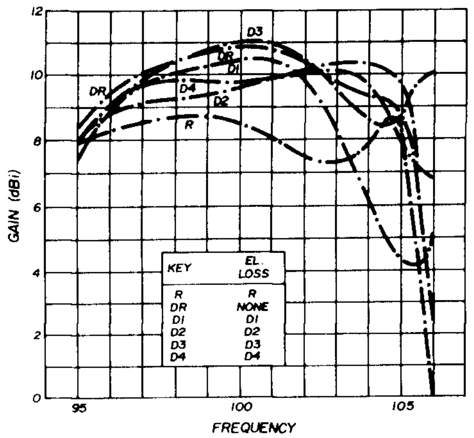


fig. 10. Forward gain of a 6-element Yagi showing performance when one element is missing. *F/B* under similar conditions is plotted in fig. 11.

It is now apparent that a Yagi-Uda antenna really "wants to work." Even major changes, such as a missing inner director, due to automatically readjusted element currents, works surprisingly well. It is now perfectly obvious why the Yagi-Uda antenna is so popular: it will provide *reasonable* performance no matter how it is constructed. It will provide top performance, especially in the *F/B* ratio, only if carefully made in accordance with the design rules presented in this article.

summary

In this article I have explored the effects of departures from the simplistic design previously given. The results show:

1. Director length taper schedules have no apparent beneficial effect on gain or *F/B* for boom lengths smaller than one wavelength. The important design parameter is the *average* director length — not the taper schedule.

2. Element placement schedules on the boom also have a marginal effect on gain or *F/B* for boom lengths less than one wavelength.

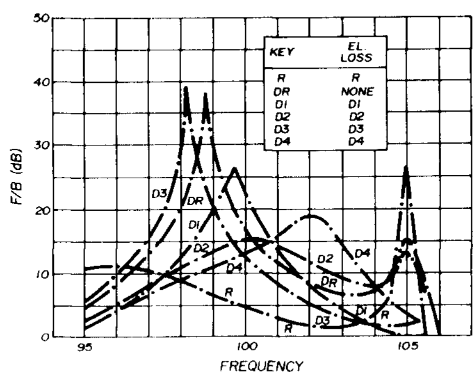


fig. 11. Front-to-back ratio of a 6-element Yagi showing the effect of a missing element. Forward gain under similar conditions is shown in fig. 10.

table 21. Performance of the 6-element Yagi where the positions of the driven element (*DR*) and director (*D3*) have been optimized through computer iteration.

frequency (F)	gain (dBi)	F/B ratio (dB)	feedpoint resistance (ohms)	feedpoint reactance (ohms)
0.970	9.432	9.225	31.638	−40.673
0.972	9.554	9.938	31.476	−38.105
0.974	9.667	10.688	31.276	−35.547
0.976	9.770	11.483	31.037	−32.990
0.978	9.866	12.331	30.758	−30.427
0.980	9.955	13.243	30.441	−27.848
0.982	10.039	14.234	30.089	−25.246
0.984	10.118	15.325	29.704	−22.613
0.986	10.193	16.545	29.294	−19.943
0.988	10.265	17.936	28.862	−17.229
0.990	10.334	19.565	28.414	−14.466
0.992	10.401	21.540	27.955	−11.651
0.994	10.465	24.069	27.493	−8.779
0.996	10.528	27.615	27.035	−5.846
0.998	10.588	33.654	26.587	−2.852
1.000	10.646	150.334	26.157	0.208
1.002	10.701	33.673	25.751	3.334
1.004	10.754	27.655	25.379	6.528
1.006	10.803	24.132	25.048	9.792
1.008	10.848	21.629	24.766	13.126
1.010	10.888	19.684	24.543	16.533
1.012	10.922	18.092	24.388	20.014
1.014	10.950	16.744	24.312	23.569
1.016	10.970	15.574	24.327	27.201
1.018	10.982	14.542	24.448	30.910
1.020	10.984	13.620	24.690	34.698
1.022	10.976	12.788	25.072	38.566
1.024	10.957	12.033	25.618	42.515
1.026	10.927	11.342	26.355	46.545
1.028	10.885	10.710	27.317	50.658
1.030	10.831	10.131	28.544	54.849

3. The simplistic design is as good as any design for boom lengths less than one wavelength.

4. A Yagi-Uda linear array on a given boom is best when it involves one and only one reflector element.

5. The *F/B* ratio at a given design frequency can, in principle, be increased without limit by iterative design procedure.

6. Very high values of *F/B* will be available only over very narrow bandwidths.

7. The Yagi-Uda antenna is basically very tolerant of major faults. Even missing parasitic elements cause surprisingly little deterioration in gain.

references

1. J.L. Lawson, W2PV, "Yagi Antenna Design: Performance of Multi-Element Simplistic Beams," *ham radio*, May, 1980, page 18.
2. J.L. Lawson, W2PV, "Yagi Antenna Design: Gain and Front-to-Back Ratio of Multi-Element Beams," *ham radio*, June, 1980, page 33.
3. P. Viezbicke, *Yagi Antenna Design*, NBS Technical Note 688 U.S. Department of Commerce, Washington, D.C., December, 1976.

ham radio

checking transmission lines with time-domain reflectometry

Most hams are familiar with the frequency-domain reflectometer, commonly known as the SWR bridge. In its various forms, this device can report the status of a transmission line under operating conditions at a single frequency (that of the transmitter — one frequency at a time). It shows the reflection coefficient, or SWR, depending on the scales employed. But if something is amiss (high SWR), the operator can't tell either the exact nature of the problem or its location. Both shorted and open lines will give the same reading regardless of length. Time-consuming tests may be required to localize the fault to the antenna, the transmission line itself, or the connectors. For as simple a test as continuity/absence of shorts, climbing the tower (in winter, yet) may even be required to disconnect the antenna and gain access to the distant terminals of the line.

An alternative approach, time-domain reflectometry, permits all measurements to be made in armchair comfort at the station end of the line. It can reveal the presence of open or short circuits or excessive resistance in the line. Additionally, the location of the problem can be determined, often within a foot or two, without going outside. The idea is not new but is not widely published in the Amateur literature.

time domain reflectometry

Here's the principle. A step of voltage is applied to the end of the line by a square-wave generator or pulse generator. The pulse has a very fast rise time. Initially the generator sees only the characteristic impedance, Z_0, of the line, which determines the current according to Ohm's law.

When the pulse reaches a discontinuity in the line (such as a break, high resistance, short circuit, or the end), it is reflected, or absorbed, or a combination of the two. The amount and phase of reflection are determined by the impedance of the discontinuity. Any reflected current travels back to the generator where it combines with the outgoing current to produce a resultant. The generator, if mismatched, will cause a second reflection, and so on, until the pulse dies away due to line attenuation.

Now if we connect an oscilloscope across the generator terminals (in the station) we'll see (if the sweep rate is right) the initial voltage step (a in **fig. 1A**), the resultant voltage caused by the sum of reflected and incident voltages (b), and so on for each stage in the reflection process. We'll see these reflections in real time as they occur (hence the name of the technique).

The time required for the first reflected pulse to return (t_1 in the figure) is twice the travel time for signals in the line. Thus:

$$t_1 = \frac{(2\,\ell_1/V)}{C} \qquad \textbf{(1)}$$

where ℓ_1 is the line length, V is the velocity factor,* and C is the speed of light, $3 \cdot 10^8$ meters/sec.). Of course if the line is resistively terminated in Z_0, no reflection occurs, and the generator waveform will be unmodified by returning pulses (**fig. 1G**).

Multiple reflections can be seen under favorable circumstances, such as when two types of line are connected in series, or a lossy connector is used. The nature of a resonant antenna can even be determined when the resonant frequency is within the oscilloscope bandpass.

*The apparent velocity factors seen by this technique seem to be lower than those published. Thus foam dielectric cables typically showed a $V = 0.6\text{-}0.7$ vs the 0.8 found in the handbooks. This may be partly due to inaccuracy in the scope time-base generator, or partly a real effect — perhaps V is a function of frequency.

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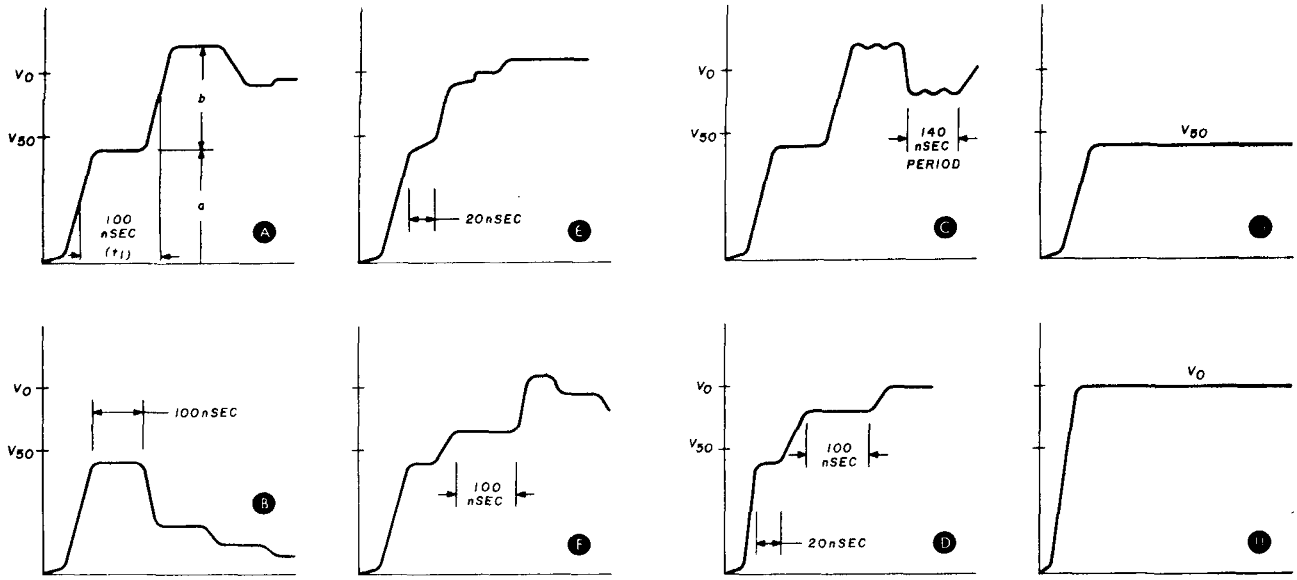

fig. 1. Time-domain reflectograms using the test setup in *fig. 2* for various line lengths and termination impedances (see *table 1*). *A* through *C* show the effect of termination on a single length of line. *D* through *E* show the effect of a series resistance at an intermediate point. *F* shows what happens when two lines of different characteristic impedance are connected in series. *G* shows a properly terminated line of any length. *H* shows the effect of no load (open-circuit generator output).

instrumentation

The apparatus for a practical setup is shown in **fig. 2.** To get sharp reflections the scope must have a wide bandwidth and the pulse or square-wave generator must have a fast rise time. My 15-MHz scope gives quite good results with lines of about 10 feet (3 meters) or more in length. TTL logic oscillators are quite good as generators if buffered or padded to 50 ohms. The scope calibrator can sometimes be used. Repetition rate (square-wave frequency) is slow enough to let the reflections die away after each pulse, but fast enough to give a good bright trace on the scope (typically 100 kHz-1 MHz). When testing real antennas, the voltage level should be kept to a minimum to avoid QRM. The minimum will be determined by the scope sensitivity.

A transformer is needed when using lines other than 50 ohms. For 300-ohm line (TV twinlead), a simple transformer on a FT-82-43 toroid worked well (primary, 50 ohms, 28 turns, secondary, 300 ohms, 43 turns). This transformer passes the frequencies involved (100 kHz-15 MHz) quite well, appearing transparent to the pulses. However, other baluns, matching devices, or transformers are not usually designed for this frequency range, so they will usually appear to be near short or open circuits, depending on their dc characteristics.

test patterns

A number of typical patterns are shown in **fig. 1** using the test setup of **fig. 2. Figs. 1A-C** show the effect of termination on a single length of line. Note

table 1. Data for determining line length and resistance parameters for responses shown in fig. 1.

figure	$_1$length feet	(meters)	$_2$length feet	(meters)	Z_o' (ohms)	R_{series} (ohms)	Z_1 (ohms)	Z_2 (ohms)
1A	34.0	(10.5)	0		—	—	∞	—
1B	34.0	(10.5)	0		—	—	0	—
1C	34.0	(10.5)	0		—	—	(note 1)	—
1D	6.6	(2.0)	34	(10.5)	50	100	∞	∞
1E	6.6	(2.0)	34	(10.5)	50	100	∞	50
1F	6.6	(2.0)	34	(10.5)	75	0	∞	∞
1G	(note 2)		0		—	0	50	—
1H	0		0		—	—	∞	—

Notes:
1. Termination was a 10-25-20 meter trap dipole. Note superposition of 14 and 21 MHz oscillations giving ~ 140 ns period (*7 MHz = 21-14*).
2. Any length. 2m and 10.5m were tried, using a dummy load.

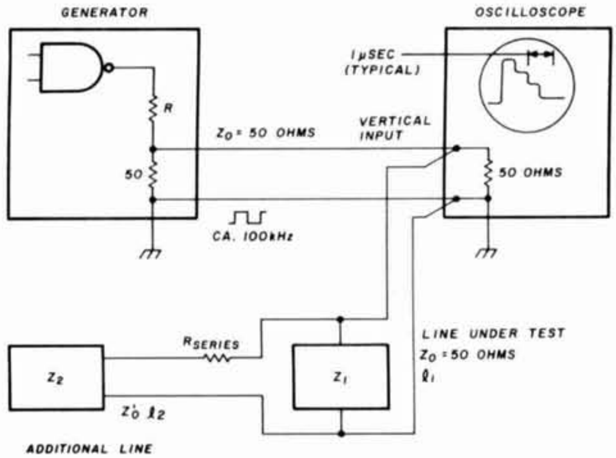

fig. 2. Test setup for line measurements. Oscilloscope must have a wide bandwidth and generator output must have a fast rise time. A 15-MHz scope gives good results with lines of moderate length (10 feet or 3 meters). *R* is adjustable to vary the generator output level. Notation corresponds to *table 1*.

that the trap dipole (no balun) is nearly an open circuit. In all three cases, the same length is obtained, within the accuracy of the scope time base. **Table 1** provides line lengths and impedance terminations for the time-domain reflectograms in **fig. 1**.

Figs. 1D-E show the effect of a series resistance at an intermediate point (such as a corroded connector). Note that in E, the properly terminated line looks like a pure resistance. Thus the effective length is ℓ_1 and the effective termination is 150 ohms.

Fig. 1F shows what happens when two lines of differing Z_0 are connected in series. The impedance bump shows clearly. And finally, **fig. 1G** shows a properly terminated line of any length. **Fig. 1H** is included to show that the voltages in the other cases are reduced from the generator open-circuit output, since the 50 ohms of the transmission line/load are in parallel with the generator output impedance.

closing comments

Since the technique is based on a step function, it covers the bandwidth from dc to *1/(rise time of scope or generator)*. Thus it's not suited for critical vhf applications except to show the location of a gross defect, such as a short circuit in the line. Neither will it tell much about the steady-state characteristics of an antenna at a fixed frequency. For this you need an impedance bridge or SWR meter. But when something goes wrong, it's sure nice to know exactly where — and time-domain reflectometry gives the answer to that.

bibliography

Allen, David M., "A Practical Experimenter's Approach to Time-Domain Reflectometry," *ham radio*, May, 1971, pages 22-27.

ham radio

More Details? CHECK—OFF Page 94

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More Details? CHECK—OFF Page 94

open quad antenna

An interesting approach to quad antenna design using phased radiators

In this article I describe a novel approach to the classic two-element quad antenna. I call it the "open quad." The designs described are for the Amateur 144- and 432-MHz bands, but they can be scaled for lower-frequency bands.

design approach

In this design I've used the regular driven element and reflector of the quad and have added director elements in the form of V and inverted V elements in a quasi-Yagi configuration. **Fig. 1** shows the general idea, and **fig. 2** shows the physical arrangement for a 10-element array for the 432-MHz band. Note that the parasitic elements are aligned with the two points of maximum current. Note also that the ends of the directors have been bent horizontally for a short distance. (This modification might be more useful for a larger antenna.)

advantages

The advantages of the open quad over the classic quad or Yagi derive not from any revolutionary concept, but rather from an attempt to combine the advantages of both designs:

1. Easier adjustment of the quad reflector for better front-to-back ratio.

2. Easier excitation of the driven element because of relatively higher impedance at the feed point.

3. Absence of feeders allows the exact phase of excitation, for top and bottom parasitic element.

4. High Q of director elements will increase gain.

5. Double V configuration makes for a collinear effect, which lowers the vertical radiation angle.

6. Possibility of eventually inserting a smaller antenna for operation on more than one band.

open quad for 432 MHz

In this design I attempted to duplicate the 13-element two-meter antenna described in reference 1. It is reproduced here in proportional scale. (I lacked a boom long enough for 13 elements.) Its characteristics include the following points:

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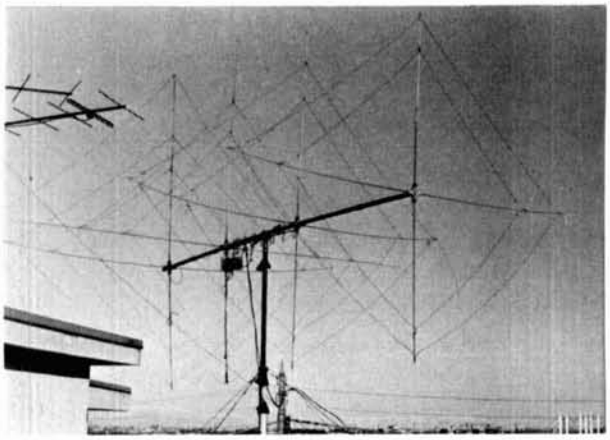


Open quad for the high-frequency bands based on the principles discussed in the text.

1. Forward gain: 17 dBd
2. Front-to-back ratio: 28 dB
3. Front-to-side ratio: 46 dB
4. Feedpoint impedance: 100 ohms

construction dimensions

I made the element lengths as follows for 432 MHz:

radiator: 27.9 inches (70.8 cm)
reflector: 29.7 inches (75.4 cm)
directors: 13.4 inches (34.0 cm) per element (two required for each director)

Element spacing for the 432-MHz quad was as follows:

reflector-to-radiator:	3.7 inches (9.5 cm)
radiator-to-director 1:	2 inches (5.5 cm)
director 1 to 2:	2.3 inches (6 cm)
director 2 to 3:	2.3 inches (6 cm)
director 3 to 4:	5.3 inches (13.5 cm)

The remaining directors were spaced 10 inches (26 cm).

The boom is about 6.4 feet (2 meters) long and made of fiberglass; the quad is fed with a 50-ohm coax cable and a 75-ohm quarter-wave matching section.

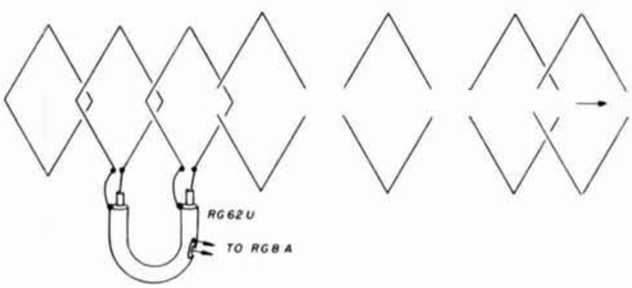


fig. 1. Design approach for the open quad antenna using two phased driven elements, with directors in the form of Vs and inverted Vs in a quasi-Yagi configuration.

open quad for 144 MHz

In this design I tried to achieve the desirable effect of two phased radiators spaced one-quarter wavelength apart. It's a well-known fact that maximum energy transfer between two antennas is a function of their spacing. The exact spacing follows a sequence of minimum and maximum current occurring at one-quarter wavelength between the two antennas. Tests have shown that an open quad with two in-line radiators, properly phased, will produce a signal not less than 0.7 times the maximum obtainable with the radiators spaced at the ideal distance.

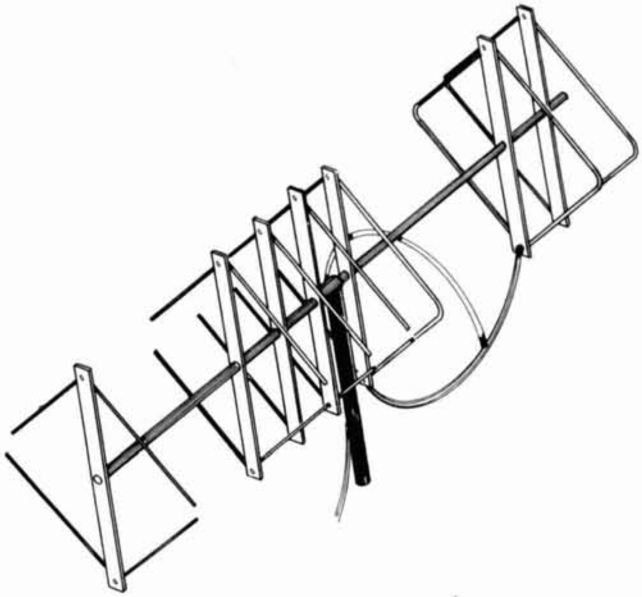


fig. 2. Physical arrangement of the open quad designed for the 432-MHz band.

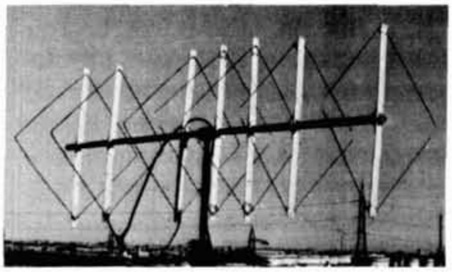
Characteristics of the 144-MHz open quad are:

Forward gain:	15 dB/d
Front-to-back ratio:	24 dB
Front-to-side ratio:	40 dB
Input impedance:	50 ohms

construction

I made the elements from aluminum tubing 0.2 inch (5 mm) in diameter. The boom was of fiberglass, as for the 432-MHz antenna. (The fiberglass boom was necessary to arrange one or more antennas across the boom of my high-frequency antenna.)

The matching section for the two radiators used lengths of 93-ohm coax cable (RG-62/U). The phasing line to the first (rearmost) radiator element was 6 inches (15 cm): that to the second radiator was 22



Author's seven-element open quad for the 2-meter band.

inches (56 cm). This matching section provided the proper phase shift between each radiator element for the energy transfer described above. The junction of these lines was soldered to a 50-ohm feedline.

Element lengths for 144 MHz were:

reflector:	6.98 feet (2.13 meters)
radiators:	7.40 feet (2.25 meters)
directors:	3.35 feet (1.02 meters) each part

Element spacing for the 144-MHz antenna was:

reflector to radiator 1:	11 inches (27.9 cm)
radiator 1 to radiator 2:	20 inches (51 cm)
radiator 2 to director 1:	8.2 inches (21 cm)
director 1 to director 2:	*8.2 inches (21 cm)*
director 3 to director 4:	15.75 inches (40 cm)

I tested a new, improved version using ten elements on a 13-foot (4-meter) boom with results better *than those of the big commercial antennas used* in Europe.

open quad for 10, 15, and 20 meters

It should be an easy matter to add one or two "open" directors to an existing high-frequency quad. I recommend the following spacing:

reflector to radiator:	0.14λ
radiator to director 1:	0.12λ

on-the-air tests

I tested the high-frequency open quad shown in *the photo with a station 20 miles (32 km) away with* the following results:

Direct line to station:	S9 + 10 dB
180 degrees from station:	S3
90 degrees from station:	*signal audible but* not readable

concluding remarks

The open quad can surely be improved, both from a mechanical and electrical standpoint, by someone with more sophisticated and precise instrumentation. I shall be glad to correspond with anyone wishing more information.

reference

1. William Orr, W6SAI, *The Radio Handbook*, 18th edition, Howard Sams, Indianapolis, 1969.

ham radio

More Details? CHECK—OFF Page 94

Low Cost...High Performance

DIGITAL MULTIMETER

$99.95 WIRED

Low cost, high performance, that's the DM-700. Unlike some of the hobby grade DMMs available, the DM-700 offers professional quality performance and appearance at a hobbyist price. It features 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 high LED display, with automatic decimal placement, automatic polarity, and overrange indication. You can depend upon the DM-700, state-of-the-art components such as a precision laser trimmed resistor array, semiconductor band gap reference, and reliable LSI circuitry insure lab quality performance for years to come. Basic DC volts and ohms accuracy is 0.1%, and you can measure voltage all the way from 100 μv to 1000 volts, current from 0.1 μa to 2.0 amps and resistance from 0.1 ohms to 20 megohms. Overload protection is inherent in the design of the DM-700, 1250 volts, AC or DC on all ranges, making it virtually goof proof. Power is supplied by four 'C' size cells, making the DM-700 portable, and, as options, a nicad battery pack and AC adapter are available. The DM-700 features a handsome, jet black, rugged ABS case with convenient retractable tilt bail. All factory wired units are covered by a one year limited warranty and kits have a 90 day parts warranty.

Order a DM-700, examine it for 10 days, and if you're not satisifed in every way, return it in original form for a prompt refund.

Specifications

DC and AC volts:	100 μV to 1000 Volts, 5 ranges
DC and AC current:	0.1 μA to 2.0 Amps, 5 ranges
Resistance:	0.1Ω to 20 megohms, 6 ranges
Input protection:	1250 volts AC/DC all ranges fuse protected for overcurrent
Input impedance:	10 megohms, DC/AC volts
Display:	3½ digits, 0.5 inch LED
Accuracy:	0.1% basic DC volts
Power:	4 'C' cells, optional nicad pack, or AC adapter
Size:	6"W x 3"H x 6"D
Weight:	2 lbs with batteries

Prices

DM-700 wired + tested	$99.95
DM-700 kit form	79.95
AC adapter/charger	4.95
Nicad pack with AC adapter/charger	19.95
Probe kit	3.95

TERMS: Satisfaction guaranteed or money refunded. COD, add $1.50. Minimum order $6.00. Orders under $10.00, add $.75. Add 5% for postage, insurance, handling. Overseas, add 15%. NY residents, add 7% tax.

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600 mHz COUNTER

$99.95 WIRED

The CT-70 breaks the price barrier on lab quality frequency counters. No longer do you have to settle for a kit, half-kit or poor performance, the CT-70 is completely wired and tested, features professional quality construction and specifications, plus is covered by a one year warranty. Power for the CT-70 is provided by four 'AA' size batteries or 12 volts, AC or DC, available as options are a nicad battery pack, and AC adapter. Three selectable frequency ranges, each with its own pre-amp, enable you to make accurate measurements from less than 10 Hz to greater than 600 mHz. All switches are conveniently located on the front panel for ease of operation, and a single input jack eliminates the need to change cables as different ranges are selected. Accurate readings are insured by the use of a large 0.4 inch seven digit LED display, a 1.0 ppm TCXO time base and a handy LED gate light indicator.

The CT-70 is the answer to all your measurement needs, in the field, in the lab, or in the ham shack. Order yours today, examine it for 10 days, if you're not completely satisfied, return the unit for a prompt and courteous refund.

Specifications

Frequency range:	10 Hz to over 600 mHz
Sensitivity:	less than 25 mv to 150 mHz less than 150 mv to 600 mHz
Stability:	1.0 ppm, 20-40°C; 0.05 ppm/°C TCXO crystal time base
Display:	7 digits, LED, 0.4 inch height
Input protection:	50 VAC to 60 mHz, 10 VAC to 600 mHz
Input impedance:	1 megohm, 6 and 60 mHz ranges 50 ohms, 600 mHz range
Power:	4 'AA' cells, 12 V AC/DC
Gate:	0.1 sec and 1.0 sec LED gate light
Decimal point:	Automatic, all ranges
Size:	5"W x 1½"H x 5½"D
Weight:	1 lb with batteries

Prices

CT-70 wired + tested	$99.95
CT-70 kit form	75.95
AC adapter	4.95
Nicad pack with AC adapter/charger	9.95
Telescopic whip antenna, BNC plug	7.95
Tilt bail assembly	3.95

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More Details? CHECK—OFF Page 94

microwave-frequency converter

for vhf counters

Theory and construction of a frequency scaler that can be used to convert 500-2500 MHz to frequencies acceptable to a 500-MHz counter

Measuring frequency accurately to 500 MHz is inexpensive and easy today. Several good 500-MHz counters are available for about $100.00. These are of little value to the microwave experimenter, however, whose world is just beginning at 500 MHz. For lower-frequency microwave measurements, up to 2500 MHz or so, a good counter will set you back $1000 or more.

This article describes a heterodyne frequency converter that will convert frequencies from 500-2500 MHz to below 500 MHz, enabling you to use your 500-MHz counter for microwave measurements. Depending on your resourcefulness at pleading with your local manufacturers' representatives to supply you with a few key parts, your parts cost will be anywhere from nothing to $100.00. Construction techniques should be followed closely unless you're familiar with uhf construction practices.

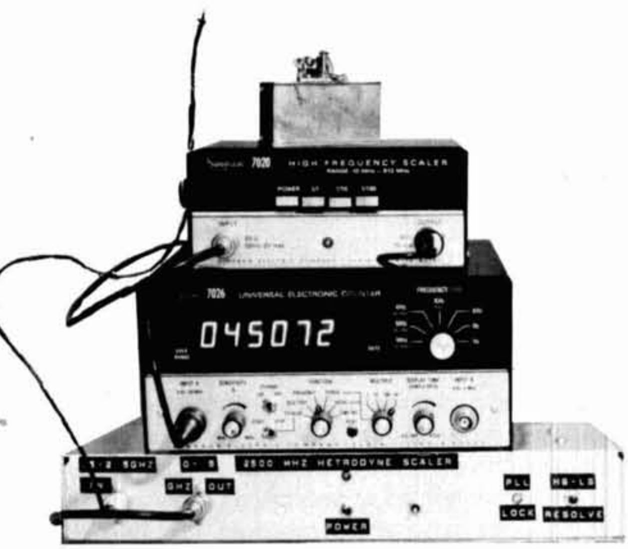
theory

The heart of this project is a double-balanced mixer, the Engelmann Microwave MLP-101 (see **fig. 6**). I recommend it because of its low cost — $35.00. Neglecting third- and higher-order products, a balanced mixer output spectrum contains two frequencies — $F_{RF} + R_{LO}$ and $F_{RF} - F_{LO}$ as in **fig. 1**.

For our purposes, $F_{RF} - F_{LO}$ will be the spectral component of interest since it translates an unknown radio frequency to a 0-500 MHz i-f by a known LO frequency: in this case 1000 or 2000 MHz. This makes your present counter direct reading in the lower two Amateur microwave bands, 1215 - 1300 MHz and 2300 - 2450 MHz, by simply mentally adding a 1 or a 2 in front of the displayed frequency, depending on the LO frequency you choose. To gain an overview of how the local-oscillator signal is generated, see **fig. 2**.

Local oscillator. The local oscillator string consists of a 1-MHz crystal reference frequency oscillator, a phase-locked-loop, X500 frequency multiplier stage, and two frequency-doubler stages. The 1-MHz reference oscillator is a conventional CMOS inverter using a parallel-resonant crystal. The loop phase detector is a 4046 CMOS IC using the sample-and-hold phase detector. All other stages are fairly conventional with a

By David R. Pacholok, KA9BYI, 437 North Crystal, Elgin, Illinois 60120



The heterodyne scaler in operation. Oscillator frequency measured 2450.72 MHz.

few exceptions. The loop filter is a two-pole design as opposed to the usual single-pole variety. The second pole resides in the "Spurrie filter" (see **fig. 2**). Its purpose is to attenuate unwanted phase-detector outputs ($N \times 1\ MHz$), which would otherwise frequency modulate the VCO, producing sidebands at 500 ± 1, 2, 3 . . . MHz. As it stands, all such reference frequency spurs are at least 50 dB down from the 500-MHz fundamental.

Frequency multipliers. The frequency doublers may seem a bit unusual to some at first glance. This is because uhf transistor frequency multipliers don't operate in the same way as their high-frequency counterparts. High-frequency multipliers operate on the principle of reduced conduction angle (60°-120°) collector-current pulses containing large quantities of harmonics, which may be filtered by a high-Q tank circuit to yield the desired harmonic output. Alas, very few transistors can switch fast enough to provide low conduction angles at 1000 MHz. Rise storage and fall times preclude operation as a conduction-angle-based frequency multiplier above a few hundred MHz or so.

Fortunately, a thoughtful electron god has included other nonlinearities into transistors that make them useful frequency multipliers. The most important of these is the varactor effect in the collector-base (C-B) junction. (See **fig. 3** for a general explanation of how vacactor multipliers work.)

Transistor-varactor-effect frequency multipliers must do three things simultaneously: 1) they must amplify the input frequency, 2) they must apply this amplified F_{IN} efficiently to their own C-B junction for frequency multiplication, and 3) they must extract and filter the desired harmonic from the C-B junction.

analysis

The circuit of **fig. 4** shows how to accomplish the objectives outlined above. To be a good F_{IN} amplifier, the transistor's conjugate input impedance must be matched; in this case to a 50-ohm input. From **Z**-parameter data, the transistor's input is $Z = 13 + j14$ at 1 GHz. This translates into a parallel-equivalent conductance of about 0.04-0.04B. Since $Mho = 1/ohm$, the transistor looks like a 25-ohm resistor in parallel with a 25-ohm inductor. My input-matching strategy is to cancel the 25-ohm inductance with a parallel 25-ohm capacitive reactance (LN2), leaving only the 25-ohm resistance. A quarter-wave impedance transformer, LN1, of $Z_0 = 35$ ohms converts 25 ohms resistive into 50 ohms resistive impedance.

$$Z_0 = \sqrt{Z_{IN}Z_{OUT}} = \sqrt{50 \times 25} = 35\ ohms \qquad \textbf{(1)}$$

The 25-ohm capacitive reactance was especially chosen, however. It consists of an eighth wavelength, at F_0IN, of $Z_0 = 25\text{-}ohm$ line open circuited at the far end. At $2F_0IN$, the desired output frequency, it is a quarter-wave open-circuited line and therefore an effective $2F_0IN$ short circuit.

This line shorts the transistor base to ground at $2F_0IN$, meaning that $2F_0IN$ energy can flow from the transistor C-B junction to the $2F_0IN$ filter without dissipating in the resistive B-E junction. This $2F_0IN$ trap greatly improves frequency-doubler efficiency.

On the collector side of the frequency doubler, L1C1 form an "F_0IN idler." This tank circuit is roughly analogous to the sine-wave current source in **fig. 3**. Note that when properly adjusted, L1C1 are not series resonant at F_0IN. (Series resonance would short-circuit the F_0IN source, leaving no energy for the current source to drive F_0IN through the varactor to generate harmonics.) Instead, the series combination of C1 and C_{CB} (average value) and L1 are parallel resonant to provide a high circulating F_0IN current

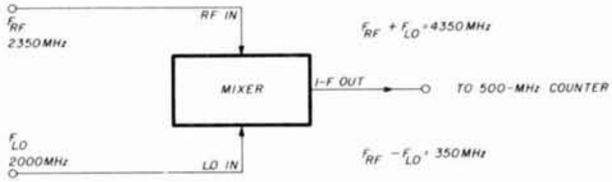
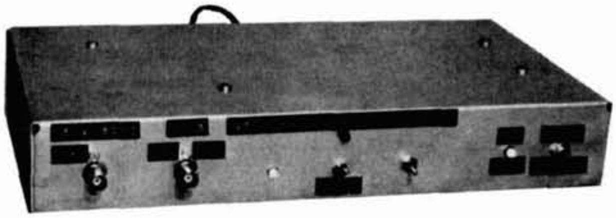


fig. 1. The heart of the heterodyne frequency scaler is a double-balanced mixer. The spectral component of interest is $F_{RF} - F_{LO}$, which makes your frequency counter direct reading in the lower two Amateur microwave bands, 1215-1300 MHz and 2300-2450 MHz.



Front panel of the heterodyne scaler.

through C_{EB}, where harmonic generation occurs.

LN3, in conjunction with C2, forms a half-wavelength line, high-Q output filter tuned to $2F_0IN$. The ratio of C2 to C3 allows for input and output matching.

The broadband amplifier (Q_{AMP} in **fig. 6**) makes up for the 7-dB loss in the mixer, giving the frequency converter as good or better sensitivity than the counter with which it's used.

The only part you may have difficulty obtaining inexpensively (or for free) is the 2500-MHz double-balanced mixer. The most inexpensive suitable unit I've found is the MLP101 (or MLF101) made by Engelmann Microwave; it cost me $35.00.

The 1-MHz crystal should be of time-base quality, as its frequency is multiplied by 2000 on the converter high-MHz range. If your counter has a 1-MHz time base, you can omit the crystal.

The 0.032-inch (0.8-mm) Teflon/glass double-sided board specified for this project may be obtained from Oak Laminates. Cincinnati Millacron polyester glass board (Milliclad[R]) is a very good, inexpensive substitute for the TFE glass but is only available in 0.062-inch (1.6-mm) double-sided board. Its dielectric constant is a function of frequency, and some experimentation will be necessary to make it work, except as indicated in **fig. 5**.

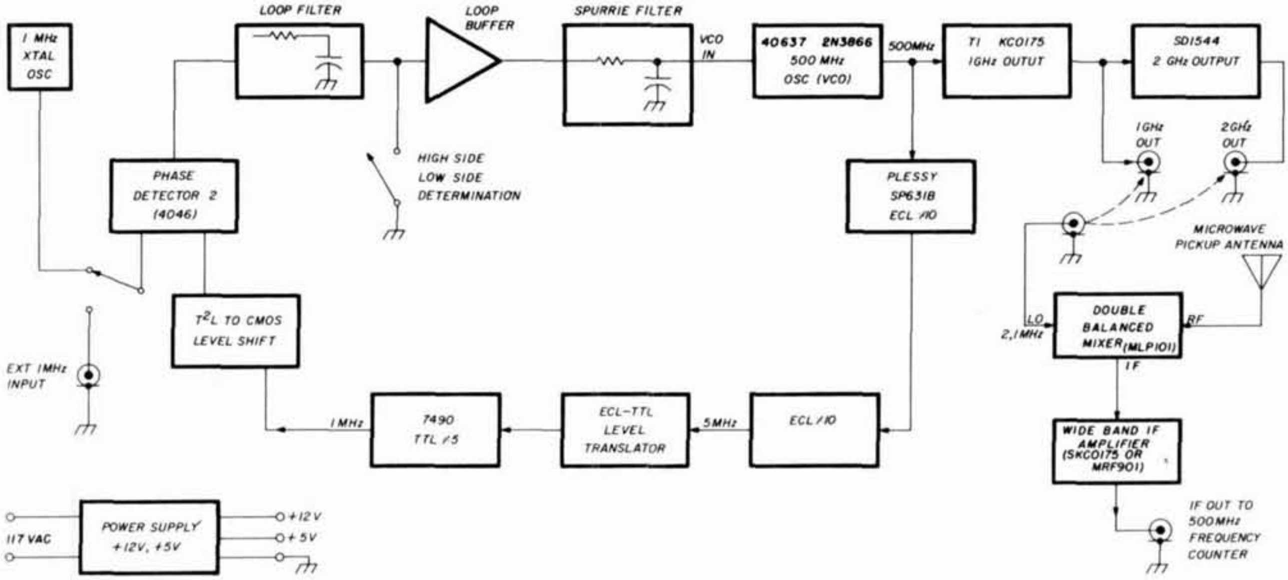


fig. 2. Heterodyne frequency scaler block diagram. Local-oscillator string consists of a 1-MHz crystal reference frequency oscillator, a phase-locked x 500 frequency multiplier, and two frequency doublers. The varacter effect in the frequency-multiplier transistors is used for efficient frequency multiplication in the Gigahertz region.

Amplifier gain is 10 dB ± 1.5 dB between 5-500 MHz. Try to obtain an SKC0175 for this stage, as any other device probably won't work too well without extensive circuit mods. The broadband amplifier can be omitted at some loss (about 10 dB) of scaler sensitivity.

obtaining parts

The only nonstandard parts used in the converter (**fig. 6**) are transistors QD1, QD2, the Plessy SP631B, and of course the MLP101 double-balanced mixer IC. Possible substitutes for the Texas Instruments SKC0175 (QD1) are the 2N5770 and the 2N3866. Possible substitutes for QD2, a Solid State Microwave SD1544, are the 2N5108 and the MRF8009. The Plessy SP631B may be replaced by any ECL 500-MHz divide-by-ten prescaler chip.

Rear panel of heterodyne scaler showing L.O. changing jacks.

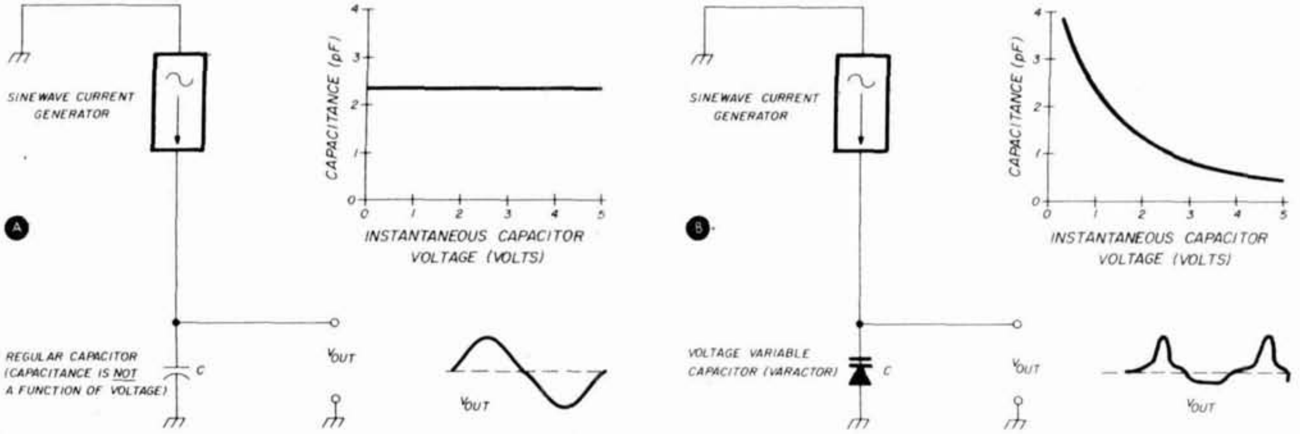


fig. 3. Comparison of a sine-wave current generator terminated by a regular capacitor, A, and one using a varactor, B. In A, V_{OUT} is an undistorted, phase-lagging sine wave (i.e., a cosine wave). In the varactor application, B, as V_{OUT} increases in amplitude, C decreases, making V_{OUT} rise ever faster. V_{OUT} will then become badly distorted, creating harmonics that may be filtered out to yield the desired output signal.

construction

Construction of the 2500-MHz heterodyne scaler may be divided into two parts: 1) power supply, reference oscillator, phase detector, loop filters and buffer; and 2) the VCO, divider string, doublers, mixer, and broadband i-f amplifier.

Part 1 may be built using any construction practices you desire, as layout and grounding aren't critical. Part 2, however, should be built using plain unetched 0.064-inch (1.6-mm) or thicker copper-clad board (material uncritical) as a goundplane.

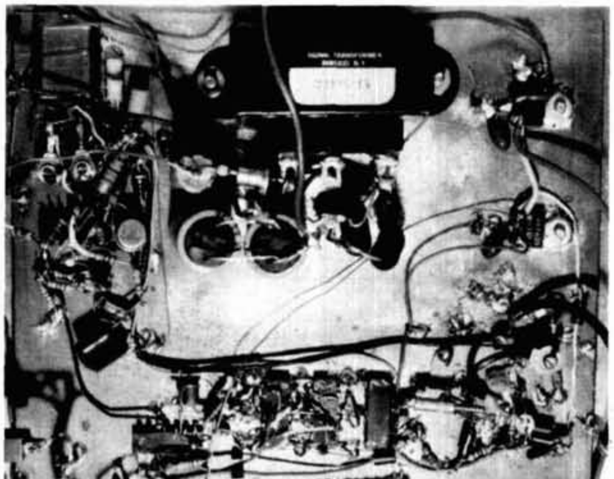
Mount the divider ECL and TTL ICs by turning them upside down and bending all their ground pins to meet the copper-clad board, then solder them in place. All ECL point-to-point wiring can then be made using *very short* lengths of solid hookup wire. By proper forethought as to layout, it's possible to eliminate any inter-IC lead length greater than 5/8 inch (16 mm). If for some reason you must make a longer run in the divider chain, use shielded 50-ohm cable. Run all bypass capacitors from their IC pins to the groundplane with lead length as short as possible. possible.

The same basic construction practices outlined above apply to the VCO and broadband amplifier, except that here you deal with transistors instead of ICs. The doublers, as they use stripline circuitry, deserve special mention. Stripline, as done in industry, is a photoetching process that leaves dual-sided Teflon copper-clad board with all its copper intact on one side (the groundplane) and etched strip-like conductors on the other side. The strip conductors work with the groundplane to form a transmission line whose impedance is a function of strip width, height above groundplane, and dielectric constant of the Teflon glass board.

For our purposes, however, we can use a different technique, which is much more inexpensive, easier, and alterable too. It involves cutting the desired lines out of Teflon PC board stock, and then cementing them to a far-less-expensive, unetched G10 or phenolic-base circuit board.

To duplicate this technique, obtain a 6 × 6 inch (152 × 152 mm) piece of single-sided, unetched, copper-clad board (material not critical). Use steel wool to polish the copper side until it shines. Then refer to **fig. 5** to obtain stripline dimensions for each line used.

Using your stock of Teflon-glass board, cut out lines to these exact dimensions. Polish both sides of



Internal view of scaler showing power supply and L.O. frequency synthesizer. This board outputs 500.000 MHz.

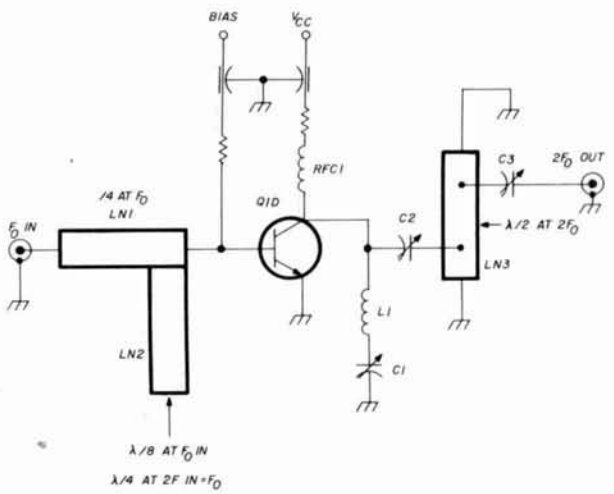


fig. 4. How to accomplish the objectives of a transistor-varactor-effect frequency multiplier. Stripline techniques are used for impedance transformation.

each line with steel wool. Using all doubler, mixer, and i-f amplifier parts, come up with a suitable layout for this assembly. When you're happy with the layout, drill holes for all feedthrough capacitors and transistors in the 6 × 6 inch (152 × 152 mm) circuit board chassis. Use *Super Glue* to mount all polished striplines to the *copper side* of this board to form instant, repairable striplines. Finally, mount both transistors and other parts with leads as short as possible. Mount the MLP101 mixer by soldering it to the circuit board in two or three spots, being very careful not to overheat the device.

A well-shielded box is recommended to house the heterodyne scaler, as it radiates on many harmonics of 1 MHz. It's a good idea to use feedthrough capacitors on the power-line cord to prevent radiation leakage.

tune up

Tune up begins by locating a volt-ohmmeter, oscilloscope (*bandwidth* $\geq$ *5 MHz*), a frequency counter, and a set of nonmetallic alignment tools. Make yourself comfortable and expect several hours of entertainment (or frustration).

1. Break the connection between the 4700-μF filter cap and the 7812 regulator. Insert a milliammeter here to measure total supply current. If you read 250-350 mA, all is probably well.

2. Connect a voltmeter to the +12 and +5 volt sources, just to make sure the regulators are wired correctly. The collector of Q_{D1} should read 7 volts ± 1 volt, that of Q_{D2} 8 volts ± 1 volt, and that of Q_{AMP} about 5 volts ± 1 volt. If these voltages are other than specified, adjust stage bias accordingly; that is, if the collector voltage is too low increase the bias resistor value and *vice versa*.

3. Ready your frequency counter, and, with the reference frequency switch set to INTERNAL, measure the 1-MHz reference frequency at the 4046, pin 14. Set this frequency to 1.0000 MHz.

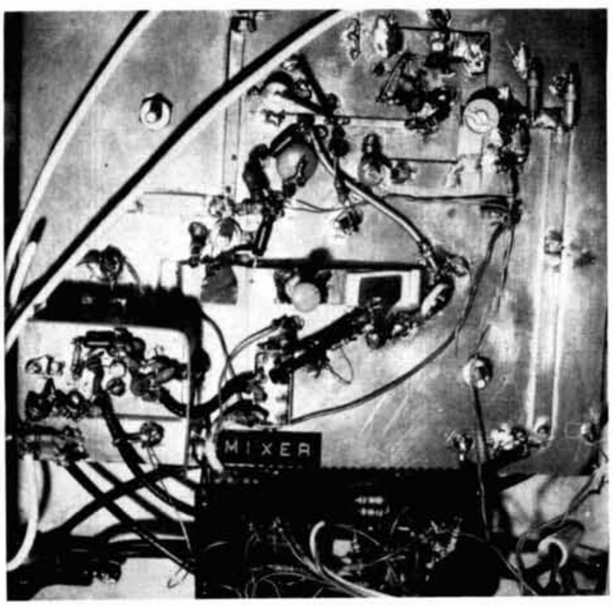
4. Loosely couple your counter to the VCO emitter through a 1-pF capacitor, after closing C_{TUNE} completely and disconnecting the 500-MHz OUT terminal. A frequency between 350 and 500 MHz should be observed here.

5. Observe the waveform at TP1. It should be a clean, jitter-free, TTL-level square wave of 700 kHz to 1 MHz. If not, adjust the 0.8-6 pF piston trimmer on the VCO emitter until the proper signal is obtained. You'll want 1-2 turns more capacitance (from the trimmer) than the minimum required to get a clean waveform at TP1.

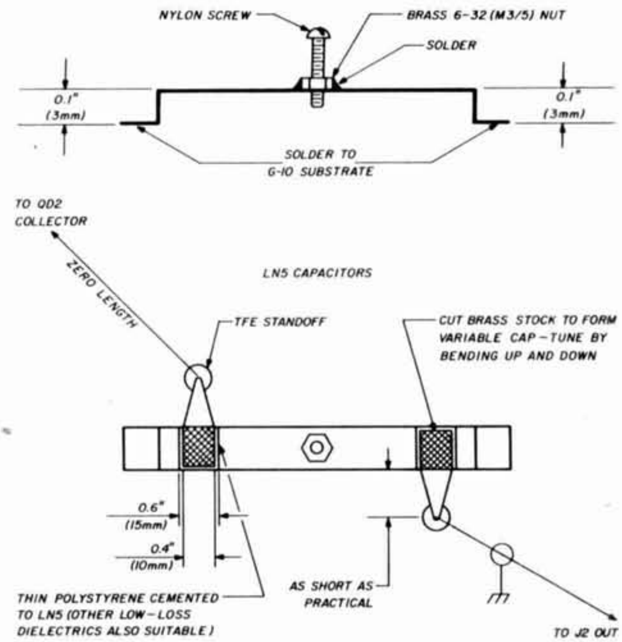
6. Slowly reduce the capacitance of C_{TUNE} until the loop locks. Lock is established when the OUT-OF-LOCK pilot goes out and the measured frequency of the VCO is exactly 500.0 MHz.

7. Connect a 47-ohm, 1/4-watt resistor, with *short leads*, between 500-MHz OUT and ground. The VCO will probably lose lock at 450 MHz or so.

8. Tune C_{TUNE} and C_{LOAD} for maximum power *and* a locked loop by measuring the dc voltage at TP2. You should measure at least 1 volt, but don't tune for more than 1.5 volts.



Internal view of scaler showing the two frequency multipliers and the broadband i-f amplifier.



LN1 3.9 × 0.2 × 0.032 inch (99 × 5 × 0.8 mm) glass Teflon

LN2 3 × 0.2 × 0.064 inch (76 × 6 × 1.6 mm) Cincinnati Millacron *Milliclad* or 3.5 × 0.2 × 0.032 inch (89 × 4 × 0.8 mm) glass Teflon

LN3 2 × 0.1 × 0.032 inch (52 × 3 × 0.8 mm) glass Teflon

LN4 0.975 × 0.2 × 0.032 inch (25 × 5.5 × 0.8 mm) glass Teflon

LN5 2.7 inches (70 mm) of 0.39-inch (10-mm) brass with hole drilled in exact center and a 6-32 (M 3/5) brass nut soldered squarely over hole, then bent as shown

L1 1.4 inches (35 mm) no. 12 (2.1-mm) bare copper wire bent around a 0.5-inch (13-mm) coil form. A wooden dowel is suitable

L2 3 turns no. 20 (0.8-mm) enamelled wire wound on a 0.1-inch (3-mm) coil form, closely spaced to start with

L3 20 turns no. 30 (0.25-mm) enamelled wire wound around a 0.5-watt resistor lead. Form the coil to occupy 0.4 inch (9 mm)

fig. 5. Construction details for the microwave striplines and inductors used in the frequency scaler.

9. Disconnect the 47-ohm resistor and reconnect *LN1* to 500-MHz OUT.

10. Move the voltmeter to TP3 and terminate BNC connector J1 $_{OUT}$ with about 100 feet (30.5 meters) of RG-58/U or RG-174/U cable (far-end termination is not critical).

11. Tune both doubler 1 variable caps until you obtain maximum otuput at TP3. You should obtain 1.5-2.5 volts. If not, stretch *L2* turns for maximum output.

12. Verify that the loop OUT-OF-LOCK pilot light is still out. If not, slightly adjust the VCO C_{TUNE} control until loop lock is re-established.

13. Connect the lossy coax load to $J2_{OUT}$. Jumper $J1_{IN}$ and $J2_{OUT}$ with a coax cable.

14. Adjust the 1-GHz idler and the nylon screw in *LN5* and its two flap caps (**fig. 5**) for maximum output at TP4. If the signal suddenly disappears, try a little more 1-GHz idler capacitance. Several iterations will be required to get the maximum voltage level at TP4. About 2 volts should be obtainable if a quality Schottky detector or point-contact mixer diode is used at TP4. If your output is a little low here, don't worry. Accurate voltage measurements using crude peak detectors are a joke at 2 GHz anyway!

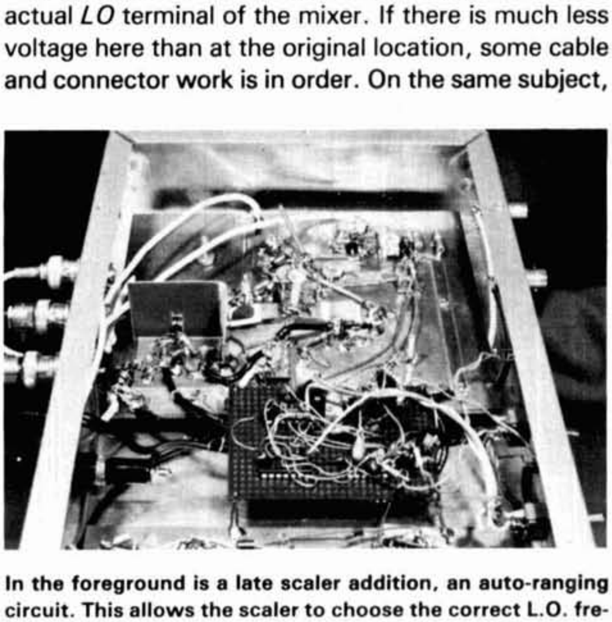
15. Remove the coax dummy load from J2 and insert the *LO* cable into J2. The reading at TP4 should be little different than with the dummy load.

further hints

If you build this unit, I assume you have some pet microwave project (or at least an oscillator) that you'd like to test and improve. Now is the time to do it. If you think your unit operates between 1.5 and 2.5 GHz, leave the heterodyne scaler set up the way it is. If you think your unit's output is between 500 and 1500 MHz, connect the LO_{IN} jack to $J1_{OUT}$. Connect your frequency counter to 0-500-MHz OUT. A suitable pickup antenna for this frequency is a half-wave coaxial dipole with a 6-inch (152-mm) overall length for the low band and a 2-1/2 to 3 inch (64 to 77 mm) length for the high band.

If you have any problems with scaler sensitivity — especially on the high band — and have followed my instructions, try one more thing.

Move the peak detector associated with TP4 to the actual *LO* terminal of the mixer. If there is much less voltage here than at the original location, some cable and connector work is in order. On the same subject,



In the foreground is a late scaler addition, an auto-ranging circuit. This allows the scaler to choose the correct L.O. frequency automatically or allow manual selection by a front panel switch.

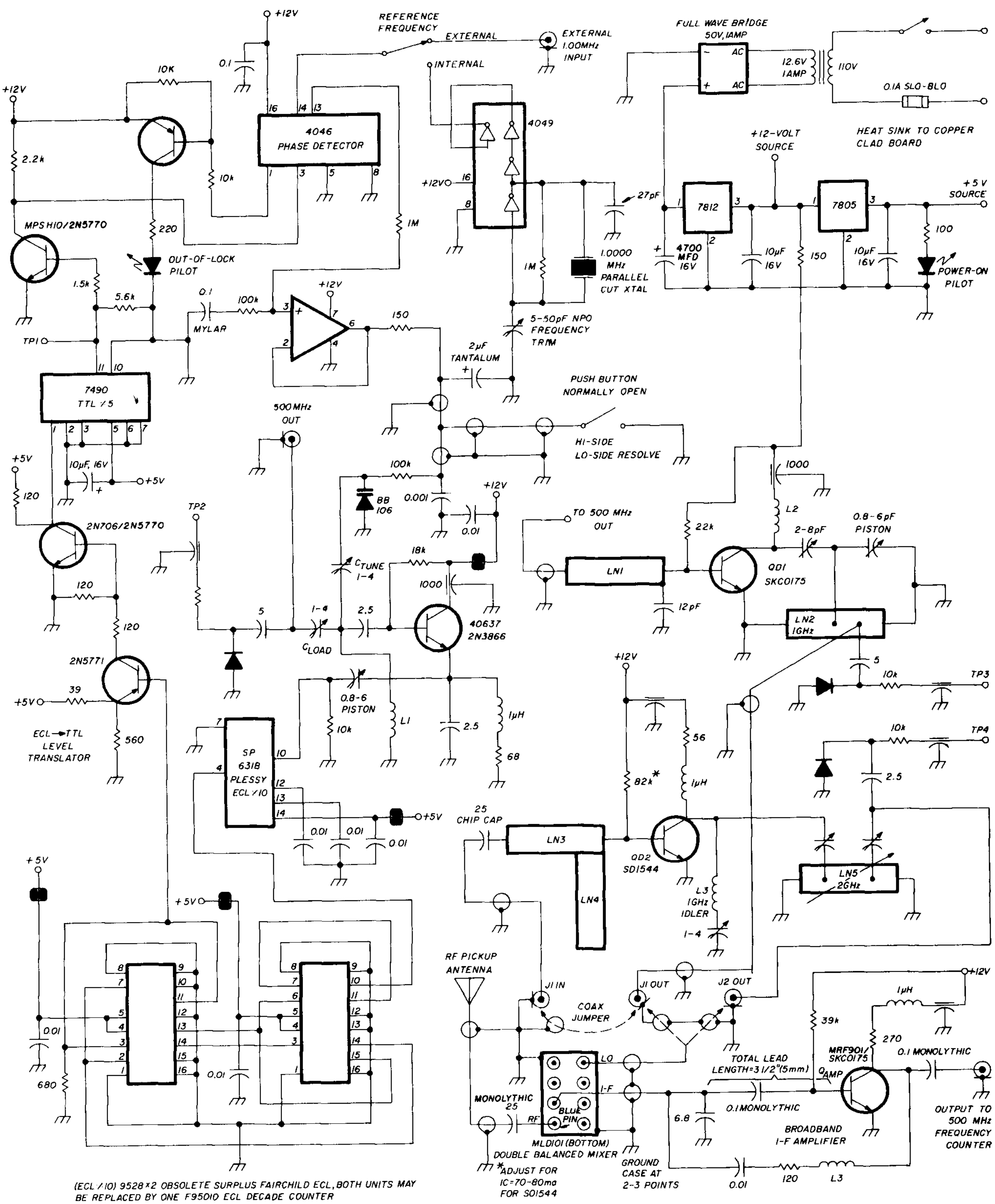


fig. 6. Schematic diagram of the 2500-MHz heterodyne frequency scaler.

if you're interested in only one band, forget about the back-panel jacks and handwire the connections.

operation

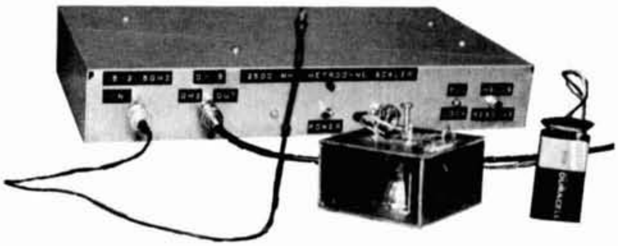
Operation of the heterodyne scaler is nearly as simple as using your counter with one or two exceptions. The first is the need to depress the HIGH SIDE-LOW SIDE RESOLVE switch when making a measurement. If, when this switch is depressed, the indicated frequency *increases*, your microwave signal is on the high side of the 1- or 2-GHz local-oscillator frequency, and your actual frequency is the *LO* frequency *plus* the counter displayed frequency. Conversely, if the measured frequency *decreases* when this switch is depressed, your microwave signal is on the low side of the *LO* frequency, and its actual frequency is the *LO* frequency in use *minus* the frequency your counter is measuring.

The second exception is overload protection. Your counter probably has fairly good overload anti-burnout circuitry; this scaler does not. The usual back-to-back hot-carrier diodes that provide protection at low frequencies are simply too reactive to do much good at 2.5 GHz. So be careful, and this scaler will serve you well. Don't do anything rash like trying to measure the frequency of your microwave oven by inserting the pickup antenna directly into the oven cavity!

Caution must also be used when trying to measure frequencies closer than 2 MHz away from the 1- or 2-GHz *LO* frequency. The double-balanced mixer as well as the broadband amplifier frequencies fall off very rapidly in this region; and with high input signal levels, i-f harmonics may well be stronger than the i-f fundamental. Besides, to measure closer than 10 MHz to the *LO*, the *dv/dt*-sensitive ECL prescaler in your counter will have to be bypassed. Some ECL devices become very confused with slowly rising and falling wavefronts.

other uses

The heterodyne scaler has other uses besides accurate frequency measurement in the lower microwave bands. Most notable among these is its use as a receiving converter for 1296 or 2304 MHz (or anywhere in between). This project was designed as a high-level mixer, so I made no attempt to characterize *LO* noise skirts or system noise figure. Because *LO* noise drops as you move farther away from the *LO* frequency, I would guess receiver noise figure would be best between 600 and 900, 1100 and 1400, 1650 and 1850, and 2150 and 2350 MHz. In a nutshell, my advice to anyone using this circuit as a receiving converter is to use a high gain antenna, a 3 dB (or better) noise figure preamp, mast-mounted if possible, and a feedline such as 3/4-inch (19-mm) hardline.



Close-up of scaler, scaler pickup antenna, and a microwave oscillator under test. Note polarization of oscillator.

Another possible use of the heterodyne scaler is a stable 2.000-GHz *LO* source for a transmitting converter. For this you need a 432-MHz transmitter, mixer, and a preamplifier-power amplifier chain.

If you have any problems, questions, or comments concerning the scaler, or live around Chicago and would like to attempt communications at or above 2300 MHz, please get in touch with me.

addresses of electronic parts manufacturers

Texas Instruments
Semiconductor Components Division
P.O. Box 5012
Dallas, Texas 75222

Engelman Microwave
Skyline Drive
Montville, New Jersey 07045

Solid-State Microwave
Montgomeryville, Pennsylvania 18936

Plessey Semiconductor Products
1674 McGraw Avenue
Santa Ana, California 92705

Fairchild Camera and Instrument Corporation
464 Ellis Street
Mountain View, California 94042

addresses of rf PC-board manufacturers

Oak Materials Group
Laminates Division
174 North Main Street
Franklin, New Hampshire 03235

Cincinnati Millaron
Molded Plastics Division
Blanchester, Ohio 45107

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KENWOOD
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15 memories/offset recall, scan, priority, DTMF

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TR-7800

Kenwood's remarkable TR-7800 2-meter FM mobile transceiver provides all the features you could desire for maximum operating enjoyment. Frequency selection is easier than ever, and the rig incorporates new memory developments for repeater shift, priority, and scan, and includes a built-in autopatch DTMF encoder.

TR-7800 FEATURES:

- **15 multifunction memory channels, easily selectable with a rotary control**

  M1-M13 ... memorize frequency and offset (±600 kHz or simplex).

  M14 ... memorize transmit and receive frequencies *independently for nonstandard offset.*

  M0 ... priority channel, with simplex, ±600 kHz, or nonstandard offset operation.

- **Internal battery backup for all memories**

  All memory channels (including transmit offset) are retained when four AA NiCd batteries (not Kenwood-supplied) are installed in battery holder inside TR-7800. Batteries are automatically charged while transceiver is connected to 12-VDC source.

- **Priority alert**

  M0 memory is priority channel. "Beep" alerts operator when signal appears on priority channel. Operation can be switched immediately to priority channel with the push of a switch.

- **Extended frequency coverage**

  143.900-148.995 MHz, in switchable 5-kHz or 10-kHz steps.

- **Built-in autopatch DTMF (Touch-Tone®) encoder**

- **Front-panel keyboard**

  For frequency selection, transmit offset selection, memory programming, scan control, and selection of autopatch encoder tones.

- **Autoscan**

  Entire band (5-kHz or 10-kHz steps) and memories. Automatically locks on busy channel; scan resumes automatically after several seconds, unless CLEAR or mic PTT button is pressed to cancel scan.

- **Up/down manual scan**

  Entire band (5-kHz or 10-kHz steps) and memories, with UP/DOWN microphone (standard).

- **Repeater reverse switch**

  Handy for checking signals on the input of a repeater or for determining if a repeater is "upside down".

- **Separate digital readouts**

  To display frequency (both receive and transmit) *and memory channel.*

- **Selectable power output**

  25 watts (HI)/5 watts (LOW).

- **LED bar meter**

  For monitoring received signal level and RF output.

- **LED indicators**

  To show: +600 kHz, simplex, or -600 kHz transmitter offset; BUSY channel; ON AIR.

- **TONE switch**

  To actuate subaudible tone module (not Kenwood-supplied).

- **Compact size**

  Depth is reduced *substantially.*

- **Mobile mounting bracket**

  With quick-release levers.

See your Authorized Kenwood Dealer now for details on the TR-7800 ... the remarkable 2-meter FM mobile transceiver!

NOTE: *Price, specifications subject to change without notice and obligation.*

MATCHING ACCESSORY:

- KPS-7 fixed-station power supply

TRIO-KENWOOD COMMUNICATIONS INC.
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Easy tuning, digital display, professional quality

R-1000

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. Its easy-single-knob tuning and high sensitivity, selectivity, and stability make the R-1000 a favorite amongst Radio Amateurs, shortwave listeners, engineers, maritime communicators, and others who demand high quality in a general-coverage communications receiver.

R-1000 FEATURES:

- **Continuous frequency coverage from 200 kHz to 30 MHz**
  Receives shortwave, medium-wave, and long-wave bands.
- **30 bands, each 1 MHz wide**
  Easy-to-use band switch with large knob.
- **Five-digit frequency display and analog dial**
  Accurate digital display with 1-kHz resolution and illuminated analog dial with precise gear dial mechanism.
- **Built-in quartz digital clock with timer**
  Precise 12-hour clock with AM and PM indicators. Timer turns on radio for scheduled listening, and even controls a recorder through remote terminal.
- **Up-conversion PLL, wideband RF circuits**
  Provide exceptional performance and easy operation without the need for bandspread, preselector, or antenna tuning. Excellent sensitivity, selectivity, and stability.
- **Three IF filters for optimum AM, SSB, CW**
  12-kHz and 6-kHz (adaptable to 6-kHz and 2.7-kHz) filters for AM wide and narrow, and 2.7-kHz filter for high-quality SSB (USB and LSB) and CW reception.
- **Communications-type noise blanker**
  Eliminates ignition and other pulse-type noise. Superior to noise limiter.
- **Step attenuator**
  0-60 dB in 20-dB steps. Prevents overload.
- **Recording terminal**
  For external tape recorder.
- **Tone control**
  For desired audio response.
- **Built-in 4-inch speaker**
  For quality sound reproduction.
- **Dimmer switch**
  Controls S-meter and other panel lights and digital-display intensity.
- **Three antenna terminals**
  Wire terminals for 200 kHz to 2 MHz and 2 MHz to 30 MHz. Coax (SO-239) terminal for 2 MHz to 30 MHz.
- **Selectable operating voltage**
  AC voltage selector for 100, 120, 220 and 240 VAC. Also adaptable to operate on 13.8 VDC. (With optional DCK-1 kit.)

Ask your Authorized Kenwood Dealer about the easy-to-operate R-1000 communications receiver.

NOTE: *Price, specifications subject to change without notice and obligation.*

MATCHING ACCESSORIES

- SP-100 external speaker
- HS-5 deluxe headphones

Other accessories not shown:

- HS-4 headphones
- DCK-1 easy-to-install modification kit for 12-VDC operation

HC-10 Digital World Clock

- **Two 24-hour displays with quartz time base**
  Right display: local (or UTC) hour, minute, second, day. Left display: month, date, world time in various cities, memory time (QSO starting time), and time difference (in hours from UTC)
- **Time in 10 cities around the world**
  Plus two additional programmable time zones.
- **"TOMORROW" and "YESTERDAY" indicators**
- **Memorizes present time**
  And recalls later, for logging purposes.
- **High accuracy**
  ±10 seconds/month

variable-inductance variable frequency oscillators

A comprehensive discussion of VFO circuits including a variometer VFO, an iron-vane VFO, and copper-vane VFO

This article deals with a number of variable-frequency oscillators, which I developed over seven or eight years, mainly for portable use.

Inductance-tuned VFOs are not a new idea. Long before the days of solid-state VFOs some manufacturers were using them in commercial ham gear. The Collins PTO (permeability-tuned oscillator) is an example. The original PTO was a precision-built factory product, but the inductance-tuned VFOs described here are simple enough to be built by any construction-minded Amateur. This might no longer be the case, however, if it becomes impossible to obtain good dial drive mechanisms through retail sales outlets.

The importance of simple VFO design may seem questionable at a time when the industry appears to be rushing into synthesizers. The answer is evident when one notes that data in an advertisement by a leading manufacturer indicated that its latest transceiver drew over an ampere from a 12-volt source in "receiving standby" position. In contrast, the receivers I've designed for portable use have drawn less than 50 milliamperes; something of the order of 10 milliamperes at 9 volts is possible with direct-conversion designs. Thus there continues to be a need to develop VFOs that are smaller, more economical, and more stable.

The reasons for the variable-inductance approach for both portable- and fixed-station use are these:

1. An inexpensive variable capacitor may cause noise and frequency-jitter when tuned.

2. Even a good variable capacitor, if at all available, may be large and heavy by modern standards, besides being expensive.

3. A variable-inductance tuning system can be low cost.

4. Variable-inductance tuning lends itself best to bandspread tuning, as in the Amateur bands.

5. Vernier dial drives that ensure backlash-free tuning are available from Japan and also from England (Jackson Bros.).

design considerations

The basic circuits used in my experiments are the series Colpitts (originally known as the Clapp oscillator), **fig. 1**, and, in one instance, the Hartley oscillator, **fig. 2**. The Seiler circuit, not used, is a modification of the series Colpitts. It is, of course, desirable to make the oscillator frequency as stable as possible consistent with cost and size.

One source of frequency instability is the active element, the transistor, which is much more unstable with current and temperature changes than a vacuum tube. At a given frequency, the transistor equivalent internal capacitances change with power-supply voltage and biasing changes, thus causing frequency shift. Temperature changes have the same effect. One source of temperature change is the dc operating currents and the other is any change in the ambient (i.e. surrounding) temperature.

By Richard Silberstein, WØYBF, 3915 Pleasant Ridge Road, Boulder, Colorado 80301

Oscillator stability. An important factor in ensuring oscillator stability is to make the frequency much dependent on the LC circuit and little dependent on the transistor parameters. The usual way to do this is to start with a high-Q LC resonant circuit and couple it loosely to a very low impedance circuit — actually to two low impedance circuits, one for actuating the transistor, and the other at 180 degrees phase reversal for feedback from the transistor.

I've usually worked with fets because, like vacuum tubes, they have high input and output impedances, whose changes with temperature and voltage variation will have a minimum effect on the low impedances with which they are made to interface. The fets I've used have usually been N-channel, dual-gate, gate-protected mosfets such as RCA's 40673 or 3N211, because the impedances and mutual conductances are higher in mosfets than in jfets. No information in the Motorola HEP literature is given as to whether their mosfets are gate-protected. Older fets such as RCA's 3N128 and 3N140 are not gate protected and thus are very hard to use in this application.

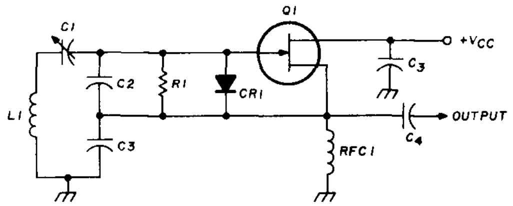


fig. 1. Series-Colpitts oscillator. C1 for tuning. Feedback ratio is determined by C2 and C3. C4 is an rf bypass capacitor and C5 a coupling capacitor.

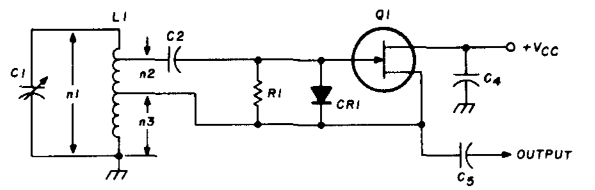


fig. 2. Hartley oscillator. C1 is for for tuning. Feedback ratio is determined by n2 and n3. C2 and C4 are coupling capacitors and C4 is an rf bypass capacitor.

Power source. To further minimize the effects of transistor instability in small portable equipment, I believe it's most economical of space and battery drain to power each oscillator with its own 9-volt battery rather than use elaborate voltage regulators with a single, common power source. Note that a battery has internal resistance, so that attempts to power another circuit from the same battery will result in frequency shifts whenever the latter circuit's current drain is changed. This effect results from the additional voltage drop in the battery's internal resistance, caused by the current to the second circuit. Besides rf shielding can be made very effective when the battery is kept in the shield box.

Battery current. If the battery current, which is normal for oscillator operation, is too high, oscillator frequency will drift after the voltage is applied, while the transistor is stabilizing to a slightly higher internal temperature. In experiments with mosfets a number of years ago I found that, in high-frequency oscillators, the drain current should not exceed about 2 milliamperes. Much lower values than this can be used to advantage, except that a too-low value can produce a noisy circuit or a hard-starting oscillator.

The current can be decreased by decreasing the voltage on the agc gate of a dual-gate mosfet (G2 in the 40673). It can also be controlled by changing the feedback ratio, which is C2/C3 in **fig. 1**, or n2/n3 in **fig. 2**.

The relationship of feedback-ratio-to-performance is not simple, but fortunately a 1:1 feedback ratio is frequently best. If the resonant-circuit Q is high enough, isolation can be improved by making C1 relatively small compared with C2 and C3 (**fig. 1**) or by placing the high n2 tap considerably below the top of L1 in **fig. 2**. Carrying this procedure too far produces high drain current, noise, parasitic oscillations, or no oscillation.

Oscillator keying. If you wish to key a transmitter in the oscillator circuit, a low-current design is desirable to minimize chirp caused by transistor heating when the key is depressed. Conditions for low current frequently mean closer coupling of the transistor to the LC circuit, so that there is a limit to the effectiveness of this way of reducing chirp (e.g., making C3 in **fig. 1** too small).

On the other hand some designs with good isolation diminish chirp but introduce slow drift because of high current (e.g., making C3 in **fig. 1** too large). However, another factor makes direct keying undesirable. A key-click filter usually entails RC circuits with slow time constants. Charging and discharging of the capacitors produces transients in the dc voltages across the transistor terminals, causing frequency shift and, hence, chirp, as the voltage rises and falls.

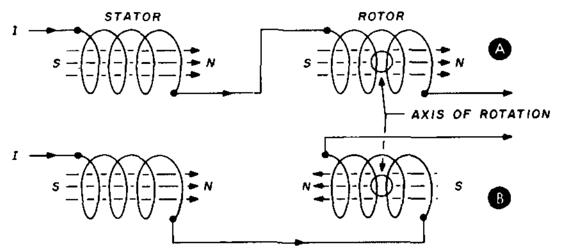


fig. 3. Principle of variometer. (A) fields aiding; (B) fields opposing.

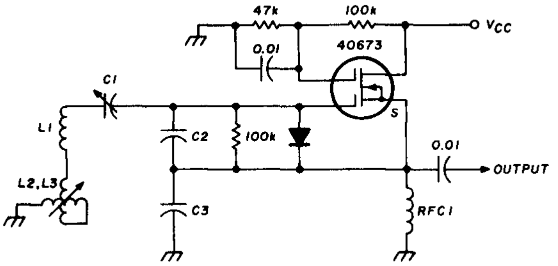


fig. 4. Variometer-tuned series-Colpitts VFO. Tuning is by L2L3. C1 is for calibration. Feedback ratio is determined by C2 and C3.

Temperature effects on inductors. In the matter of ambient temperature effects, it has been pointed out that even powdered-iron toroid inductors have a temperature coefficient too great for many purposes.[1] However, I've used these inductors for compactness in the VFOs of portable QRP equipment and have found that they did not drift too rapidly when the equipment was used under a canopy on a ship in the tropics, even with the sun setting.[2] Outdoor operation in a simple metal box in the sunlight is more difficult.

Air-core toroids. Ferrites have much worse temperature characteristics than powdered iron, so almost the only alternative is an air-core coil. However, there's still the issue of what kind of air-core coil to use. A number of years ago I evolved a simple construction procedure for making *air-core toroids*. These are much larger than their powdered-iron equivalents, but they still can be confined to a small shield box; since a toroid's magnetic field is closely contained, the shield box has little effect on inductance and *Q*, producing mainly an increase in capacitance.

One of my early air-core toroids had a *Q* of about 300 and was used in a signal generator tuning from 12 to 23 MHz. With a 200-milliwatt final amplifier preceded by a buffer, it was possible to key the oscillator and transmit on the fundamental frequency at 14 and 21 MHz without very bad chirp. But key clicks were

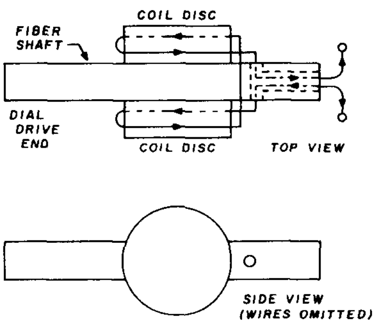


fig. 5. Split-rotor coil construction.

bad, so that direct keying became undesirable, as discussed above.

Temperature stability of air-core toroids built for a recent series of tests was disappointing, as described further on. Perhaps the toroidal shape itself produces a high temperature coefficient when subjected to the expansion and contraction of copper and plastic.

Capacitors. Capacitors are temperature sensitve too. In frequency-determining circuits, experimenters have found the small polystyrene types to be the best. Both DeMaw[1] and Eaton[3] correctly note that overall frequency drift can be compensated for by experimenting with capacitors of different temperature coefficients, but that this is a difficult process, the results of which might be hard to duplicate. And enclosing the oscillator in a constant-temperature oven, an obvious alternative, would not be feasible for small, portable equipment.

In addition to being temperature sensitive, certain capacitor types are highly sensitive to humidity. Mica compression trimmers in rf circuits have been known

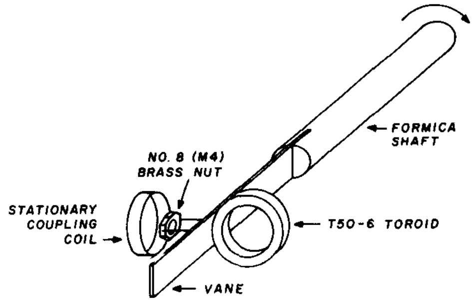
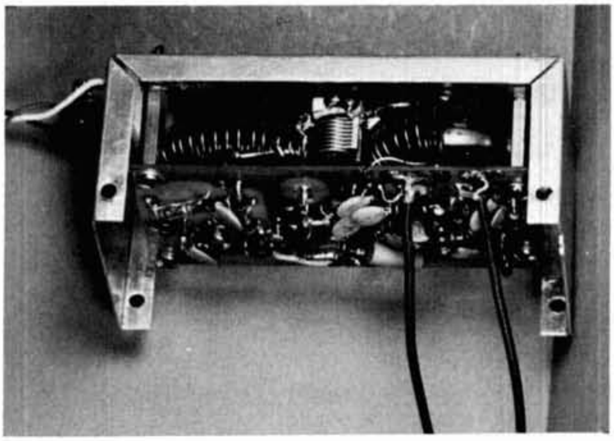


fig. 6. Tuning couplet. Vane is of epoxy circuit-board material. Leave two small copper spots for the supporting wires. Alternatively, support nut with insulating spacer block. Hold T50-6 toroid to vane with cyanoacrylic cement.

for many years to have some of the properties of barometers. It used to be considered a cute trick to place a sensitive back-biased vacuum-tube voltmeter across the output of an i-f stage and watch the effect of blowing one's moist breath through a glass tube onto a tuning capacitor.

Even ceramic trimmers leave a lot to be desired, which makes it appear that adjustable air-dielectric trimmers might be the best for stable oscillators. Some very small air trimmers, with very close plate spacing, are available. Having once worked with installations on yachts, I wonder if the spacing in some capacitors is small enough that precipitation, which sometimes forms when hot, moist air cools on a metal surface, might not short-circuit capacitor plates; i.e., is a "dew-drop shunt" a possibility, especially if the dew drop happens to be salty?

Shielding. In small, portable equipment, a limited amount of frequency drift can be corrected by resetting the oscillator to a reference calibration point, as in zeroing the VFO to a frequency marker. The important thing here is to be sure that the VFO does not drift appreciably during a contact. To obtain improved results with a small increase in size, the temperature of the oscillator can be made to change slowly under conditions where the oscillator might have drifted rapidly, as when moving equipment outdoors into the sun or into a chilly breeze. This can be done by using both an internal and external shield box. Such an arrangement produces double rf shielding, which not only decreases rf interference locally, but also is one way of reducing feedback from the higher stages of the transmitter to the VFO, which may be a source of instability. For best rf shielding, as in rf screened rooms, grounding the boxes to each other should be at *one* point, with precautions in treating rf cables and power leads the same as in the screened-room case.

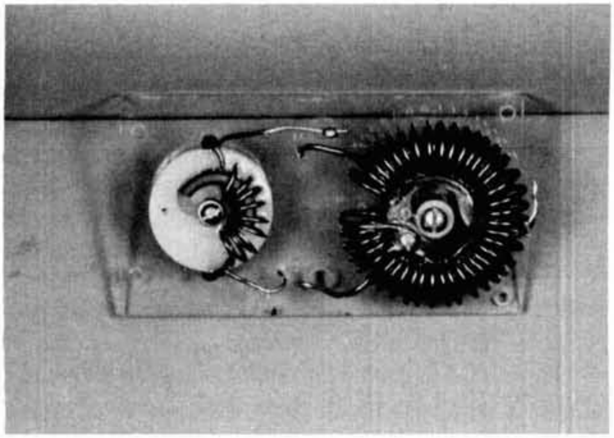


Coil board for VFO H-2. Air-core toroid and iron-vane tuning unit are shown in position.

the variometer VFO

In the early 1920s the variometer was used to tune stages of the then-prevalent trf (tuned radio-frequency) broadcast receivers. It consisted of one stationary coil, inside of which was one rotatable coil. **Fig. 3** shows the principle, with the rotor shown beside the stator for illustrative purposes. Drawing **A** shows the rotor in the "fields-aiding" position. The magnetic fields of the two coils are mutually aiding, which results in a larger value of inductance than the sum of the two. In the "fields-opposing" position, **B**, the circuit inductance becomes small.

Commerical variometers were made with split stators and rotors, which were shaped to ensure the maximum coupling of electromagnetic field lines, and thus the maximum range of inductance values,

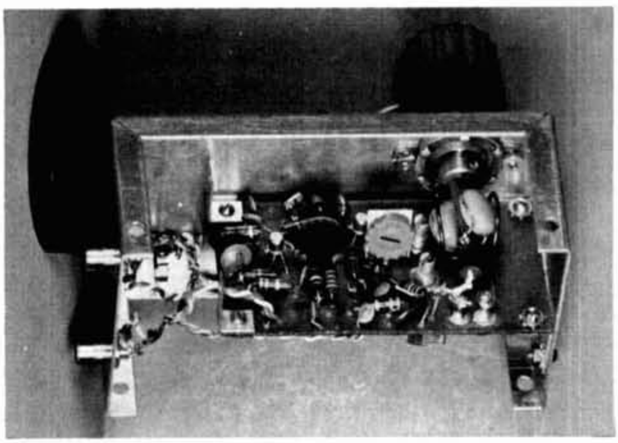


VFO H-2 assembly. Air-core toroid and copper-vane tuning unit are visible under circuit board.

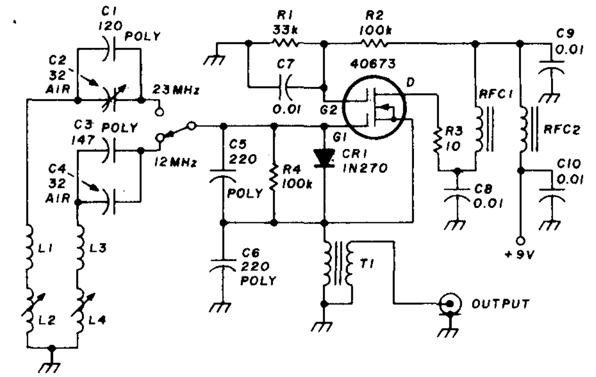
as the rotor was turned 180 degrees inside the stator. The main problem with the variometer is that most of the circuit resistance remains unchanged for all rotor positions, so that the Q becomes very small at one end of the tuning range. This results in broad tuning and low gain at the high-frequency end of the dial in trf receivers. However, for bandspread tuning, which is required for covering only a few hundred kHz in an oscillator, a small variometer can be placed in series with the main inductor, so that the worst-Q position of the rotor will have only a small effect on the circuit Q.

Fig. 4 shows a series Colpitts circuit tuned by variometer L2L3 in series with L1. Capacitor C1 can be used for calibration.

Construction. Fig. 55 shows how the split rotor in the experimental model was built. The tuning shaft is a 1/4-inch (6.35-mm) phenolic rod. Cemented to its sides are two circular disks of insulating material,



Variometer VFO. Variometer tuning unit is in corner attached to Jackson dial drive behind tuning knob.



C1	120-pF polystyrene
C2,C4	32-pF air trimmer
C3	147-pF polystyrene
C5,C6	220-pF polystyrene
C7-C10	0.01μF
L1	9-1/2 turns no. 18 (1-mm) wire on T50-6 core
L2	1-1/2 turns no. 16 (1.3-mm) wire close to couplet
L3	21 turns no. 20 (0.8-mm) wire on T50-6 core
L4	4-1/2 turns no 18 (1-mm) wire close to couplet
T1	primary 19-1/2 turns no. 30 (0.25-mm); secondary 4 turns no. 24 (0.5-mm) wire on FT37-63 form or equivalent
RFC1, RFC2	23-1/2 turns no. 32 (0.2-mm) wire on FT-37-63 form or equivalent

fig. 7. Couplet-tuned VFO for 15- and 20-meter receiver with 9-MHz i-f.

each carrying a one-turn coil section. Leads from the rotor go through a drilled portion of the shaft to the rear, and come out to stationary terminals in the rear. These leads are twisted, plastic-covered, stranded no. 24 (0.5-mm) wire. They emerge at terminals close behind the shaft to minimize the adverse effect of continual flexing. The dial drive is at the front of the shaft. The spring effect of the twisted wires was enough to rotate the shaft against dial friction. (This could have been eliminated by a friction bearing as described later.)

The disks of insulating material in the rotor could be replaced by powdered-iron cores to some small advantage. The split stator consists of two self-supporting, one-turn, coils connected in series with the rotor, and into which the rotor meshes.

The variometer-VFO design illustrates the principle. The oscillator was never used, because better approaches became evident. However, it did serve as a stimulus for thinking about inductance tuning. In other respects, this particular oscillator might be called the "nostalgia special."

Performance. In all-inductance tuning, as you might guess from simple considerations, for a fixed ratio of inductance change to total circuit inductance, the frequency coverage or bandspread achieved decreases directly as frequency decreases. This is accurate when the bandspread is a small fraction of the main frequency; it becomes less accurate as the fraction increases. Design for the lower high frequencies suffers from this frequency effect, especially at 3.5-4.0 MHz, where a large bandspread may be desired.

a rotating-couplet VFO

This design allowed the most compact construction of all that were built. The heart of the unit is a "rotating-couplet," whose design I evolved from recollection of an aligning tool of the 1930s. This magic wand had a powdered-iron slug at one end and a brass slug at the other end. Inserting the iron slug into an rf solenoid coil increased the coil inductance. If this act tended to increase the receiver output at a given frequency, the rf coil needed more inductance; a decrease meant it needed less. The brass slug worked in the opposite manner, since induced eddy currents in the brass reduced the coil inductance.

Construction. The couplet shown in **fig. 6** consists of a powdered-iron toroidal core (I used T50-6) placed opposite a brass slug and in the same plane. The whole couplet is rotated by a formica shaft. At one extreme of a 180-degree rotation, the toroid is coupled closely to a stationary coil, increasing its inductance a maximum amount. Rotation away from the maximum position decreases the effect of the iron. At the 90-degree point, the coil inductance should be the same as if the coil were alone. Past the 90-degree point, the brass slug rotates into the coil, reducing its inductance. An 8-32 (M4) brass nut was used as the slug. The core and the slug are mounted

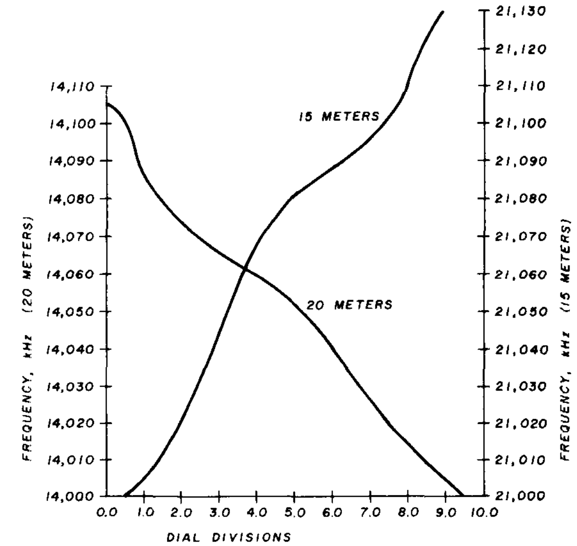
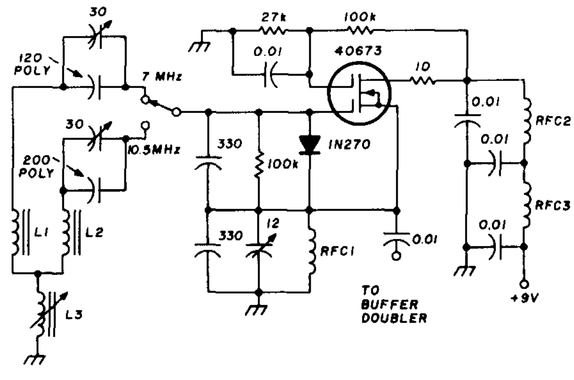


fig. 8. Frequency calibration of receiver using couplet-tuned local oscillator.



L1	33-1/2 turns no. 26 (0.3-mm) wire on T50-6 form
L2	18-1/2 turns no. 22 (0.6-mm) wire on T50-6 form
L3	see text
RFC1, RFC2	Miller 70F475A1
RFC3	Miller 70F125A1

fig. 9. Iron-vane VFO for 15- and 20-meter transmitter.

on a small Fiberglas epoxy vane made from a piece of circuit board from which most of the copper was removed. The nut is supported on wires soldered to two small copper spots remaining on the vane. (One large spot might have changed the coupling pattern.) A better method would have been to mount the nut in its offset position by using a small plastic block.

For homebrew equipment, when multiband operation is desired, it's probably better to go back to early commercial practice and use a bandswitching VFO, since the penalty of separate frequency calibrations for each band may be less than that of using an extra mixer stage with frequency conversion and the accompanying difficulty of suppressing additional unwanted frequencies.

Fig. 7 is a circuit diagram of a VFO used as an LO in a 15- and 20-meter bandswitched portable CW receiver with a 9-MHz i-f. For 20-meters the oscillator operates in the 23-MHz region and for 15-meters in the 12-MHz region. This is the LO used in my "Minicruiser" receiver.[2]

Two switched coupling coils, L2 and L4, are placed on opposite sides of the rotatable couplet. Each is in series with a main inductance coil. These coils are L1 and L3.

Performance. The calibration curves of the receiver using this oscillator are shown in **fig. 8**; they are obviously far from linear. Even worse, in the 90-degree region such an arrangement can produce a small but annoying defect. As the powdered-iron core rotates past the minimum-effect position at 90 degrees and the brass slug rotates in, the iron slug may momentarily increase the coil inductance again to a greater extent than the brass slug decreases it, producing a tiny doubling back of the calibration curve. Similarly, the effect of the brass slug may predominate over a small interval on both sides of 90 degrees.

an iron-vane VFO

This VFO has some similarity to the previous one, except that the variable inductor in series with the main inductance consists of a semi-circular coil and a rotatable powdered-iron vane of the same general type as that originally designed for a QRP Transmatch by W1CER and K1KLO.[4]

Fig. 9 is the circuit for a two-frequency switched VFO used in the transmitter of the *Minicruiser.*[2] The generated frequencies are near 7 and 10.5 MHz, to be doubled to cover the CW frequencies near 14 and 21 MHz. **Fig. 10** shows some design details of an iron-vane tuning unit similar to the one used here.

Construction. I made the iron vane from a toroidal core 1 inch (25.4 mm) in diameter, of material said to be similar to Micrometals T2. I sawed the core in half and cemented it to a sickle-shaped flat plate from which the copper had been etched. Shaping the sickle was easy. I cut an approximate disk from the circuit-board material and turned it in a lathe to a diameter of 1 inch (25.4 mm), although without a lathe a tedious sawing and filing job would have been satisfactory. The material between the sickle and the "hub" was removed with bandsaw files and occasionally a small hacksaw.

The best adhesive for fastening the powdered-iron segment to the sickle was cyanoacrylic (like Eastman 910). To my surprise, epoxy worked badly, perhaps because it did not adhere to the lacquer on my particular core segment. Rubber cement was even better. The assembly withstood a voyage around South America and a truck trip from San Francisco to Denver.

I screwed the now-finished vane to the end of a 1/4-inch (6.35-mm) Formica shaft, which had been drilled and tapped for a 6-32 (M 3/5) screw. I made the shaft long enough to fit into the dial drive. A

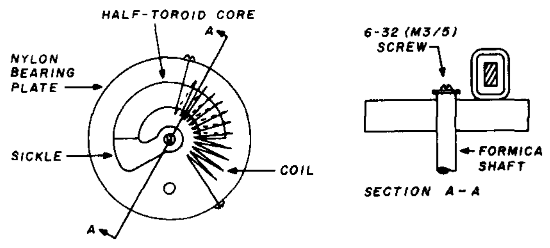


fig. 10. Design features of an iron-vane tuning unit. This unit was used in some tests of VFO H-2 in lieu of the copper-vane tuner of figs. 12 and 13.

good fit, essentially free of backlash, was made by drilling and tapping the dial drive collar for 4-40 (M3 screws with the shaft already inserted, using two screws at right angles and set at different lengths along the shaft.

To eliminate wobble of the iron vane as it moved in the pickup coil, I devised a bearing made of a circular piece of solid nylon. The material flows when hot, so the shaft hole had to be drilled with a 1/4-inch (6.3-mm) drill and then filed carefully for a tight fit to the Formica shaft. Very slow rotation, as in a back-geared lathe, can eliminate this problem. The nylon material provided a tight grip on the shaft, yet did not produce frequency jitter as the dial was turned.

In the design for the circuit of **fig. 9**, I wound the semicircular coil, L3, with five turns of no. 18 (1 mm) wire, using a rectangular mandrel 1/4 × 1/2 inch (6.35 × 12.70 mm) and cemented it, using epoxy, to the outside circumference of a small nylon bearing in the region where the bearing was mounted to an acrylic plate. In the design of **fig. 10**, I merely cemented the coil to the flat surface of the fairly large bearing shown, using Duro plastic-mending cement. I found it desirable to make the coil cover less than 180 degrees of arc, otherwise the end turns would act as stops (detents) as the core was rotated, with considerable potential for damage.

Performance. Fig. 11 is a set of calibration curves for the iron-vane VFO of **fig. 9** at the frequencies observed after doubling. They are quite linear over a fairly wide range. Irregularities are no doubt caused by imperfections in the hand-wound variable inductance.

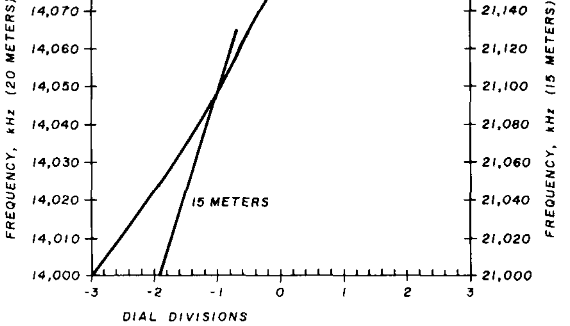


fig. 11. Calibration of "Minicruiser" iron-vane transmitter VFO.

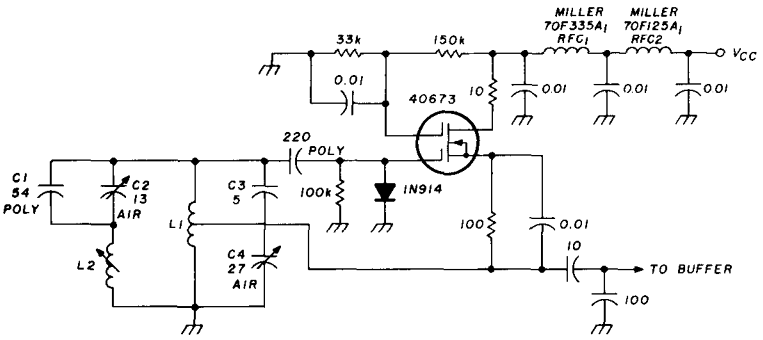


fig. 12. Circuit of inductance-tuned Hartley VFO H-2. L1 and L2 are described in text. C4 is for calibration setting.

a copper-vane Hartley VFO

In the design of the iron-vane VFO of **fig. 9**, the main inductance coil for each frequency is a powdered-iron toroid. Hoping to increase the temperature stability of the system, I decided to sacrifice some compactness and eliminate powdered-iron cores altogether. The main inductance coil now became an air-core toroid. The series tuning coil became an air-core half-toroid into which was rotated a sickle of copper tubing which, through eddy-current action, decreased the inductance (and also the Q) of the semi-toroid as more of the sickle was rotated into the coil. The effect was the opposite of that experienced with the iron-vane VFO. As a further innovation I decided to use the simpler Hartley circuit of **fig. 2. Fig. 12** is the circuit of my test VFO H-2. The earlier H-1 was similar.

Construction. For coil L1, I had evolved a simple means of making an air-core toroid as mentioned above. Coil of the order of 2 inches (51 mm) in diameter demonstrated Qs of the order of 180-200 at frequencies from 5 to 15 MHz, with shunt capacities of 60-100 pF.

Coils L1 and L2 of **fig. 12** are shown mounted on an acrylic coil board in **fig. 13**; their construction is described below.

First I constructed a spool consisting of two 1-inch (25.4 mm) acrylic disks separated by a cylindrical fiber spacer 5/16 inch (8 mm) in diameter and 0.400 inch (102 mm) long. The spacer was drilled for a no. 8 (M4) brass screw for holding the assembly together, and for fastening to a mounting board. The disks were drilled and tapped for short no. 4 (M3) brass screws to serve as terminals.

I made the main toroidal coil, L1, by inserting a 1/2-inch (12.7-mm) dowel with screw terminals into a lathe chuck and winding forty two turns of no. 14 (1.6 mm) soft-drawn enameled copper wire onto the dowel. Upon release, the coil I finally made was 40-1/2 turns, which was then carefully wrapped around the spool and held with a rubber band.

The next step was to cement the coil to the spool, using epoxy, although plastic mending cement was later found to be easier to use and worked just as well. I used this coil in experimental VFO H-2 for 5-MHz tests. It was reduced to 38-1/2 turns for 9-MHz tests, the purpose being to produce an appreciable gap to isolate the high-impedance end from lossy and temperature-unstable materials.

I made L2, the half-toroid for tuning, in a similar way except that eighteen turns were used for VFO H-2. I mounted a nylon bearing with a 1/4-inch (63.5 mm) shaft hole flat on the Plexiglas coil board. The outer diameter of the bearing was 3/4 (190.5 mm) and the height 5/16 inch (79 mm). Then I wrapped the half toroid around the bearing to an angle slightly less than 180 degrees and also made it rest against the board. Next I cemented it into place. The copper-tube sickle was a half-circle of 1/4-inch (6.35-mm) tubing flattened at one end and offset from the shaft end by an insulated crank arrangement.

In some of the temperature tests described later, I replaced the air-core toroid of L1 by one of similar inductance and *Q*, consisting of a T50-6 core wound with twenty six turns of no. 24 (0.5 mm) enameled copper wire. Also, in some tests I replaced the copper-vane tuning unit at L2 with the iron-vane unit of **fig. 10**. Here coil L2 consisted of nine turns of no. 18 (1 mm) wire initially wound on a mandrel 1/4 × 3/8 inch (6.35 × 9.25 mm).

With different values of C1 I could make the VFO function at frequencies from roughly 4 to 12 MHz using essentially the same coil at L1, which was center-tapped. I did most of the experimenting near 9 MHz with C1 consisting of two 27 pF polystyrene capacitors in parallel.

Performance. Fig. 14 shows a frequency calibration

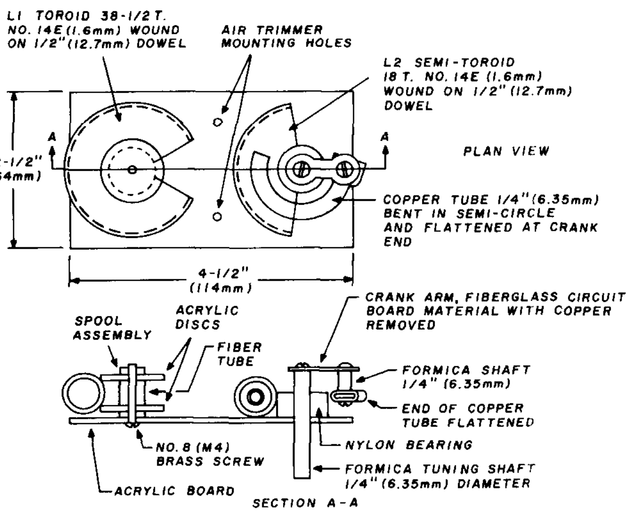


fig. 13. Coil board for copper-vane VFO. The etched circuit board is of the same size and is mounted above on 1-inch (25.4 mm) brass spacers.

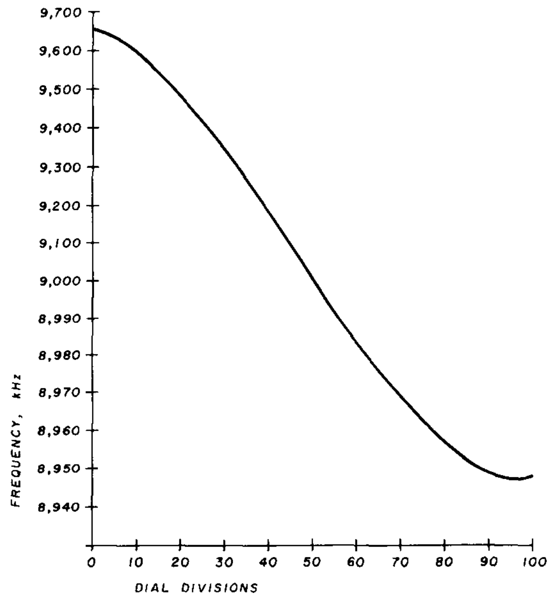


fig. 14. Copper-vane VFO H-1 frequency calibration.

of the copper-vane VFO H-1 around 9 MHz. The coils were similar to their H-2 counterparts, except that L1 had 41-1/2 turns and L2 had 17 turns. The linearity of most of the curve is evident. There was less success with temperature stability in this type of design, as described below.

temperature experiments

I had originally hoped to make readings of frequency drift with change of temperature of the VFO, but it soon became evident that this would not be an easy matter. Any kind of structure, whether a VFO or a house, when subjected to temperature changes acts like a radio circuit to the extent that it has a time constant. However, whereas radio circuits may have time constants as short as several nanoseconds or as long as several seconds, a house being heated or cooled may have a time constant of the order of a day and a half. A piece of radio equipment may require many hours to come to something approaching temperature equilibrium.

Heat transfer. Heat introduced into a VFO from the outside gradually travels to the inside. If the heating takes place inside an insulated box, with the room temperature and the heat source constant, everything eventually will reach an equilibrium temperature. The setup I decided on only approximated the ideal described above.

I mounted the VFO box on an electric plate warmer and covered it with a large chassis taped to the heating surface. Leads for the dc power and the temperature sensor were brought out at one corner. The rf output, after passing through a buffer stage, was brought through a piece of RG-174 cable at another corner and fed to a frequency counter.

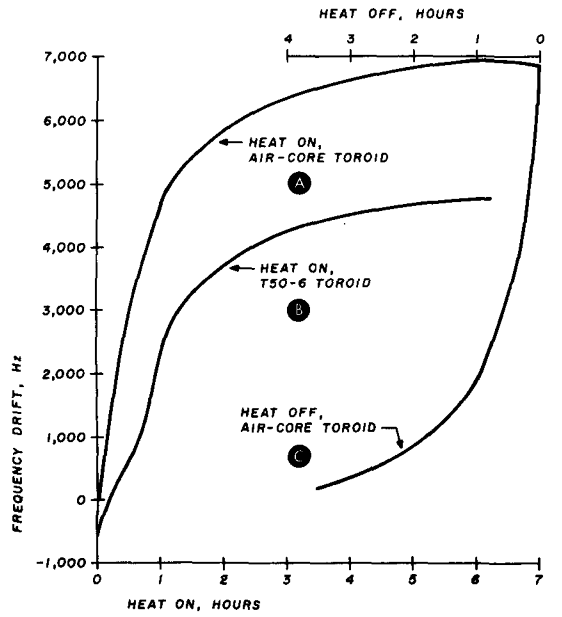


fig. 15. 5-MHz heat runs on VFO H-2 with copper-vane tuning unit.

Test procedure. For the temperature-sensing device, I used an IC made by Analog Devices, type AD590. This IC was fastened to the VFO box. The test procedure I used was to a) turn on the heat at the LOW setting when I thought the VFO to be in equilibrium with room temperature, and b) make periodic readings until the frequency stabilized. At this point, I assumed that the temperature of everything of importance inside the VFO was equal to the peak temperature reading of the run, about 34F (19C) above room temperature.

Some of the circuit components were undoubtedly more temperature sensitive than others and some might have even had temperature coefficients of opposite senses. There was, of course, no simple way of knowing when each circuit component reached what temperature, nor which ones were doing what during the transient warmup process. Besides, to reach a single equilibrium temperature took the better part of a day. The best thing I could do was to plot frequency drift *versus* time for the procedure described above and repeat the test under the same thermal conditions after making changes in the components in which I was interested. These components were usually L1 and L2 of **fig. 12**. In tests of both iron-vane and copper-vane tuning units, I always placed each vane in the position of closest coupling with its coil. I assumed that polystyrene capacitors and air trimmers were consistent in their temperature behavior, and spent only a little time on mica capacitors, which apparently differ one from another.[1]

Test results. In **fig. 15**, curve **A** shows frequency drift *versus* time at 5 MHz, with the 40-1/2-turn air-core toroid as L1 and the 18-turn copper-vane tuning unit as L2 (**fig. 12** circuit). C1 was a 220-pF polystyrene capacitor in parallel with one of 27 pF. The frequency drifted upward to 4700 kHz in the first hour, but it was not until the start of the sixth hour that the peak of 6900 Hz had been reached.

Curve **B** shows what happened after I turned the heat off and removed the VFO at the beginning of the seventh hour; the frequency decrease was very rapid at first. Since frequency drift is related to temperature, there was an obvious resemblance between the rise and fall of temperature and of current as the dc voltage in an inductive circuit is switched on or shunted out.

Curve **C** (**fig. 15**) is for the same oscillator but with the powdered-iron toroid in the L1 position. I was surprised to note that, after a small negative excursion, the frequency had drifted to only about 4700 Hz in six hours. There was later evidence, however, that the air-core toroid would have been more stable with a larger gap between the ends, although this did not make the air-core toroid competitive in the 9-MHz tests described below.

Fig. 16 shows the 9-MHz tests, with C1 at 54 pF. For the test of curve **A**, I reduced L1 to 38-1/2 turns, but the copper vane remained as before. In this test the frequency drift was 6350 Hz in fourteen hours and still rising.

For curve **B**, I used the T50-6 toroid in the L1 position, but continued to use the copper-vane unit at L2 as before. Here the frequency drift remained very small for the first twenty minutes, then went sharply negative to − 1750 Hz, then went up again, crossing the zero axis at about 4-1/2 hours and reaching 1665 Hz at 14-1/2 hours, where it was still increasing.

I next tried the T50-6 toroid in series with the iron-vane tuning unit of **fig. 10**. Curve **C** (**fig. 16**) shows the results. Again, there was an initial large swing in the negative direction, to − 2000 Hz in the first hour. Then the frequency began to increase with time, crossing the zero axis at 7 hours and peaking at 400 Hz at 13 hours 45 minutes.

Curves **B** and **C**, being more irregular than curve **A**, show some undetected mechanical instability, probably in the T50-6 toroid. However, the negative excursion and return of the two curves has some significance. Product literature issued by Micrometals for their toroids show curves of temperature coefficients *versus* temperature for all of their powdered iron, including no. 6 material. All have a transition from a positive to negative coefficient at 77F (25C). It's possible that, for curves **B** and **C**, the change of slope at about one hour occurs near where the T50-6 coil reached that temperature.

Fig. 15 and 16 are not directly comparable, because the air-core toroid was improved for the 9-MHz *heat runs, as described above.* Nevertheless, they do show that, contrary to expectations, the air-core toroids were not relatively immune to temperature changes. From the viewpoint of relative stability over hours, the combination of **C** might be termed best, with **B** next best. From the viewpoint of the first 20 minutes, **B** seems the most stable. However, the steep negative excursions of **B** and **C** are worse than the steep positive excursion of curve **A**, in the first few minutes after start.

In one brief test at 9 MHz, where I used all air-core coils and a 51-pF silvered-mica capacitor at C1, the temperature coefficient was negative. This test was discontinued for reasons mentioned previously.

Observations on coil construction. Thinking that the poor temperature behavior of the air-core toroid at L1 might be caused by the epoxy cement, I built an identical one using Duro plastic mending cement, which yielded essentially the same temperature curve. My guess was that perhaps the narrow diameter of the air toroid, relative to the wire diameter, was responsible for a greater proportionate reduction of the magnetic-flux-carrying cross-section area as the wire expanded with heat. In none of the tests did I try any *temperature-compensation techniques* because of difficulties in reproducing results when temperature-compensating components are introduced.[3]

conclusions

The Hartley oscillator uses fewer components than the series Colpitts, and may be adjusted more easily in an experiment. However it is possibly more prone to producing spurious oscillations and noise if not adjusted correctly. Apparently poor temperature characteristics of the Hartley oscillators with air-core toroids might be due to the fact that, because of spurious oscillations, it had not been possible to achieve isolation of the resonant circuit of **fig. 2** by bringing the n2 tap below the top of L1. In respect to isolation from temperature-sensitive transistor parameters, the Seiler circuit would probably have yielded the best temperature stability, with the series Colpitts as good but harder to adjust experimentally.

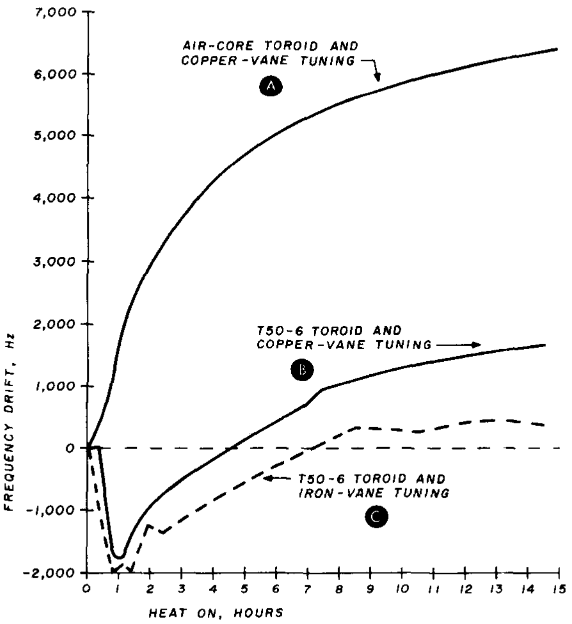


fig. 16. 9-MHz heat runs on VFO H-2.

Of the inductance-tuned VFOs the variometer VFO is an interesting antique; and the rotating-couplet VFO can be the most compact but yields a poor calibration curve. So far, air-core toroids are a disappointment in temperature behavior; one can make a practical VFO for portable use with an iron-core toroid and iron-vane tuning, *especially if a second* shield box is used. Frequent recalibration of a reference point against a simple crystal frequency standard is very helpful. Both iron-vane and copper-vane inductance tuning yield linear calibration curves.

My experiments over a number of years have demonstrated that the home constructor can build a useful VFO for either fixed-station or portable use without resorting to precision variable capacitors. In these experiments, frequency drift with temperature was greater than might be desired, but I believe that future experimenters should be able to determine whether improvements can result from the use of conventional air-core solenoid coils along with inductance tuning. Also, I regret that frequency-stability tests were not made on the Seiler or series-Colpitts circuit with inductance tuning. So there is plenty of opportunity for future learning.

The experimenter should certainly read the VFO references mentioned so far, besides many others, including Jim Fisk's early article[5] with its references going back many years. Bill Wildenhein's careful experiments[6] indicate that frequency drift can be conquered. In performing experiments, bear in mind one fact, which caused Reed Easton[3] to turn to a synthesis control technology: There are so many temperature-dependent variables that apparently identical VFOs may have different drift characteristics.

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ham radio

linear-amplifier cost efficiency

Consider gain per dollar rather than watts per dollar for a cost-efficient linear amplifier

Many Amateurs ask the question, "Should I invest several hundred dollars in a linear amplifier and what good will it do me?" The watts-per-dollar criterion is misleading because signal gain at the receiver does not increase linearly with power increase. A more valid measure would be gain per dollar. This article shows that the greatest signal gain-per-dollar cost is obtained at some easily calculated level of amplification. Amplification beyond this optimal level should be avoided by the cost-conscious Amateur.

background

It's well known that the increase in signal strength resulting from an increase in power is determined by the following formula:

$$dB = 10 \log \frac{P_2}{P_1} \qquad (1)$$

where P_1 is the initial power level output and P_2 is the increased power level output.

If we begin with $P_1 = 100$ watts output from the exciter and increase the power in steps of 50 watts, the increase in signal strength will be (for $P_1 = 100$ *watts*) as follows:

linear amplifier output (P_2)	increase (dB)	incremental increase (dB)
150	1.76	
200	3.01	1.25
250	3.98	0.97
300	4.77	0.79
350	5.44	0.67
400	6.02	0.58

An increase in output from 100 to 150 watts will increase the signal by 1.76 dB. Notice that as power increases the gain also increases but in *progressively smaller steps*. This is true for any level of power input and amplification. Each increase in power output results in a smaller increase in gain at an *increased cost* of energy and components. Clearly it's wise to stop short of the point where an increased cost would result in no significant incremental increase in signal strength.

efficiency

Most will agree that an increase in power will result in a greater cost of energy, components, insurance, space, and weight. We can define a cost-efficient amplifier as one that produces the strongest signal per unit power output, or the strongest signal per cost. In Pentagon terminology, a cost-efficient amplifier produces the "biggest bang per buck."

For example, if two signals of identical strength are generated from two amplifiers, the cost-efficient amplifier would be the one costing less. Another example is to compare the signal received from two amplifiers of equal cost. The one producing the greatest gain would be the most cost efficient.

Cost efficiency can be expressed mathematically:

$$efficiency = \frac{dB}{cost} \qquad (2)$$

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	DenTron 1200	DenTron Clipperton	ETO Alpha 76A	ETO Alpha 770X	Heath 201	Heath 221	Swan 1200Z	Swan 1500Z
power input (PEP)	1200	2000	2000	4000	1200	2000	1000	1500
gain (P_1 = 100W)	10.8	13	13	16	10.8	13	10	11.7
price (1979)	380	600	1495	3995	385	569	500	600
efficiency = $\frac{\text{gain}}{\text{price}} \times 100$	2.9	2.18	.87	.4	2.8	2.28	2	1.95

This article assumes costs increase proportionally but not linearly with the power level of the amplifier. I attempted to compare the selling prices of various amplifiers and to compute their relative efficiency. This proved to be a futile effort because prices are influenced by factors other than power, such as paying a premium for a name (Collins), paying for features not found on amplifiers of different manufacture, and not paying for costs of labor (Heath).

What could be done, however, is to compare amplifiers of the *same* manufacturer. This was done, and in every case the increase in power was accompanied by a commensurate increase in price and a decline in efficiency, where efficiency = gain divided by cost.

Above are calculations of comparisons of amplifiers manufactured by DenTron, ETO, Heath, and Swan which demonstrate this thesis.

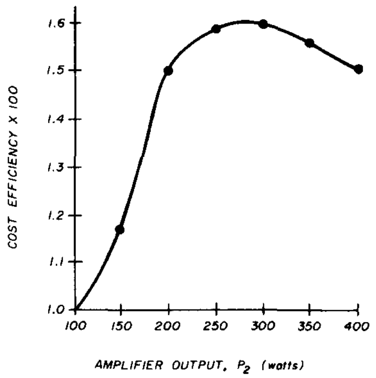

fig. 1. Cost efficiency obtained when increasing power-amplifier output. Best cost efficiency, in terms of costs of energy, components, insurance, space, and weight of the linear amplifier, occurs when amplifier output is about three times that of the exciter. In other words, $P_2 = 3P_1$ where P_2 and P_1 are the amplifier output and input power respectively.

where costs increase proportionally to power level. **Fig. 1** shows that the cost efficiency curve rises, levels off, and falls as a function of amplifier power. This is because of (a) the increasing costs that accompany an increase in power, and (b) the progressively smaller *incremental* increases in dB. Each additional dollar spent in power increases results in smaller and smaller signal increases. At some point the dB increase becomes insignificant but the costs continue to rise, resulting in reduced efficiency (*for* P_1 = *100 watts*):

linear amplifier output (P_2)	increase (dB)	cost efficiency × 100
150	1.76	1.173
200	3.01	1.500
250	3.98	1.590
300	4.77	1.590
350	5.44	1.550
400	6.02	1.500

An examination of **fig. 1** shows that optimal cost efficiency occurs when the linear amplifier output is approximately *three times* that of the exciter. Because we're speaking of the ratio of two power levels, optimal amplification of three times exciter input is true for all power levels.

summary

We've defined cost-efficient amplification as that which gives us the biggest bang per buck. This occurs when $P_2 = 3P_1$. This won't give you the strongest signal on the air but it will give you the *strongest signal per dollar*.

This definition of efficiency might be modified to consider other factors. For instance, a weight-efficient transmitter could be defined as one having the highest ratio of dB to weight, thereby being useful for comparing mobile or portable equipment.

The cost-conscious Amateur should consider a linear amplifier that increases exciter output three times — all other things, such as the antenna, being equal. After this point, diminishing returns occur.

ham radio

vhf techniques

improved accuracy when measuring small inductances with a dip oscillator

The time-honored method of measuring inductance has been to parallel the coil with a capacitor of known value, determine the parallel-resonant frequency with a dip oscillator, and calculate the inductance from the formula

$$L = \frac{1}{4\pi^2 f^2 C}$$

For small inductance values this method is, at best, approximate because of **a**) uncertain calibration of the oscillator frequency, **b**) normal tolerance of the capacitor value, and **c**) lead inductance of the capacitor, which adds to the inductance being measured. The uncertainty of **a** can be eliminated by coupling a frequency counter to the dip oscillator when obtaining a dip. The effect of **b** can be minimized by using a 1 or 2 per cent capacitor, or by actually measuring the capacitance on a bridge or equivalent instrument. The problem of capacitor lead inductance, which can amount to 10 to 20 nanohenries even with minimum lead length, is solved as follows.

If suitable capacitance-measurment equipment is available:

1. Prepare one side of a small piece of double-sided copper-clad board as shown in **fig. 1**.

2. Clip off the leads of a dipped mica capacitor of any convenient value; 100 pF is suggested.

3. Carefully crack the conformal case of the capacitor by gradually applying pressure in a bench vise and then picking off the case fragments; be sure not to loosen the end crimps of the denuded capacitor.

4. Place the capacitor on the prepared board so that it straddles the bared strip, with the remainder of the leads facing upward, and solder the end crimps to the copper cladding.

5. Clip the remaining leads from the capacitor. You now have a leadless capacitor, with only its structure contributing any inductance.

6. Measure and record the capacitance between the two isolated sections of the board. (The measured capacitance will be greater than the capacitor value by approximately one-half the capacitance between the back of the board and either top section.)

If you do not have access to capacitance-measurement equipment, follow steps **1** through **5** above, using a small piece of *single-sided* copper-clad board. Assume the final capacitance to be the same as that of the capacitor.

The coil to be measured can be soldered onto the board so that it bridges the bared strip, placing it in parallel with the capacitor with virtually no added series inductance. If you must use a single-sided board, be sure that it is placed on a non-metallic surface before dipping the L-C combination; otherwise an error will be introduced.

Purists may want to use a chip capacitor to further reduce the parasitic inductance, but this is probably overkill at frequencies where a dip oscillator can be used.

Wilkinson power dividers

One of the accepted techniques used by Amateurs in up-conversion to the vhf through microwave regions is that of utilizing the same local oscillator (LO) for both transmitting and receiving. This requires either a coaxial relay to switch the LO from transmit to receive or a power divider. If the LO has sufficient power output to drive both the transmit and receive mixers, the power divider is the simpler approach. Such dividers are available commercially, but their price and that of a coax relay are comparable.

An elegant solution, both technically and economically, is to use a Wilkinson power divider. This handy circuit may be configured from discrete components or from transmission-line sections; the latter can consist of either actual coaxial line sections or microstrip. Regardless of the physical realization of the divider, it will provide two isolated outputs, at each of which one-half of the input power is available.

Fig. 2 shows a Wilkinson power divider circuit using lumped components that is useful to several hundred megahertz. The output R_o is the impedance of

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each load (R_L) and that of the load impedance which the divider presents to the input source. The values of the components which make up the divider are calculated by using the following relationships:

$$C1 = \frac{1}{2\pi f R_o}$$

$$C2 = \frac{C1}{2}$$

$$L1 = L2 = \frac{R_o}{2\pi f}$$

$$R1 = 2R_o$$

where f is the operating frequency in megahertz
C is in microfarads
L is in microhenries.

For example, a 50-ohm system at 116 MHz, C1 is 27.4 pF, C2 is 13.7 pF, L1 and L2 are 68.6 nH, and R1 is 100 ohms. Five per cent mica capacitors of 27 and 13 pF may be used (although bridging to closer tolerance is desirable), and the coils may be wound and measured using the technique described earlier.

Normal vhf wiring practices of using short, direct leads should be followed. Space the two output connectors so that C2 and R1 may be connected between them with minimum lead length. There should be no mutual coupling between L1 and L2; at higher frequencies a shield between the coils may be necessary.

Between about 300 and 1000 MHz, the values of discrete components become too small to be realizable, especially the inductances, which tend to approach transmission lines. Therefore a coaxial transmission line version of the Wilkinson divider is preferable, although it may be used at any lower frequency, limited only by the amount of space physically required by the line sections. This configuration is shown in **fig. 3**. As in the lumped-constant version, $R1 = 2R_o$. The two coaxial transmission-line sections, $T1$ and $T2$, are identical and are one-quarter

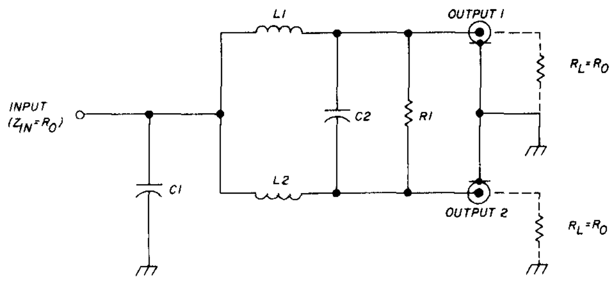

fig. 2. Wilkinson power divider using discrete components is useful below 300 MHz. Component values are discussed in the text.

wavelength long at the frequency of interest. The characteristic impedance, Z_t, of each line section is determined from the expression:

$$Z_t = 1.414R_o$$

and the electrical length, d_t, from

$$d_t = \frac{2951 V_p}{f} \text{ (inches) or } d_t = \frac{7495 V_p}{f} \text{ (cm)}$$

where f is the frequency in megahertz and V_p is the velocity factor of the cable.* For a 50-ohm system, the calculated value of Z_t is 70.7 ohms. Practically, any low-loss cable having a characteristic impedance between 70 and 75 ohms will prove satisfactory.

It is essential to ground the coax shields at both the input and output ends. If BNC connectors are used, they should be UG-260/U flange types rather than the UG-1094/U threaded variety. This will permit the transmission-line sections to lie closer to the connector mounting surface and reduce the length of the ground lead at each end. In addition, by closely spacing the flanges of the two output connectors, the spacing between the center pins will be such that a half-watt resistor can be placed directly on the mounting surface ground plane and soldered to the connector pins with minimum lead length.

At frequencies above 1 GHz, the circuit of **fig. 3** is easier to construct using microstripline instead of coaxial cable. The length and width of the transmission-line section will depend on the dielectric material and thickness in the copper-clad board. G-10 glass-epoxy board is is adequate below about 1.3 GHz, but becomes excessively lossy at higher frequencies, where Teflon or Duroid should be used. Reference 1 contains the necessary information to determine the transmisssion-line dimensions when microstrip on G-10 board is to be used.

Both the discrete component and coaxial line versions of the Wilkinson power divider have been built for use at frequencies between 100 and 408 MHz. In

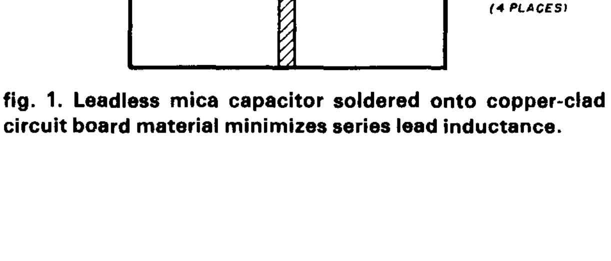

fig. 1. Leadless mica capacitor soldered onto copper-clad circuit board material minimizes series lead inductance.

* V_p is 0.66 for solid polyethylene dielectric, 0.78 for foam polyethylene, and 0.695 for Teflon.

each case the two outputs were balanced within 0.2 dB, and the isolation between output ports exceeded 20 dB. The VSWR looking into any port with the others terminated was 1.1:1 or less.

terminating double-balanced mixers

The use of the commercially packaged double-balanced mixers, such as those manufactured by Anzac, Mini-Circuits, Vari-L and others, has become commonplace in Amateur design. Their convenience and relatively good performance, insofar as intermodulation products are concerned, make them useful in both transmitting and receiving upconverters.

Without going into the details of intermodulation distortion, which has been covered extensively in published literature, [2,3] it is sufficient to state that the intermod specifications of double-balanced mixers can be attained only if at least two of the three mixer ports are terminated in 50 ohms at *all* frequencies. If only two ports are properly terminated, one of these must be the local oscillator input. Terminating the LO port is the easiest requirement to implement, since it entails only inserting a loss pad between the oscillator and mixer. A 2 or 3 dB loss pad is generally sufficient to provide a match, since the oscillator output impedance will normally have been optimized for maximum output into 50 ohms, thereby making its source impedance very close to that value.

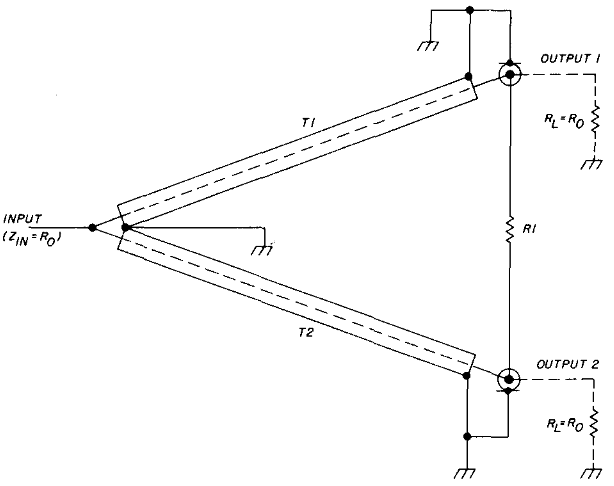

fig. 3. Transmission-line version of the Wilkinson power divider. Parameters of the coax sections are covered in the text.

Of the two remaining ports, the i-f output is the easier to match because it generally feeds an invariable, narrow-band load. Many techniques have been devised to ensure that the mixer i-f output sees 50 ohms at both the signal and the image frequencies, but most of them require two or more tuned circuits and 50-ohm terminating resistors. Another approach, suggested by Alan Podel (who designed many of the Anzac mixers), is to use a properly biased common-base amplifier following the mixer. This not only terminates the mixer output for all frequencies, but changes the gain of the mixer block from between −6 and −9 dB to one as high as +13 dB.

Fig. 4 shows a double-balanced mixer directly coupled to a grounded-base NPN amplifier. (Note that a loss pad is connected between the local oscillator and the mixer LO input.) The input resistance of a grounded-base amplifier is approximately $28/I_c$, where I_c is the collector current in milliamperes. Therefore if the collector current is set to 0.56 milliamps with the 5 kilohm potentiometer, the mixer will see about 50 ohms at all frequencies. The desired intermediate frequency is selected by means of the collector tuned circuit, *L1* and *C1*. A regulated 12-volt supply is necessary to maintain a constant base current, and the 82.5k and 21.5k resistors should be metal film for temperature stability. The ubiquitous 1N914 *idiot* diode is also included to save the transistor in case of cockpit error (reversing the power supply polarity).

The type of transistor used at Q1 depends on the intermediate frequency and the desired gain. The gain-bandwidth factor (f_T) of the transistor should be eight to ten times the intermediate frequency for maximum stable gain. Transistors having the highest dc current gain (h_{FE}) at the lowest collector current will also yield maximum gain. In the circuit shown, the overall conversion gains from 432 to 28 MHz for several types of transistors, used with the same mixer, are tabulated below.

Q1	gain (dB)
2N708	10.5
2N3646	12.5
2N3692	2.5
2N3693	8.0

The collector current may be monitored by inserting a low-range milliammeter between the 12-volt supply and the mixer-amplifier unit. Set the wiper arm of the 5000-ohm potentiometer to the ground end, which will cut off the transistor. The meter will thus read only the bleeder current drawn by the 21.5k resistor in series with the pot. Then adjust the pot so that the meter indicates the bleeder current plus 0.56 mA. (The base current will also be measured, but it is negligible compared with the bleeder and collector current.)

It should be noted that **fig. 4** is not an error; the rf input signal is applied to the mixer i-f port, and the i-f output is taken from the rf port. This was done to provide a direct ground return for the emitter of Q1.

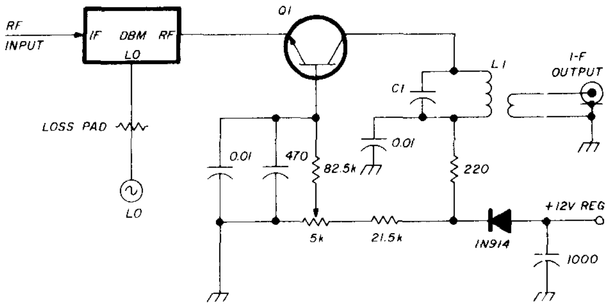

fig. 4. Schematic diagram of a double-balanced mixer with its LO and output ports terminated. L1 and C1 are resonant at the intermediate frequency; the text covers selection of the type of transistor. The mixer i-f and rf ports are reversed in this application.

As shown in **fig. 5**, the dc path from the rf port is directly through the transformer winding, but passes through the diodes from the i-f port. It can also be seen that an rf signal applied to either the i-f or the rf port will reach the same points in the diode quad with the same phase relationship. Therefore, except for the fact that the i-f port is usable down to dc, the two ports are interchangeable. As long as the desired intermediate frequency is above the minimum usable frequency of the rf port, typically 0.5 to 5 MHz for 500 MHz mixers, interchanging the ports has no effect.

VSWR measurements below 450 MHz

Measurement of VSWR using conventional slotted lines, such as those manufactured by Hewlett-Packard, GenRad, and General Microwave, is generally restricted to frequencies above 400 MHz because of the line length limitation. However, another technique is available which does not utilize anything more complicated than a signal generator, an SWR indicator (Hewlett-Packard model 415B or equivalent), and a resistive VSWR bridge. The use and applications of the SWR indicator have been published previously.[4] The resistive VSWR bridge, which has been used by many vhf experimenters for several years, was described by Joe Reisert in his article on antenna matching.[5] Although that article covered use of the bridge only to achieve minimum VSWR for antenna matching, the bridge can be used to measure the VSWR of any device from 3 to 450 MHz with a reasonable degree of accuracy, and does not require any sophisticated test equipment for calibration.

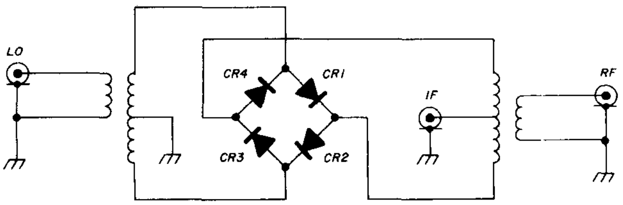

fig. 5. Schematic diagram of a typical commercial double-balanced mixer.

A few words are in order about the bridge, which is shown schematically in **fig. 6**. For best performance, resistors R1 and R2 must be matched to within one per cent, be virtually leadless, and should rest on a ground plane. The ground plane can be made of copper-clad board which has been notched to clear the connectors, and should be fastened in the enclosure just below the center pins of the four connectors.

Fig. 7 shows the test setup for both calibrating and using the bridge. The signal generator must be modulated at 1000 Hz, and must have sufficient output power to enable the SWR indicator to register 0 dB on its 30-dB range with J3 open-circuited or shorted. As large a pad as possible should be used between the signal generator and the bridge to keep the load presented to the generator as nearly constant as possible. If necessary, the sensitivity of the bridge may be increased by removing R3, which serves no useful purpose in this application nor in the one described in the original article.

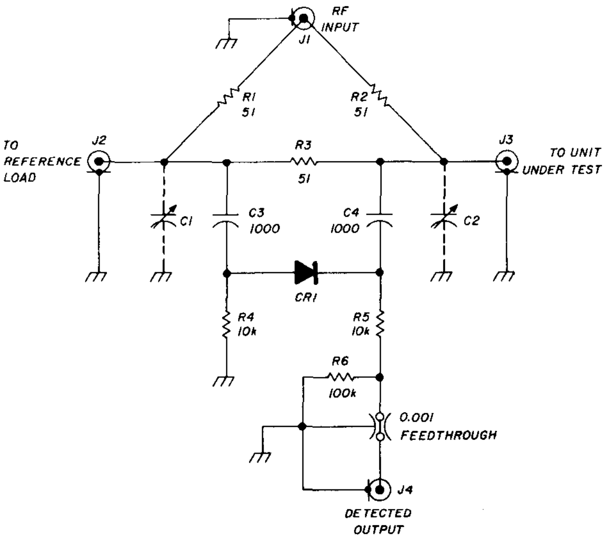

fig. 6. Schematic diagram of the VSWR bridge described in reference 5. Resistors R1 and R2 must be matched within one per cent. Either C1 or C2 (not both) is used to compensate for capacitive unbalance. CR1 is a 1N82 or equivalent germanium diode.

Before calibrating the bridge, the capacitive tab (C1 or C2) must be adjusted to yield a *return loss* of at least 40 dB at 450 MHz or at the highest frequency within your signal generator range, if less than 450 MHz. (I use the term "return loss" to indicate the relative balance of the bridge, as displayed on the dB scale of the SWR indicator meter, because it corres-

table 1. Calculated VSWR of open or shorted 50-ohm resistive attenuators.

attenuation (dB)	VSWR	attenuation (dB)	VSWR
1	8.7	11	1.17
2	4.4	12	1.13
3	3.0	13	1.11
4	2.3	14	1.08
5	1.92	15	1.07
6	1.67	16	1.05
7	1.50	17	1.04
8	1.38	18	1.03
9	1.29	19	1.025
10	1.22	20	1.02
		23	1.01
		30	1.002

ponds to return loss in a directional coupler, although the terminology is not absolutely correct in regards to an unbalanced bridge.) After the 0 dB reference level has been set on the 30 dB range, connect a 50-ohm load of known accuracy to J3. Then probe, with a small strip of metal touching the ground plane, the area adjacent to the pin of J2 or J3, to determine if C1 or C2 must be added. Solder a copper or brass tab at that point and bend the tab for minimum indication on the meter. If your load has a known VSWR of 1.1, the meter should read at least 25 dB below the reference level; if the VSWR is 1.05, the reading should be down at least 35 dB. A really good 50-ohm termination will result in a return loss of more than 40 dB.

To calibrate the SWR meter dB scale in terms of VSWR obtained with the bridge, a set of known mismatches is required. Since such mismatches are not readily available, unterminated or shorted coaxial resistive attenuators can be used instead. **Table 1** lists the calculated VSWR for open or shorted attenuators of standard values. A shorted attenuator is preferable, because the short eliminates fringing in an open connector, but either is sufficiently accurate in this application.

Because the bridge is an imperfect device, it is somewhat frequency sensitive. **Fig. 8** shows the theoretical curve of VSWR plotted against return loss in dB, as well as a set of curves taken on my bridge at various frequencies, using shorted attenuators as mismatches. Note that at low values of VSWR the errors are minimal, but increase as the VSWR increases. Fortunately, this is what we would have hoped for.

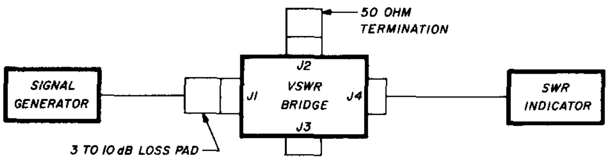

fig. 7. Test setup for calibrating and using the VSWR bridge. Known mismatches for calibration, or the device under test in actual VSWR measurements, are connected to J3.

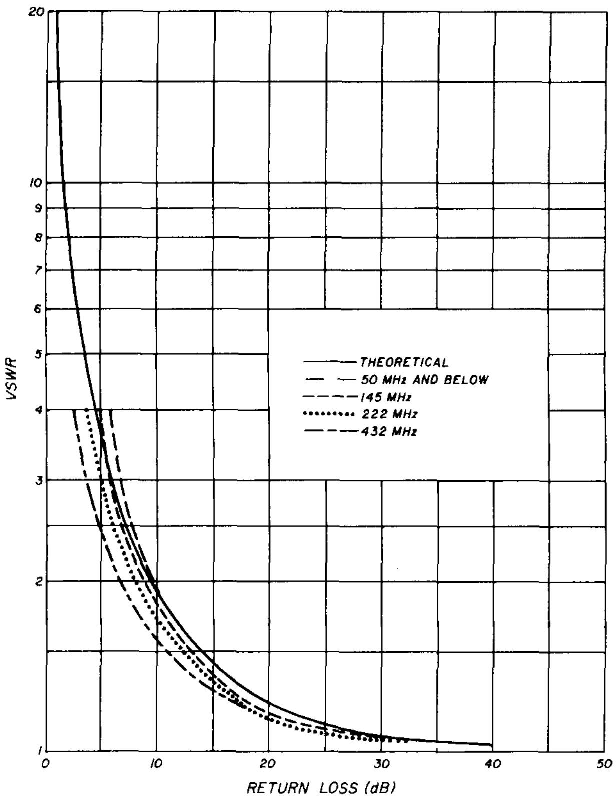

fig. 8. VSWR plotted against return loss for the VSWR bridge at several frequencies. Return loss is read on the SWR indicator shown in fig. 7.

To make your own calibration curves, establish the 0 dB reference level with J3 open, and determine the return loss for each known mismatch. It is also important to record the signal generator output and modulation percentage used for each calibration curve. When actually using the bridge, the same output and modulation level should be used. Otherwise an additional error may be introduced because the bridge diode is not in its square-law region at the high signal level required to set the 0 dB reference.

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1. James R. Fisk, W1HR, "Microstrip Transmission Line," *ham radio*, January, 1978, page 28.
2. Dan Cheadle, "Selecting Mixers for Best Intermod Performance," *Microwaves*, Part I, November, 1973, page 48; Part II, December, 1973, page 58.
3. "Reactive Loads — The Big Mixer Menace," Technical Note, Anzac Electronics Division of Adams-Russel, 38 Green Street, Waltham, Massachusetts 02154.
4. Robert S. Stein, W6NBI, "Using the SWR Indicator," *ham radio*, January, 1977, page 66.
5. Joseph H. Reisert, W1JAA, "Matching Techniques for VHF/UHF Antennas," *ham radio*, July, 1976, page 50.

ham radio

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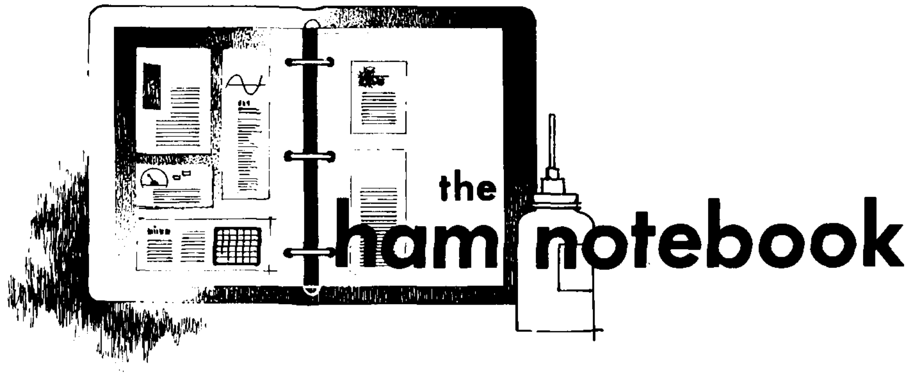


a method for measuring inductance or capacitance

Given a frequency counter and a calculator with buttons for square, reciprocal, and *pi*, there's an easy way to measure inductance in terms of a known capacitance, or a capacitance in terms of a known inductance. **Fig. 1** shows a complete setup for this

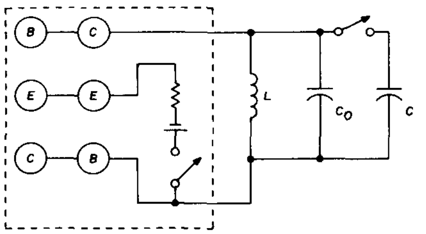


fig. 1. Arrangement for measuring inductance or capacitance in terms of known values. *E, B, C* refer to emitter, base and collector of a transistor pair. Plug equation into a calculator to determine the unknowns. Frequency is measured with a counter.

purpose. Inside the dotted enclosure is any sort of arrangement that will cause an inductance with parallel capacitance to oscillate. Other arrangements may well be better for the purpose, but I like the one shown because of its simplicity. Mounted in a small box, it is a handy gadget to have around as it will make almost anything oscillate. The letters **E, B, C**, refer to the emitter, base, and collector leads of a pair of transistors; those on the left are for one transistor, and those on the right for the other. The resistor value is far from critical. Anything over 1500 ohms will ensure that the battery drain will never exceed 1 mA.

Whatever the arrangement used inside the dotted enclosure, the procedure is simply to measure the frequency, f_0, with the switch open and f with the switch closed. Then use the calculator and the equation to determine L if C is known, or C if L is known. In the equation, frequency is in megahertz, inductance in millihenries, and capacitance in picofarads.

The essential feature of the method is that neither the value of C_0 nor the capacitance of the oscillation-generating device enters into the equation. So long as the latter does not change when the switch is operated, the results should be quite accurate.

Of course a little common sense should be used. Don't make C_0 too large compared with C, or the frequency change will be too small for accurate measurement. On the other hand, if C_0 is too small compared with C, the frequencies will be so different that the effect of the device may change as the switch is operated. With these extremes avoided, the measurement should be independent of the value chosen for C_0. Also look out for the effect of hand capacitance and oscillator drift. Open and close the switch several times to make sure the frequency readings repeat.

The method would hardly be practical without the counter and the calculator, which may be why I've never seen it described before.

Walter van B. Roberts, W2CHO/K4EA

audio-driven DSB generators

One feature of a properly adjusted class-C amplifier is the linearity of its output power with respect to its input power. It should be applicable to many requirements if all the input power to the final stage of a DSB transmitter is provided by a high-power audio system, as suggested in **fig. 2**. Simple isn't it? No crystal filters, no phasing networks, and practically no standby power consumption. Great for portable operation!

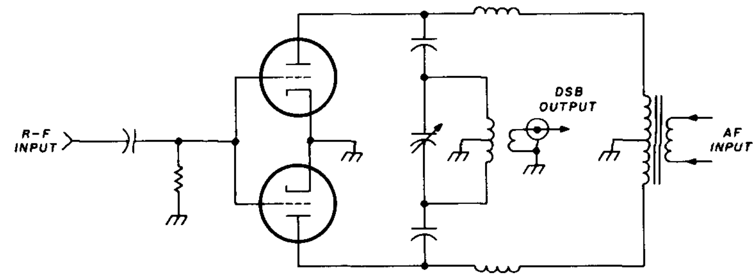


fig. 2. Suggested typical audio-driven DSB generator.

So, why should one even consider DSB when every ham knows that SSB is better and uses less space in the rf spectrum? But wait a minute! Everything has its application — SSB for the crowded hf bands, of course; but on the higher-frequency bands there's a lot more room. And DSB can be received with phase-lock detection, which will provide AFC action — ideal for the probable channelized operation. When (and if) CB surfaces on frequencies in the hundreds of MHz, it will almost certainly be of a channelized nature, and AFC will probably be a requirement.

There are at least three different methods of receiving DSB. One is to use sharp filtering, which eliminates one sideband, and then treats the remaining signal as though it were SSB. Another involves phase-reversal techniques. When polarity reversals of the modulating frequency produce phase reversals of the rf signal, a corresponding phase reversal in the detection process will compensate, resulting in normal signals.

The third, and somewhat simpler method, is to phase lock an oscillator to the sideband signal on twice the i-f of the missing carrier. The oscillator signal can be divided by two with a simple diode circuit, and there's your carrier for re-insertion. This last method also results in the AFC effect previously mentioned.

Perhaps with the approach of more and more channelized operation as the higher-frequency bands are exploited, it would behoove us to investigate this little-used mode of transmission.

If you must use SSB, however, a power-efficient transmitter can be made by combining two of these DSB generators with appropriate phasing. Methods of doing this have been amply covered in the literature during the last twenty-five years and are not repeated here.

There is, however, an interesting angle I've not seen and which could have applications as a high-power emergency CW transmitter. The joker is that two-phase power would be needed, but an arrangement similar to **fig. 3** should provide a respectable dc note when the input to the final amplifier is ac, direct from the mains! Saving the cost and weight of the rectifier and filter system just could be the deciding design consideration for some applications. It's worth thinking about.

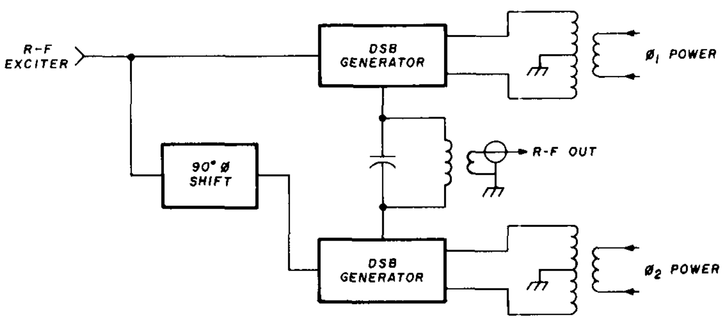


fig. 3. Possible transmitter without dc power.

Henry S. Keen, W5TRS

tuning aid for crystal-controlled vhf receivers

Aligning a crystal-controlled 2-meter or commercial high-band receiver using a VFO-based signal generator is tedious and exasperating because of generator drift and the difficulty of setting the exact frequency. For occasional use, the obvious alternative, a synthesized generator, is usually prohibitively expensive.

If one is available, a Regency *The Touch* synthesized scanner (Model ACT-T-16K) serves as a good substitute. The local oscillator in *The Touch* is 10.7 MHz below the programmed receive frequency on two meters and high band, and incidental radiation from the LO is audible at a distance of 30-40 feet (9-12 meters) on a tuned handheld with a rubber duck antenna. Thus, for a signal at 146.52 MHz, program in 146.52 + 10.7 or 157.22 MHz.

The Touch's priority feature provides an added bonus: If a nonpriority channel is used and the priority feature enabled in the manual mode, the periodic priority channel check gives a switched carrier with a distinct "chugging" sound that's easy to hear and tune to.

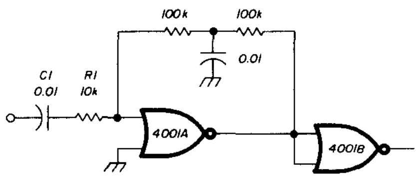
Without alteration of *The Touch* coupling is by radiated signal, and attenuation is provided by changing the distance. People in the area will affect received signal strength, so stand still while tuning.

The signal provided by *The Touch* this way will be delightfully stable and accurate enough for tuning. It will not usually provide a perfect frequency adjustment, but it will normally bring the receiver close enough to hear the repeater or base station for a final frequency tweak from the actual source.

David McLanahan, WA1FHB

no-adjust bias for VLF dip meter

When building the VLF dip meter converter described in the August, 1979, *ham radio*, it is possible to replace R2 with a fixed RC circuit. This modification, shown in **fig. 4**, pro-



vides no-adjust bias without sacrificing gain. In **fig. 4** the node between 4001A and 4001B automatically adjusts to the switch threshold of the first gate. Inherent matching in the chip means it also will be very close to the second gate's switch point.

Anthony L. Carson, WB3IDJ
Baltimore, Maryland 21234

More Details? CHECK—OFF Page 94

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11C70DC	600 MHz Flip/Flop with reset	12.30
11C58DC	ECL VCM	4.53
11C44DC/MC4044	Phase Frequency Detector	3.82
11C24DC/MC4024	Dual TTL VCM	3.82
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2N1692	15.00	2N5637	22.15	MM1553	56.50
2N1693	15.00	2N5641	6.00	MM1601	5.50
2N2632	45.00	2N5642	10.05	MM1602/2N5842	7.50
2N2857JAN	2.52	2N5643	15.82	MM1607	8.65
2N2876	12.35	2N6545	12.38	MM1661	15.00
2N2880	25.00	2N5764	27.00	MM1669	17.50
2N2927	7.00	2N5842	8.78	MM1943	3.00
2N2947	18.35	2N5849	21.29	MM2605	3.00
2N2948	15.50	2N5862	51.91	MM2608	5.00
2N2949	3.90	2N5913	3.25	MM8006	2.23
2N2950	5.00	2N5922	10.00	MMCM918	20.00
2N3287	4.30	2N5942	46.00	MMT72	1.17
2N3294	1.15	2N5944	8.92	MMT74	1.17
2N3301	1.04	2N5945	12.38	MMT2857	2.63
2N3302	1.05	2N5946	14.69	MRF304	43.45
2N3304	1.48	2N6080	7.74	MRF420	20.00
2N3307	12.60	2N6081	10.05	MRF450	11.85
2N3309	3.90	2N6082	11.30	MRF450A	11.85
2N3375	9.32	2N6083	13.23	MRF454	21.83
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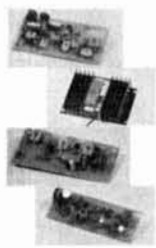

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More Details? CHECK—OFF Page 94

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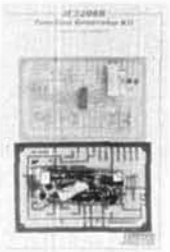

Provides three basic waveforms: sine, triangle and square wave. Frequency range from 1 Hz to 100K Hz. Output amplitude from 0 volts to over 6 volts (peak to peak). Uses a 12V supply or a ±6V split supply. Includes chip, P.C. Board, components & instructions.

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JE300 .... $39.95

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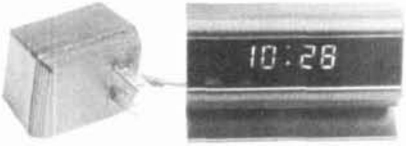
Use Intersil 7205 Chip. Plated thru double-sided P.C. Board. Red LED display. Times to 59 minutes, 59.59 seconds with auto reset. Quartz crystal controlled. Three stopwatches in one: single event, split (cummulative) and taylor (sequential timing). Uses 3 penlite batteries.

Size: 4.5" x 2.15" x .90"

JE900 .... $39.95

4-Digit Clock Kit



Bright .357" ht. red display. Sequential flashing colon. 12 or 24 hour operation. Black extruded aluminum case. Pressure switches for hours, minutes and hold functions. Includes all components, case and wall transformer. Size: 3¼" x 1¾" x 1¼"

JE730 .... $14.95

6-Digit Clock Kit

Bright .300 ht. common cathode display. Uses MM-5314 clock chip. Switches for hours, minutes and hold functions. Hours easily viewable to 20 ft. Simulated walnut case. 115VAC operation. 12 or 24 hour operation. Includes all components, case and wall transformer. Size: 6¾" x 3-1/8" x 1¾"

JE701 .... $19.95

Jumbo 6-Digit Clock Kit

Four .630" ht. and two .300" ht. comm. anode displays. Uses MM5314 clock chip. Switches for hrs., mins., and hold functions. Hours viewable to 30 ft. Sim. walnut case. 115VAC operation. 12 or 24 hour operation. Incl. all components, case and wall transformer. Size: 6¾" x 3-1/8" x 1¾"

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Uses LM309K. Heat sink provided. PC board construction. Provides solid 1 amp @ 5 volts. Includes components, hardware & instructions. Size: 3½" x 5" x 2"h

JE200 .... $14.95

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JE205 .... $12.95

Variable Power Supply Kit

Full 1.5 amp @ 5-10V output. Up to .5 amp @ 15V output. Heavy duty transformer. Three-terminal I.C. voltage regulator. Heat sink provided for cooling efficiency. PC board construction. 120VAC input. Size: 3½" x 5" x 2"h

JE210 .... $19.95

62-Key ASCII Encoded Keyboard Kit

The **JE610 ASCII KEYBOARD KIT** can be interfaced into most any computer system. The JE610 kit comes complete with an industrial grade keyboard switch assembly (62-keys), IC's, sockets, connector, electronic components and a double-sided printed wiring board. The keyboard assembly requires +5V @ 150mA and −12V @ 10mA for operation. Features: 60 keys generate the full 128 characters, upper and lower case ASCII set. Fully buffered. Two user-define keys provided for custom applications. Caps lock for upper case-only alpha characters. Utilizes a 2376 (40-pin) encoder read-only memory chip. Outputs directly compatible with TTL/DTL or MOS logic arrays. Easy interfacing with a 16-pin dip or 18-pin edge connector.

JE610 .... $79.95

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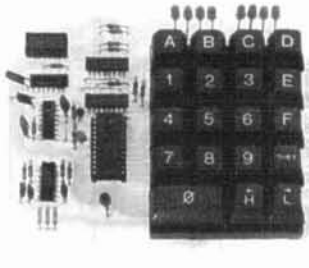

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More Details? CHECK—OFF Page 94

NEW products

new nine-band amateur transceiver has new WARC bands

The new Ten-Tec Delta transceiver is the answer to your wishes and desires — a transceiver capable of covering the new WARC bands at 10, 18, and 25 MHz, packaged in a handsome, small enclosure, and available at an affordable price. Delta is equally at home in the car or at home on SSB or CW, RTTY, or SSTV. Delta has all of the features you have come to expect from a Ten-Tec transceiver, and more.

It covers 160 through 10 meters in nine bands; 10.0 to 10.5 MHz band is fully operational on transmit and receive as received. The 18.0 to 18.5 MHz and 24.5 to 25.0 MHz bands are fully incorporated except for accessory plug-in crystals, available when these bands open to Amateurs. Delta has QSK — instant break-in. It is all solid state, and uses basic 13 Vdc circuits. Another Delta feature is wideband, no-tune final amplifier and receiver front end. Delta has new styling and small size; 4-3/4 × 11-3/8 × 15 inches. The new Model 280 Power Supply comes with over-voltage and over-current protection.

Other features are built-in VOX and PTT, built-in notch filter, hang agc for smooth operation, three selectivity responses to choose from, with optional 500 Hz, six-pole ladder i-f filter for CW. Delta also has a built-in, 20-dB receiver attenuator, excellent receiver dynamic range, and a full line of accessories is available.

For more information and complete specifications, see your nearest Ten-Tec dealer, or write Ten-Tec, Inc., Sevierville, Tennessee 37862.

new high-frequency relay from Dow-Key

Dow-Key Division of Kilovac Corporation announces the availability of the Model 401-230832, single-pole, double-throw coaxial relay. The new relay is designed for excellent rf performance in microwave systems to 18 GHz and for operation in severe environments. The 401-230832 is a failsafe relay, remotely actuated by a 28-Vdc, 475-ohm coil with a maximum operate time of 20 ms at 20°C. A balanced actuator mechanism provides very good tolerance to shock and vibration. Auxiliary contacts are provided for remote position-indicator circuits.

The 401-230832 can carry 10 watts of rf power at 18 GHz with isolation greater than 60 dB, insertion loss less than −0.5 dB, and VSWR 1.5:1, maximum. At 1 GHz, VSWR is less than 1.05:1, isolation is greater than 80 dB, insertion loss is less than 0.1 dB, and the maximum operating power is 75 W. At lower frequencies, the 401-230832 is an excellent choice for high isolation with low VSWR and insertion loss. The maximum power capacity of the relay is 200 watts up to 200 MHz. Typical applications include switching of antennas, components, receivers, and instruments in radar and communications.

Small quantities are available from stock. The price is $140 each for quantities of one to four pieces. For complete information contact Jack Dysart, Product Line Supervisor, P.O. Box 4422, Santa Barbara, California 93103.

Heath's new HM-2141 dual meter vhf wattmeter

Heath Company, the world's leading manufacturer of electronic kits, has announced the introduction of a new dual-meter vhf wattmeter for the Radio Amateur. The new HM-2141 monitors both forward and reflected power simultaneously, between 50 and 175 MHz.

According to Heath, the HM-2141 measures forward and reflected average power, forward and reflected peak envelope power (PEP), and standing wave ratio (SWR). The dual-range meter gives simultaneous readings of transmitted output up to 30/300 watts forward, and 10/100 watts reflected power for complete ease of antenna tuning.

Heath specifications give this Dual Meter kit an average forward accuracy of ±7.5 per cent of full scale. It reads SWR directly from 1:1 to 3:1. The factory-assembled and calibrated sensor can be mounted inside the 4-1/8 × 7-1/2 × 6-3/8 inch cabinet or separately. The HM-2141 can be powered by a 9-volt battery or on 120 Vac using the optional PS-2350 converter. The 9-volt battery is required for PEP operation only.

Mail order priced at $74.95, F.O.B. Benton Harbor, Michigan, the HM-2141 is featured, along with other Amateur gear and nearly 400 kits you can build yourself, in the latest Heathkit catalog. For a free copy write Heath Company, Department 350-130, Benton Harbor, Michigan 49022. Free catalogs are also available at Heathkit Electronic Centers

(Units of Veritechnology Electronics Corporation), listed in your telephone white pages.

Heath is a subsidiary of Zenith Radio Corporation.

Micronta power inverters

Just introduced by Radio Shack are two new Micronta power inverters for converting 12 Vdc and 120 Vac to power ac appliances from your car, boat, or recreational-vehicle battery.

The 300-watt inverter is said to be capable of powering color TVs, electric typewriters, small hand drills, sewing machines, and many other items requiring no more than 300 watts continuous power. The 100-watt inverter is suitable for powering small TV sets, electric razors, transistor radios, Amateur and CB two-way radio equipment, and other small appliances.

Both inverters feature a NORMAL/BOOST switch to provide extra power to compensate for low battery-input voltage. Automatic overload protection causes the inverters to turn themselves off if overloaded. Circuit breaker automatically resets 3-4 seconds after the overload has been removed.

Full-load input current is given as 25 amps for the 300-watt model; 12 amps for the 100-watt inverter.

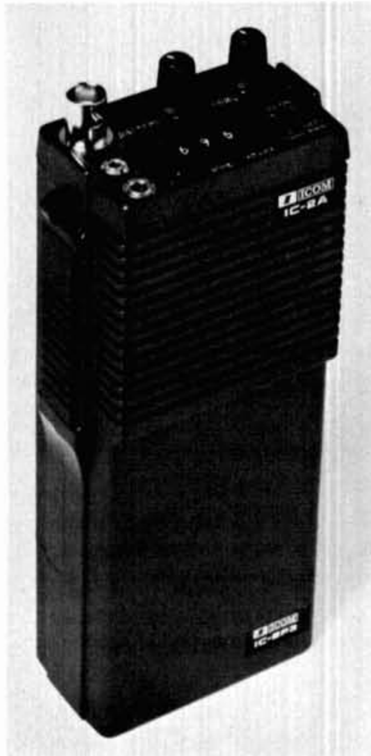
The Micronta 300-watt power inverter is priced at $79.95; 100-watt inverter, complete with cigarette lighter plug, is priced at $39.95.

Micronta power inverters are available exclusively from participating

More Details? CHECK—OFF Page 94

Radio Shack stores and dealers, nationwide.

IC-2A hand-held 2-meter rig



Icom is proud and excited to announce the availability of their compact 800-channel IC-21 hand-held rig.

The IC-2A will be marketed in three basic packages:

IC-2A with alkaline battery pack;

IC-2A with NiCd pack and wall charger;

IC-2A with NiCd pack, wall charger, and built-in tone pad.

The NiCd pack (IC-BP3), and Charger (BC-25UO), is available separately. The Alkaline Pack (IC-BP4) without batteries is also available. The tone pad is a user-installable, plug-in option. Price and availability of additional options, including speaker-mic, drop-in desk charger, and leather case will be announced shortly. All IC-2As are supplied with a flexible rubber antenna and a belt clip.

More Details? CHECK—OFF Page 94

We are sure that the IC-2A, with its extremely compact size, unique slip-on battery pack, and its easy-to-understand operational features, at a price significantly below that of any others in the industry, will be a most popular item.

In addition, the IC-502A is now available. Similar to the IC-502, it incorporates a fine-tuning control to provide excellent band spread in an economical 6-meter portable rig. See your nearby Icom dealer, or write Icom America, Inc., 2112 116th Ave. N.E., Bellevue, Washington 98004.

free Heathkit winter catalog

The new 104-page Heathkit winter catalog describing the latest in electronic kits is now available from Heath Company, Dept. 350-200, Benton Harbor, Michigan 49022.

The free catalog lists nearly 400 kits for home, work, and pleasure, including the latest in home computers, color TVs, Amateur Radio, audio components, precision test instruments, educational self-instruction programs, and innovative electronic devices for the home.

In this catalog, Heath Company is introducing foreign-language self-instruction programs for the first time. Heath claims that now anyone can learn Spanish, French, German, or Italian in the comfort of his own home with these new programs.

The foreign-language programs are aided by a special electronic translator that displays the spelling and at the same time pronounces the foreign-language equivalent of the word entered in English.

The catalog also introduces new self-instruction programs in statistics, and, for the Amateur Radio buffs, a new high-gain, tri-band antenna.

The complete catalog is available free by writing Heath Company, Benton Harbor, Michigan 49022. Heath Company, the world's largest manufacturer of electronic kits, is a subsidiary of Zenith Radio Corporation.

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Coming Events

PENNSYLVANIA: The Broadcasters' Amateur Radio Club will conduct its Hamfest on July 13th from 9 a.m. to 4 p.m. at the Pocono Downs Race Track, Rt. 315, Wilkes-Barre. Unlimited outdoor and indoor space, refreshments, prizes; admission $2.50 — XYL's and children free. No additional charge for sellers. Gates open at 8 a.m. for set-up. Talk in 147.66/.06 and 146.52. Contact: Charles Baltimore, WA3NUT (717) 823-3101; B.A.R.C. 62 S. Franklin St., Wilkes-Barre, PA. 18773.

VERMONT: The Burlington A.R.C. is holding their International Hamfest on August 9th and 10th at the Old Lantern Campgrounds, 14 miles south of Burlington. Many events are planned — flea market, commercial exhibitors, traditional "Can-Am" tug-of-war, prizes. Admission $4 — talk-in on .34/.94, W1KOO/RPT. For more details write: Hap Preston, W1VSA, Box 312, Burlington, VT 05402.

NEW YORK: HAM-O-RAMA '80 on September 12th and 13th at the Erie County Fairgrounds in Hamburg. Exhibits, tech programs, prizes, flea market. Plenty of free parking, free RV hookups. Advance tickets $3 — contact Ron Brodowski, KC2P. 260 Hilltop Drive, Elma, NY 14059. (716) 652-6754.

SOUTH DAKOTA: The Black Hills A.R.C. is holding their Hamfest and Picnic, Friday, July 25th at 4 p.m. through Sunday, July 27th, at the South Dakota School of Mines Campus, Rapid City. Registration is $6.50 before July 1st; $7 at door. Forums, tours, exhibits, flea market, contests, prizes, YL activities — flea market tables are free. Sunday noon meal will be catered — tickets available at door. Call in frequencies 146.34 - .94 — contact WØBLK. For pre-registration and further details write: Black Hills A.R.C., P.O. Box 1014, Rapid City, SD 57709.

WASHINGTON: Seattle National Amateur Radio Convention (SEANARC), July 25-27, SEA-TAC Airport Red Lion Motor Inn, Seattle. Seminars, displays, forums, major equipment exhibitors. Write: SEANARC, P.O. Box 68534, Seattle, WA 98168.

MONTANA: Gallatin Ham Radio Club's Ghost Town DXpedition from 1800 UTC Saturday, July 5th, until 1800 UTC July 6th from Bannack. Callsign will be W7ED, on the following frequencies: 7235 kHz SSB, 14060 kHz SSB and 21360 kHz SSB plus or minus 5 kHz. Special certificates for those sending QSL's, SASE and $1 to: Bannack DXpedition, 417 Staudaher St., Bozeman, MT 59715.

PENNSYLVANIA: The Beaver Valley Hamfest will be held on Sunday, July 20 from 9 a.m. to 5 p.m. at the Community College of Beaver County in Monaca. Tickets $2 each or 3 for $5. Free indoor vendor space, free outside flea market space, free parking. Talk-in on 146.25/85, 223.26/86, or 146.52 simplex. Contact: Adam Horniak, WB3JZN, 182 Edgewood St., Aliquippa, PA 15001 (412) 378-9667; or Gary Mohrbacher, WB3FKE, 3417 47th St., New Brighton, PA 15066 (412) 843-9546.

FLORIDA: The Jacksonville Hamfest and ARRL Florida State Convention sponsored by the Jacksonville Hamfest Assoc., will be held August 2nd and 3rd at a new location — the Orange Park Kennel Club, Jacksonville. Programs, forums, manufacturer and dealer exhibits; inside swap tables $5 a day per table — order from: Andy Burton, WA4TUB, 5101 Younis Rd., Jacksonville, FL 32218. Special DXers forum and dinner banquet — tickets $11.50 — write: N4KE, 258 Wesley Rd., Green Cove Spring, FL 32043. For more information write: Jacksonville Hamfest Association, 911 Rio St., St. Johns Dr., Jacksonville, FL 32211.

RADIO EXPO "80" Lake County fair grounds, Rt. 45 & 120. Sept. 6 & 7 — advanced tickets $2.00, $3.00 at gate. Write: Radio Expo Tickets, P.O. Box 1532, Evanston, IL 60204. Exhibitor information call (312) BST-EXPO.

MINNESOTA: The Iron Range Hamfest will be held on July 13th from 9 a.m. to 5 p.m. at the St. Louis County Fair Grounds, in Hibbing. Camping facilities are available at $3.50 per night/day including elec. & water; other accommodations in town. Free tables for flea market — lunch available. Talk-in on 19/79.

ILLINOIS: Fox River Radio League Hamfest, Sunday, August 24th, Kane County Fairgrounds, St. Charles. Free outside flea market - inside display area. Table discounts available. Contact Gary Senesac, KA9ADP, 926 Britta Lane, Batavia, IL 60510. Tickets: $1.50 advance; $2 at gate. Contact Jerry Frieders, W9ZGP, 1501 Molitor Rd., Aurora, IL 60505. Talk-in on 146.94.

COLORADO: The RMRL's annual Field Day Demonstration and Swapfest, Sunday, July 20th at Karl Ramstetters (WAØGUN) Ranch, Golden. Activities start at 10 a.m. — bring own food, chairs, blankets — soft drinks provided. Talk-in on 34/94 — bring your family!

WASHINGTON: The Radio Club of Tacoma (W7DK) HAMFAIR will be held August 23-24, on the campus of Pacific Lutheran University, Tacoma. Door prizes, flea market, banquet, loggers breakfast, commercial exhibits and much more. Talk-in 88/28. Info from Joe Winter, WA7RWK, 819 No. Mullen, Tacoma, WA 98406. Tel: (206) 759-9857.

BRITISH COLUMBIA: The Maple Ridge Amateur Radio Club is hosting their Hamfest '80 on July 5th and 6th at the Maple Ridge Fairgrounds, east of Vancouver. Registration $4 - $11 including banquet. Food, prizes, swap & shop, ladies program, and camper space available - no hook-ups. Talk-in: 3.755, 146.34/94 and 146.19/79. For more info write: Bob Haughton, VE7BZH #20625-114th Ave., Maple Ridge, B.C. V2X 1S7 Canada.

PENNSYLVANIA: Harrisburg RAC Annual Firecracker Hamfest on Friday, July 4th at the Shellsville, VFW Picnic Grounds, Harrisburg. Shade trees and pavilion; parking for 1,000 cars. Food available or picnic. Admission $3 — XYL and children free. Tailgating $1.50. Prizes awarded. Write: Richard Kerlin, K3AM, 635 Lenker Rd., Harrisburg, PA 17111.

NEW JERSEY: The West Jersey Radio Amateurs Hamfest will be held on July 20th at McGuire AFB, Wrightstown from 9 a.m. to 4 p.m. Admission $2.50 — spouses and children free. Tailgate or table space, $2.50 — bring own tables. Refreshments and activities available. Talk-in on 52 and 146/925. For advance tickets and details send SASE to: Mary Lou Shontz, WB1QIU, 107 Spruce Lane, Rte. 16, Mt. Holly, NJ 08060, or call Mark Millman, N2ME at (609) 871-6691.

MINNESOTA: The Detroit Lakes Amateur Radio Club will hold its Picnic and Swapfest on Sunday, July 20th from 10 a.m. to 4 p.m. at Long Lake Park, 1½ miles west of Detroit Lakes. Tickets for drawing $1; picnic and swap tables available. Talk-in on 146.22/82 and 146.52/52. Contact Russ Berger NØARZ, 1406 Long Ave., Detroit Lakes, MN 56501.

OKLAHOMA: The Oklahoma State ARRL Convention and "Ham Holiday" will be held on July 25th through 27th at the Lincoln Plaza, Oklahoma City, sponsored by the Central Oklahoma Radio Amateurs. Talks, flea market, awards, full program for ladies available. $5 preregistration or $6 (after July 19th). Unlimited parking, full recreational facilities. Mail registration to: CORA, P.O. Box 15013, Oklahoma City, OK 73155.

TENNESSEE: The all-indoor Nashville Hamfest will be held on Sunday, July 27th beginning at 8 a.m. CDT at the National Guard Armory, Nashville. Admission $1; tables $3. Refreshments available. Talk-in on: .90/.30. For further details contact: Radio Amateur Transmitting Society, P.O. Box 2892, Nashville, TN 37219.

MICHIGAN: The Black River Amateur Radio Club will be operating a Special Event station during the National Blueberry Festival in South Haven, on July 16-20. The call will be WD8AGC and the frequencies: 3.975, 7.275, 14.275, 21.375, and 28.375 MHz. CW operations will be conducted randomly throughout Novice/Technician subbands. Any station working WD8AGC during this time can receive special certificate by mailing QSL card to: The National Blueberry Festival, P.O. Box 224, South Haven, MI 49090.

PENNSYLVANIA: The Delaware Lehigh A.R.C. (W3OK) and the Lehigh Valley A.R.C. (W3OI) presents their annual "Hamfest", "Computerfest", and "Electronics Fair" on July 20th from 8 a.m. to 4 p.m. at Franko's Farm, Bethlehem. Admission $3; Tailgaters $4; Indoor Sales $5; children free. Food and indoor facilities available. Talk-in on 52, 34-94, 10-70. For more details write: Wayne Comstock, WB3CDL, RD #1, Box 162B, Saylorsburg, PA 18353; or call: (215) 381-3674.

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INDIANA: The Indianapolis Amateur Radio Convention and Hamfest will be held on Sunday, July 13th at the Marion County Fairgrounds, Indianapolis. Write: Indianapolis Amateur Radio Assoc., Box 11086, Indianapolis, IN 46201.

NORTH CAROLINA: The Cary Amateur Radio Club's Mid-Summer Swapfest is on Saturday, July 19th at the Cary Lions Club Shelter (next to highschool) from 9 a.m. to 3 p.m. $3 registration; prizes to be awarded. Talk-in: 146.28/.88 or 146.52/.52. Tables rented or bring your own. Write: CARC, Box 53, Cary, NC 27511.

MISSOURI: The Zero-Beaters ARC will sponsor the Washington Hamfest at the Washington Fairgrounds on Sunday, July 20th. Prizes, activities for all; exhibits and traders row. Food available. Talk-in on .52 simplex. For more information write: ZBARC, Box 24, Dutzow, MO 63342.

SOUTH CAROLINA: The Charleston Amateur Radio Society is sponsoring the Charleston Hamfest on July 12th and 13th at the Omar Shrine Temple, Charleston. General admission $3.50, flea market tables, $5; Commercial booths $35; hospitality room, ladies activities. Food available — air conditioned. Talk-in on: 146.34/94, 146.16/76, and 146.19/79 for general use. Grand prize: Kenwood TS 120S. For more information contact: Charleston Hamfest Committee, P.O. Box 30643, Charleston, SC 29407; (803) 747-2324/563-2523.

MICHIGAN: The Shiawassee Amateur Radio Association is holding the SARA Hamfest and Michigan Nets Picnic on July 20th, at McCurdy Park, Corunna. For more information contact: Shiawassee Amateur Radio Assoc., 1302 W. Main St., Owosso, MI 48867.

MARYLAND: The Maryland Hamfest sponsored by the Baltimore Radio Amateur Television Society (BRATS) will be held on Sunday, July 27th at the Howard County Fairgrounds, West Friendship. Activities begin at 8 a.m. EST. Talk-in on 63/03, 16/76, and 52 simplex. For information write: BRATS, Box 5915, Baltimore, MD 21208.

ILLINOIS: The Belvidere Big Thunder Amateur Radio Club will hold its Hamfest at the Boone County Fairgrounds on Sunday, July 20th in Belvidere. Large indoor facility and plenty of outdoor space available. Camping available at 6 p.m. Saturday. Talk-in 146.52 and 147.375 Repeater. For advance tickets ($1.50), write: Mike George, 6159 Broadview, Belvidere, Illinois 61008.

PENNSYLVANIA: The South Hills Brass Pounders and Modulators Hamfest, Sunday, August 3rd, on the South campus of Allegheny Community College, south of Pittsburgh. Large indoor, air conditioned facilities, plenty of outdoor flea market area. Dealers, flea market, forums, food, prizes — doors open 11 a.m. Talk-in on 146.13 and 52 simplex. More information contact: Doug Wilson, WA3ZNP, 185 Orchard Ave., Emsworth, PA 15202.

ARKANSAS: The Arkansas Army MARS Convention will be held in Blytheville on July 19th and 20th at the National Guard Armory, highway 61 South. Registration is $7.50 and includes a catfish supper and pancake breakfast. Talk-in on 148.01 and 07/67. For more information contact: Richard Duncan, WB5CNV/AAR6SR, 209 Wilson St., Dell, Arkansas 72426.

PENNSYLVANIA: The Two Rivers Amateur Radio Club's Hamfest, Sunday, July 20th at the Penn State University, McKeesport Campus, McKeesport. Outside Flea Market vendors $5.00/car space. Door prizes, food and drink available — free admission. Talk-in on 146.22/82 MHz. For more info write: Gregory Lesko, Two Rivers A.R.C., McKeesport, PA.

WISCONSIN: The South Milwaukee Amateur Radio Club's "Swapfest '80" will be held on Saturday, July 12th from 7 a.m. to 5 p.m. at the American Legion Post #434, 9327 Shepard Ave., Oak Creek. Parking, picnic area, food & beverages, overnite camping available. Admission $2 — prizes to be awarded. Talk-in on 146-94 (2m) FM. For info write: S.M.A.R.C., P.O. Box 102, South Milwaukee, WI 53172.

ONTARIO: The Burlington Amateur Radio Club is sponsoring the Ontario Hamfest 1980 on Saturday, July 5th at the Milton Fairgrounds, Burlington. General Admission $3 — Children and YLs free. Gates open Friday, July 4th at 12:00 noon and on Saturday at 7:00 a.m. — tables are free. Camping and food available. Talk-in on: VE3/RSB 147-81, 147-21. Write: B.A.R.C., Box 836, Burlington, Ontario, Canada L7R 3Y7.

MAINE QSO PARTY: Sponsored by the Portland Amateur Wireless Assoc., the contest period is from 1600Z, July 19th to 2000Z July 20th. CW and Phone count as same contest — stations may be contacted once on each band and mode. Logs should show date/time in UTC, band and emission. Logs and summary sheet due by Sept. 1st. Suggested frequencies: CW: 1805, 3560, 7060, 14060, 21060, 28060; SSB: 1815, 3930, 7280, 14280, 21380, 28580; Novice: 3725, 7125, 21125, 28125. Write: Joe Blinick, K1JB, Portland Amateur Wireless Association, P.O. Box 1605, Portland, ME 04104.

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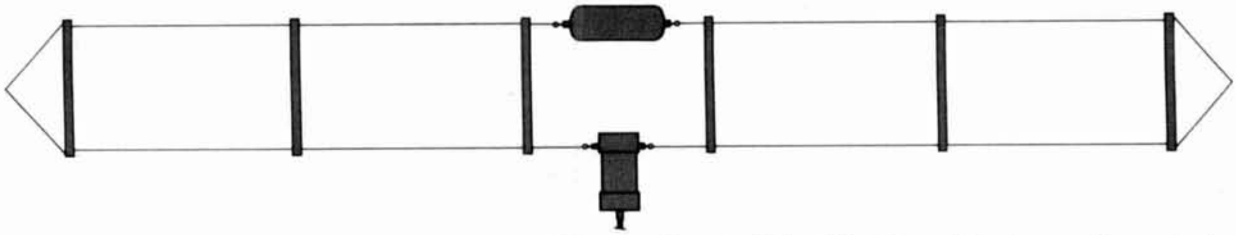

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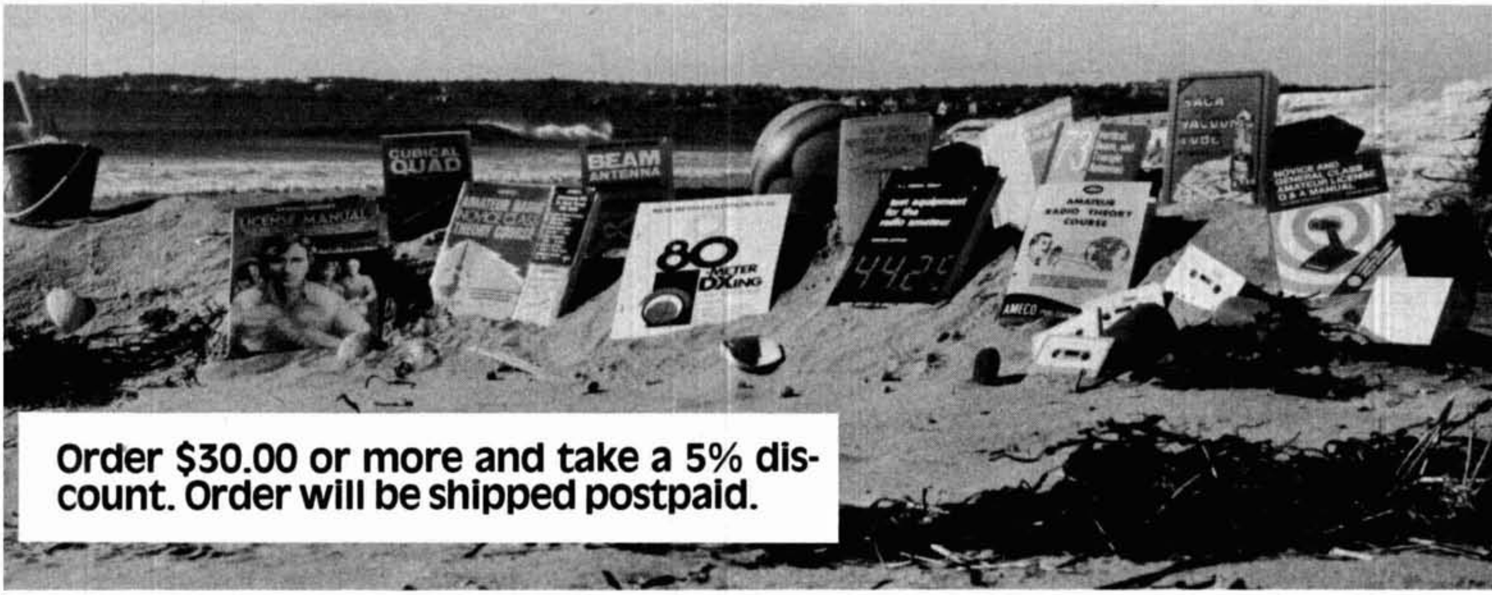

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by the ARRL staff

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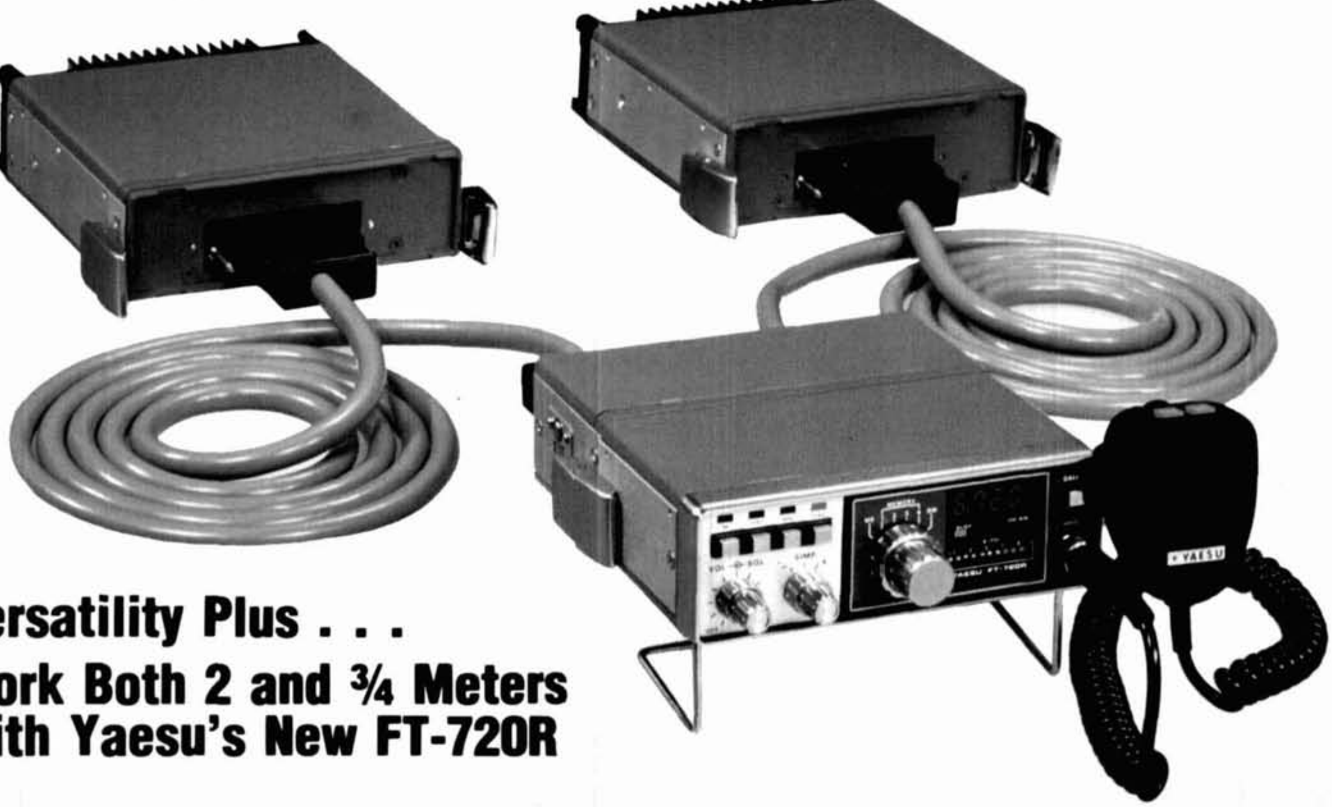

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