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ham radio

magazine

DECEMBER 1979

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More Details? CHECK—OFF Page 126

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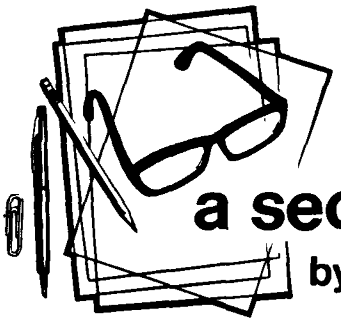

a second look

by Jim Fisk

What appear to me to be arbitrary and capricious decisions by FCC staffers in Washington have shown once again that the FCC bureaucrats are apparently interested in responding neither to public interest or public need. What I'm referring to, of course, was a recent announcement that FCC type acceptance of solid-state wideband amplifiers for the Amateur Service has been terminated *without explanation* by the FCC's Office of Chief Scientist. Are Radio Amateurs to be denied the use of modern solid-state technology because of the autocratic decision of an obscure bureaucrat in a *democratic* government? Is this not a contradiction to paragraph 97.1(c) of the FCC's own Regulations which state that one of the fundamental purposes of the Amateur Radio Service is to encourage and improve ". . . the amateur radio service through *rules* which provide for advancing skills in the . . . technical phases of the art?"

This recent action is just another in a long series of official FCC decisions which are contrary to the needs and desires of the Amateur Radio community — the linear amplifier ban, an unpopular and ridiculous callsign system, equipment type acceptance, the ASCII ban, recommending to the World Administrative Radio Conference (WARC) that CW be an "option" for the Amateur Radio service. This last item is a real dilly and stresses the need for closer congressional scrutiny of the Commission.

Several years ago, as part of the WARC preparations, the FCC formed the Advisory Committee for Amateur Radio (ACAR) and gave them the task of recommending, on behalf of Amateur Radio, what proposals should be made by the United States at WARC '79. ACAR carefully reviewed Article 41, which contains miscellaneous rules pertaining to the Amateur service including a Morse code proficiency requirement for operation below 144 MHz, and proposed no changes. As the WARC preparations proceeded, the FCC released Notices of Inquiry in Docket 20271 which requested public comment on various WARC draft proposals. The Commission requested comment on a proposal of "no change" to Article 41; those who responded supported that proposal. The FCC staff, however, chose to ignore both ACAR's advice and the public comments and recommended to the State Department that the United States' WARC position should include a proposal to delete the requirement for Morse code proficiency!

In Geneva the United States delegation proposed to make the code requirement below 144 MHz a "recommendation" rather than a requirement, a position that was supported by both Canada and Japan. Fortunately, some 15 administrations opposed the move as did every Radio Amateur in attendance. Brazil argued that any change to Article 41 would jeopardize existing reciprocal licensing agreements; Sweden proposed a 28-MHz cutoff; Papua New Guinea suggested 30 MHz. In the end the Papua New Guinea proposal won out and will be recommended for adoption by the conference. In effect, this lowers the frequency for a code-free Amateur Radio license to 30 MHz, a change which affects only six meters. Thus Amateur Radio lost a little — it could have been much worse — and the blame falls directly at the feet of unknown staffers within the FCC.

How could this happen? The *publicly stated* position of the FCC in April, 1977, regarding Article 41 reflected both the advice of ACAR and majority public opinion; its recommendations to the U.S. State Department less than two years later proposed a deletion of the Morse code requirement, in direct contradiction to the public's wishes. Apparently the change was conceived by some staff member (or members) within the Commission in direct violation of the Administrative Procedures Act, and no one in authority felt strongly enough about their public responsibility to veto it. It's no secret that the CB industry has been applying tremendous pressure for a code-free high-frequency operator's license, and this recent effort to sneak an unpopular proposal to an international forum leads one to believe they may have found a responsive element; except for Amateur Radio's friends in the international community, they would have succeeded.

Jim Fisk, W1HR
editor-in-chief

ICOM and 6 meters: Nobody Does It Better.

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ICOM's new **IC-551** is the all mode 6 meter unit in a compact, easy to use instrument, which uses a built-in microprocessor for frequency control and scanning. The no backlash, no delay dual VFO light chopper system, similar to the **IC-701** and the **IC-211**, is included as a standard feature at no extra cost, and provides split frequency operation as well as completely variable offsets.

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All stated specifications are subject to change without notice. All ICOM radios significantly exceed FCC regulations limiting spurious emissions.

comments

linear power amplifiers

Dear HR:

I would like to comment on Bill Orr's article on linear amplifier construction in the June issue. I believe his warning against military surplus tubes is much too conservative. I'm afraid most Radio Amateurs watch costs. New high-power transmitting tubes will go a long way toward driving the cost of a homebrew linear past the cost of a commercial amplifier. I have built a linear using three 813s in parallel, and at 1500 watts PEP it performs beautifully, just as predicted in author Orr's own *Radio Handbook*. Few tubes can beat the 813 for clean response or ruggedness (carbon plate version). I have acquired a collection of six JAN 813s, mostly from other hams; all have been tested in my amplifier and all work beautifully. For a small fraction of the cost of a 3-1000Z or even two 3-500Zs, I have a good set of tubes in my amplifier and a spare set on hand.

If you're considering building your own linear amplifier, read everything you can, such as W6SAI's fine articles, then build it yourself, and *scrounge*.

William Brain, KB5EY
Houston, Texas 77040

Right. There's nothing wrong with buying a surplus 813 from a fellow ham for a few bucks and trying it out in your rig; maybe he'll even take it back if it doesn't work.

It's all a matter of judgment. How about buying a surplus 8877? You can save nearly a hundred dollars over the user's price if you buy a surplus tube. But what if your "bargain" tube is bad? If you have bought it from a surplus dealer by mail, do you think you will get a refund? Fat chance. It all depends upon how much of a risk you want to take. You don't lose much with a surplus 813 or two.

Being in the power tube business, I am familiar with tear-stained letters from hams who have bought a JAN-branded, expensive, power tube and have been dismayed to find the tube bad and no warranty on it. But if you understand the limitations on warranty with respect to surplus tubes, and have the ability to test your tube immediately upon getting it home (and stand a reasonable chance of getting your money back if the tube is no good), why not? After all, plenty of people lose a wad of money every day at the horse races. But they have the fun of watching the horses run.

Bill Orr, W6SAI
Menlo Park, California

memory keyer

Dear HR:

I have just finished building the memory keyer featured in the April issue of *ham radio*; I wish to thank Robert C. Cheek and your magazine for a beautiful and accurate article on the construction of this keyer.

I substituted 21LO2 memory chips for the 2102s and then added a 7400 gate with all inputs tied to ground. A single switch in the 5-volt supply to the 7400 will give a high or inactive output on the four gates; three of these outputs are tied to the chip-enable pins of the 21LO2 memory chips. With the addition of separate switches in 5-volt supply lines to the memory board and the keyer, and a separate supply line with a 70-ohm resistor bypassing the switch to memory board, you can hold the memory in a power-down mode with a drain of less than 40 mA (compared with over 300 mA to the memory board alone at 5 volts).

I found the power supply ran a little warm when the drain was high and the keyer was left on overnight to retain the memory. The 70-ohm resistor reduces the supply voltage to 1.5 volts to the memory board; the chip-enable pins on the 21LO2s are floating at about 4.5 volts from gates of the 7400 chip. With the 5-volt supply to the 7400 switched off, the gate outputs are inactive and the 21LO2 functions as before the modification. Switching to memory power-down mode must take place before the 5-volt supply to the memory board is opened or else memory is lost. There is a zero time factor from power on to power down.

William Hansen
Glenwood, Illinois

split-band speech processor

Dear HR:

Congratulations to Wes Stewart, N7WS, for his fine article, "Split-Band Speech Processor," in September, 1979. Wes mentions that the circuit is sensitive to rf and that the proper use of ferrite beads, bypass capacitors, and rf shielding is important; he is correct. To this end, I would like to suggest that rf bypass capacitors be added in parallel with the 1N914 clipper diodes to prevent rf mixing. Values of several hundred picofarads should be sufficient.

Since symmetry is extremely important in the prevention of second order harmonics, I would also recommend replacing the 1N914 clipper diodes with diode pairs such as Motorola's MSD6150. Using two diodes manufactured on the same substrate provide close matching of V_f and other electrical characteristics. It is also an ideal means of keeping both diodes at the same temperature, ensuring the best clipping symmetry possible. The additions have proven themselves in the processor I am using.

James D. Allen, WA2SSO
Rochester, New York

(Continued on page 12)

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TS-180S *with* DFC

The TS-180S with DFC (Digital Frequency Control) is Kenwood's top-of-the-line all solid-state HF SSB/CW/FSK transceiver covering 160 through 10 meters, with outstanding performance and many advanced functions, including four tunable memories to provide more operating flexibility than any other rig!

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- Tunable noise blanker, to eliminate cross modulation from strong signals when noise blanker is on.
- Selectable wide and narrow CW bandwidth on receive (500-Hz CW filter is optional).
- SSB normal/reverse switch (proper sideband is automatically selected with band switch).
- Dual RIT (VFO and memory/fix).
- Available without DFC. Digital frequency display still included, with differential function showing difference between VFO and "digital hold" frequencies.

OPTIONAL ACCESSORIES:

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- YK-88SSB second filter for dual-filter system.

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presstop

WARC NEWS CONTINUED GENERALLY FAVORABLE for the Amateur service, ARRL's WARC team reported from Geneva in early November. Some consideration has now been given to the Amateur bands, and on a world-wide basis, the service is holding its own.

160 Meters: An exclusive 40-kHz slot in Region 1 seems to be within reach; Region 2 looks like it'll keep 1800-2000 kHz with 50 kHz exclusive (probably the bottom end) and 150 kHz shared. Region 3 is also likely to end up with 200 kHz, possibly all on a shared basis though an "Amateur Exclusive" slot is possible.

40 Meters: As of now, the present 7-MHz world-wide allocations appear to have been maintained. However, international broadcasting's desire for new frequencies should still be considered to be a potential threat to a portion of that band.

10 Meters: It was agreed in committee to maintain the 28.0-29.7 slot as Amateur exclusive, worldwide. That committee is discussing the possibility of a separate HF-bands broadcasting conference in the future, and has appointed a large working group to recommend the ground rules for such a conference.

6 And 2 Meters: Present 50-MHz allocations do not appear to have been altered at this time. 144-148 MHz is also retaining the status quo, though some reservations ("footnotes") are believed to have been proposed for the band in some areas.

220 MHz: Amateur and Mobile services have been proposed as "Co-Primary" users for this band, in Region 2 only.

1215 MHz: A reduction to 1240-1300 was voted after considerable discussion by the Working Group. 1215-1240 would go to the Radio Navigation Satellite Service, with 1240-1300 a world-wide secondary allocation for Amateurs and the 1250-1260 MHz subband for Amateur satellite uplinks. The U.S.S.R. indicated it may later suggest a further reduction to 1250-1300 for Amateurs.

Microwave: 10.45-10.50 GHz has been proposed as a world-wide "Amateur Satellite" subband. The present 24 GHz band has been sustained by the Working Group, and the outlook is good for Amateur interests in the millimeter wave area above 40 GHz.

ASCII WAS ACCEPTED, paving the way for its use by Amateurs in the near future, at the Commission's meeting in Washington in mid-October. After considerable discussion, the Commissioners agreed to let Amateurs use ASCII for RTTY and instructed the staff to prepare rules for its implementation. During the discussion identification of Amateur stations operating on RTTY was a key issue, and the present requirement that RTTY users must identify using CW was the telling argument in getting the Commissioners to agree on the ASCII okay.

ANY AMATEUR LICENSE MODIFICATION will now require filing of a Form 610 or 610B, the FCC has decided. Previously, a change in mailing address (only) simply required a note to Gettysburg, though a station location change did require a Form 610 or 610B. For most Amateurs, mailing address and station location are the same. However, most Amateurs who have moved have not been submitting 610s, thus requiring a response from Gettysburg before their request for modification could be processed.

Modified Licenses now will be extended for a full 5-year term effective on the modification date; changes were effective November 12.

WIDEBAND FM ON ALL OF 6 METERS is to be proposed in a new Notice of Proposed Rule Making due shortly from the Commission; the Commissioners decided to throw the issue of 6-meter sub-bands squarely back to the Amateur fraternity by proposing that the band have no divisions by modulation. Strong opposition to the proposed deregulation can be exexpected from SMIRK and other 6-meter DXers.

AMATEURS WHOSE LICENSES or upgrades were obtained by fraud will soon be the subject of enforcement action, as a result of a recent Commission meeting. In addition, a number of Commission employees who received callsigns in a manner "inconsistent" with Commission procedures soon will be receiving new callsigns. About 50 people in the first category were identified as a result of the FBI investigation that began in Indianapolis back in June, 1977. Many of them, along with other Amateurs who may have taken part in fraudulent licensing schemes, will be facing revocation or suspension proceedings.

No Action Is Planned against non-FCC Amateurs who received special callsign treatment in the past.

LAUNCHING OF THE UNIVERSITY OF SURREY/AMSAT UK UOAST has been approved by NASA. The launch now is scheduled for mid-1981. Plans call for the inclusion of a synthesized telemetry system and a slow-scan TV camera on board the British OSCAR. The TV system will photograph cloud cover and the earth so that Amateurs can receive the pictures directly on their equipment.

A 23-CM LINEAR TRANSLATOR was put into service near San Jose, California in September, after extensive low-level tests. The translator was developed to provide new techniques for orbiting and terrestrial translators using microwave frequencies and to encourage greater 23-cm activity by Amateurs.

The following are excerpts from unsolicited letters and registration cards received from owners of the new TEN-TEC OMNI transceiver.

"I sold a Yaesu to buy this and am very impressed" —*WB5ULA*

"My first QSO with OMNI-A was LA1SV on CW and second was EA8SK on SSB." —*N2CC*

"Excellent rig, just as advertised." —*WB5TMD*

"Very pleased with performance. QSK feature very slick." —*WB0ELM*

"This is my 5th TEN-TEC transceiver in less than 2 years. I loved them all and still have 3." —*WB0VCA*

"Through the years I have had complete Drake and Collins stations. I tried a 544 Digital and liked it the best so decided to purchase the 546 OMNI-D Digital." —*WA4NFM*

"Your OMNI is the best rig I have had in 20 years of haming." —*K4IHI*

"As a owner of Collins rig, your OMNI-D is the best." —*K9JJL*

"I already have an OMNI-A, 544 and a TRITON IV. You may ask why I own so many TEN-TEC rigs. In case there is a great RF famine, I want to be ready!" —*WD4HCS*

"You guys really know how to turn on an old timer!" —*K8ELS*

"Best operating & most conveniences of any transceiver I've ever used." —*W6LZI*

"I like CW. Compared OMNI against IC701 (rcvr) and OMNI won hands down. XYL WD6GSB really enjoys rig on SSB. Finds rig is very stable and digital readout accurate." —*AC6B*

"Have checked it out on both modes from "top band" (160) all the way to 29 MHz. Terrific!!!!" —*W4DN*

"Works well, parts layout and design much better for any possible servicing than other ham gear. The Japanese hybrid sets can't compare to TEN-TEC for audio. Audio reports excellent without special speech processors, etc., to distort the signal." —*AG8K*

"I have been using the S-Line over 15 yrs and never thought anything could outperform it. I got the biggest surprise and THRILLED with this OMNI-D even though I have been a ham since 1936." —*KV4GD*

"This must be the greatest. I've spent enough money on final tubes to almost pay for this." —*KA4BIH*

"This transceiver was recommended to me by old time hams (Xtras) whom I have known for 40 yrs. Has excellent break-in." —*N6AVQ*

"Best package job I've ever seen! First licensed 6AAV in 1926. Now in operation—a sweetheart!" —*W7LUP*

"From a 32V2/SX115 to an OMNI is a big step!" —*K6YD*

"Receiver prominent—transmitter likewise—working comfortable—pleasing design." —*OE1FAA*

"First new rig for me in 10 years but seems to be very good." —*W5GBY*

"The best transceiver I ever used or owned." —*W3TS*

"I wouldn't swap my OMNI for anything on the market, regardless of price." —*WD0HTE*

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Model 545 Series B OMNI-A . . . $949
Model 546 Series B OMNI-D . . . $1119

To add your name to the fast-growing list of OMNI owners, see your TEN-TEC dealer, or write for full details.

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Pro-mark KWM-380 is a complete base station in a smartly styled desktop configuration. The look is unmistakably professional.

All in one unit. AC/DC power supply, speaker and accessory VFO function are built in.

Microprocessor frequency control provides rate selectable tuning in 1 MHz, 1 kHz, 100 Hz or 10 Hz increments. Frequency memory provides split VFO function (half-duplex) for transmit and receiver registers.

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Specifications†

Frequency Range: Tunable in 10 Hz steps.
Receive mode — 2.0 to 30.0 MHz, 0.5 to 2.0 MHz at reduced sensitivity.
Transmit mode — SSB or CW 160- thru 10-meter amateur bands.
Mode: SSB (voice and RTTY, either sideband selectable), CW, or AM (receive only).
Power requirements: 105, 115, 125, 210, 220, 230, 240, 250, ±5% V ac (Internal strapping option) 50-60 Hz, 12 V to 15 V dc (Connector strapping). 120 W input in receive max; 600 W input in transmit max.

†Subject to change without notice.

Frequency accuracy: Accurate to within ±5 Hz when the 39.6 MHz oscillator and the 455.0 MHz oscillator are set within ± 3Hz. Warm-up time is 10 min.
Frequency stability: Stability is within ±150 Hz over the temperature range of 0-50°C.

TRANSMIT PERFORMANCE

Output impedance: 50 ohms nominal.
Power output: 100 W PEP nominal from 1.6-30 MHz. In CW or RTTY, there is automatic turndown to 50 W after 10 seconds, 50% duty cycle, key down 15 minutes max.

With the optional blower kit, power is 100 W average, 50% duty cycle, key down 1 hour max at 25°C, ½ hour max. at 50°C for all modes.

Unwanted signal suppression: (minimum values below)

Carrier suppression	50 dB
Undesired sideband, 1 kHz ref	55 dB
Harmonics (all)	40 dB
Mixer products	55 dB

Third order distortion: 25 dB below each tone of a two-tone test.
Audio inputs: Microphone — low impedance type, internal strap for HI-Z. Line — 600 ohm input unbalanced impedance; level of 40 mV sufficient to produce full output.

Audio frequency response: Not more than 5 dB variation from 300 to 2400 Hz.

RECEIVER PERFORMANCE

Antenna impedance: 50 ohms

Sensitivity: Not more than 0.5 uV for 10 dB S+N/N at antenna input for SSB and CW, 2.0 to 30 MHz. Broadcast band attenuation is a nominal 30 dB.

I.F. and image rejection: Greater than 60 dB.

Selectivity: In operating modes of USB, LSB, CW, and AM.

BW at −3 dB (min)	BW at −60 dB (max)
2.1 kHz	4.4 kHz
*1.7 kHz	3.4 kHz
*360 Hz	1.25 kHz
*140 Hz	600 Hz
*6.0 kHz	25 kHz
8 kHz	50 kHz

*optional

Audio output: Not less than 3½ W into 4 ohm load at 1 kHz, at not more than 10% total harmonic distortion. Line audio output, −10 dBm nominal into 600 ohms.

Audio frequency response: Not more than 5 dB variation from 300 to 2400 Hz.

AGC: Audio output variance not more than 8 dB as the RF input varies from 2.0 uV to 100 mV open circuit.

Intermodulation distortion: Two signals spaced 20 kHz at a level of −10 dBm each will produce IMD down 50 dB min.

Size: 15.50" W (39.4 cm); 6.5" H (16.5 cm) (w/o feet), 7.5" H (19.1 cm) (w/feet); 18.00" D (45.7 cm).

Weight: 50 lbs. (22.7 kg).

comments

(Continued from page 6)

lightning protection

Dear HR:

While I was pleased to see two letters in the July issue complimenting my article on lightning protection, I must take exception to W6RTK's suggestion that the ground wire on a wooden pole be broken into short lengths, with small spark gaps between segments.

The main ground conductor on a wooden pole is one of the most important items in the protection system. The establishment of a low impedance path from the air terminal on top of the pole to ground is necessary to send the greatest possible percentage of the total lightning stroke current directly to ground. Although lightning will certainly jump across the small gaps recommended, the presence of these gaps will have a negative effect on the performance of the overall protection system. I don't know how to quantify the amount of degradation, but I don't think it's wise to take a chance. Also, even if lightning doesn't strike the pole, the breaking up of this ground lead may allow the entire antenna system to acquire a large static charge, possibly sufficient to cause minor equipment damage.

Mr. Caldwell is concerned that this ground wire may have some undesirable effects on the performance of the antenna system; when considered from the standpoint that the ground wire only makes the wooden pole look electrically equivalent to a metal tower, I think it is safe to say that this effect must be minimal.

John E. Becker, K9MM
Prospect Heights, Illinois

quartz crystals

Dear HR:

It has been brought to my attention that a statement in my article on quartz in the February issue was misleading if not incorrect. While it is true that crystals have high inductance and low motional capacitance, this is not the reason for high Q; Q is a ratio of the charge stored to the charge dissipated. In crystals the charge is primarily stored by the inductance, so Q is determined by the value of the motional inductance divided by the motional resistance:

$$Q = \frac{2\pi f K1}{R1}$$

Most of us associate high inductance with a large piece of iron wrapped in copper wire — an arrangement which is completely ineffective at rf. With quartz, you must rethink the problem.

Don Nelson, WB2EGZ
Voorhees, New Jersey

10 meters for satellite communications

Dear HR:

I write in response to W1GMP, who suggests (July *ham radio*) that we discontinue the use of 29.360 - 29.502 MHz for satellite communications. To put this into proper perspective, the frequency spectrum reserved for this purpose is no wider than two 25-kHz wide repeaters, adding both input and output bandwidths, yet serves fifty times as many stations on an intercontinental basis.

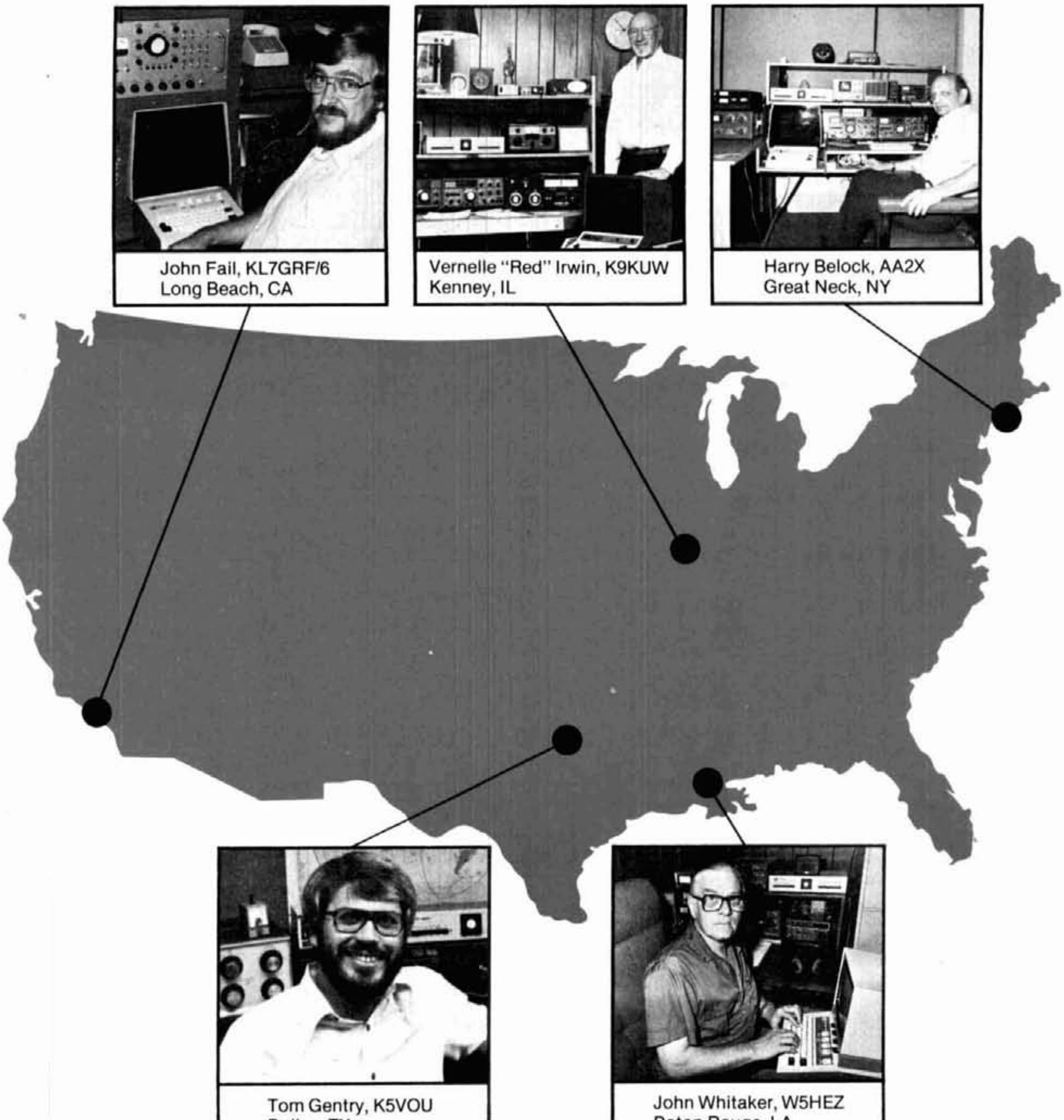
Apart from the aspect of communications, the use of 10 meters has been the basis of valuable research into sub-horizon communications, E_s and aurora detection and forecasting, and low-level signal techniques; it is also of great value in using Amateur Radio for teaching practical physics, geometry, trigonometry, astronomy, and mathematics through the use of a simple antenna and receiver.

The present maximum in the solar cycle will soon begin to decay, leaving only the satellite devotees to effectively occupy the high end of 10 meters. This will help prevent intrusion and takeover of the top part of the band, safeguarding it by regular, valuable usage.

Pat Gowen, G3IOR
Norwich, England

More Details? CHECK—OFF Page 126

WHEN OUR CUSTOMERS TALK... WE LISTEN.



John Fail, KL7GRF/6
Long Beach, CA

Vernelle "Red" Irwin, K9KUW
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CMOS 2-meter synthesizer

Construction details of a 2-meter synthesizer featuring choice of output frequency and CMOS design

Soon after joining the 2-meter fm crowd about three years ago, I learned how limiting "rockbound" mobile operation can be. At the same time, I had been wanting to learn more about frequency synthesizers, and designing and building one looked like the perfect answer to both needs. The result of my labor tunes from 146.000 to 147.995 MHz in 5-kHz steps and provides a variety of output frequencies to permit use with quite a number of rigs. This article will give full details on how the design was thought out, as well as how to build a copy. If you are interested in the subject of synthesizers, want to design one of your own but wonder where to start, or have soldering iron in hand ready to begin building, this article is for you.

design requirements

In addition to the above description, I expected the completed design to meet the following requirements:

1. Thumbwheel switch selection of receive and transmit frequencies
2. High output purity, at least 60-dB spur rejection
3. Self-contained; simple construction and circuitry
4. Minimum of test equipment needed to align and test
5. Minimum modification of 2-meter rig
6. Capable of mobile operation

operating frequency

The first choice was an operating frequency for the synthesizer. By looking at schematics and information on a number of common rigs, I learned that most use a receive crystal near 45 MHz. Transmit frequencies are less consistent and include $f/6$, $f/12$, $f/18$, and $f/24$. The circuit simplicity and purity goals ruled out the use of multipliers; therefore, I picked 45 to 49 MHz, or one-third the channel frequency.

synthesizer concepts

The next step was finding the most suitable method of synthesis. A literature search showed that the most popular type of synthesizer today is an elaboration of the phase-locked loop (PLL). **Fig. 1** shows the block diagram of such a system. In this, the VCO (voltage-controlled oscillator) is made to run at N times the reference frequency, f_R, which is normally fixed. Because of loop feedback, changing the divide ratio, N, also changes the VCO frequency to maintain the frequency relationship shown.

Because of its apparent simplicity, this kind of synthesizer seemed like an ideal approach for my design; but, after studying the logic required, some serious complications were obvious. The need for an i-f shift to go from transmit to receive made the design a real mess.

By Tom Cornell, K9LHA, RR2, Box 53A, Greentown, Indiana 46939

Several synthesizer schematics showed a different approach that looked like it had real promise; **fig. 2** shows the basic block diagram. The design freedom introduced by choice of crystal frequency allows the same variable divider ratio to be used in both transmit and receive. This results in great simplification of the overall synthesizer system.

practical design

This section will cover the more important design considerations.

VCO. The simplest form of VCO is a varactor-tuned oscillator, and a common circuit is shown in **fig. 3**. While a VCO can be developed by trial and error, it is much easier to calculate tank circuit values using the equations shown in the **Appendix** of this article.

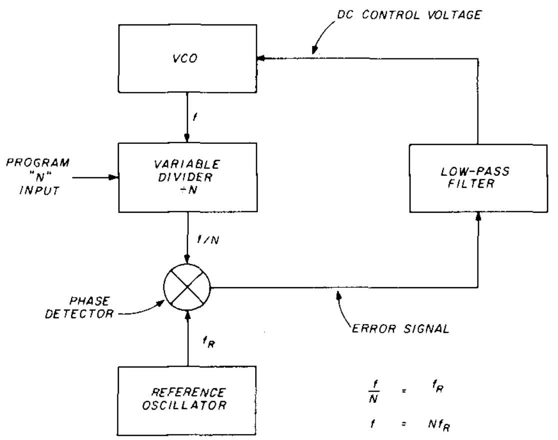


fig. 1. Block diagram of a simple synthesizer using a phase-locked loop.

Assuming a receiver i-f of 10.7 MHz, the minimum and maximum VCO frequencies are:

$$f_{max} = 147.995/3 = 49.33166\ MHz$$
$$f_{min} = (146.000 - 10.7)/3 = 45.1\ MHz$$

After picking some varactors (Motorola MV-2209) and suitable end point voltages, I was able to begin using the equations.

$$C_{max} = 50\ pF\ (1\ V) \qquad C_{min} = 28\ pF\ (6\ V)$$

Letting $C1 = 330\ pF$ and $C2 = 33\ pF$ and assuming 3 pF of transistor and stray capacitance, the total fixed capacitance, T, was 33 pF. The equations then gave $C_p = 88.2\ pF$ and $L = 0.19\ \mu H$. The circuit was built using these values and worked just about exactly as intended.

Oscillator/mixer. Design of the oscillator/mixer has quite an impact on the variable divider, and, after much study I decided on:

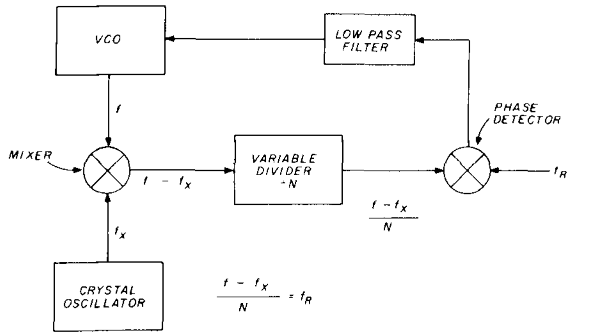


fig. 2. Diagram of a mixing-type synthesizer where the same divide ratio can be used for either transmit or receive, with the change in frequency accomplished by shifting the crystal oscillator frequency.

$$N \text{ at } 146.000\ MHz = 400$$
$$f_{xtal}(TX) = 48\ MHz$$
$$N \text{ at } 147.995\ MHz = 799$$
$$f_{xtal}(RX) = 48 - i\text{-}f/3$$

These numbers are a good illustration of synthesizer operation, and it might help your understanding if you plug them into the equation shown in **fig. 2**.

After selecting these points, design of the oscillator and mixer was relatively uncomplicated. A dual-gate MOSFET with an untuned output circuit was selected as the mixer, and two separate oscillators were used for receive and transmit.

Variable divider. The variable divider design has a lot to do with the complexity of a synthesizer circuit, and an intelligent choice is very important. Some of the divider requirements have already been covered, and the remaining important characteristic is speed. For this kind of synthesizer, the highest divider speed is:

$$f_{in}(max) = N_{max} \times f_R = 1331.66\ kHz$$

After studying the above requirements and the data sheets of a number of prospective devices, I chose the RCA CD4059 as the most suitable. This IC is a CMOS, 5-stage, BCD-programmable counter which has exactly the capability needed in this design. Additional factors favoring this choice were the inherent properties of CMOS. This logic family offers greater circuit density than TTL. It also consumes far less power, which incidentally means that there will be much less high-frequency energy produced to cause interference in other parts of the synthesizer.

Phase detector. A second CMOS IC, the CD4046, was selected as the phase detector. This is a special purpose device designed specifically for such an

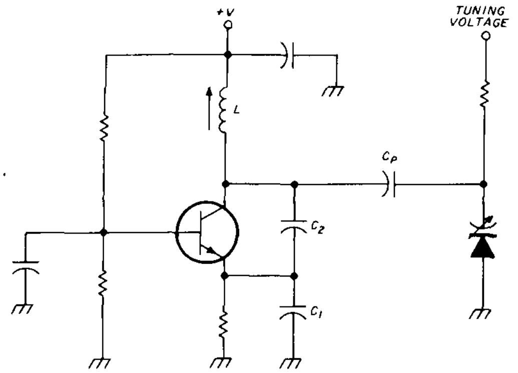


fig. 3. Schematic diagram of the basic VCO oscillator. The computations for the component values are shown in the appendix.

application. In addition to the phase detector, the CD4046 has a lock detection circuit that will be described later.

Lowpass filter. To keep the phase detector switching products from frequency modulating the VCO, a lowpass filter is inserted in the tuning line to the VCO. Because of its inherent phase shift, this filter is also to a large degree responsible for determining the PLL stability. And, while there are formulas for calculating filter values, I felt that the complex relationships involved would dictate some "cut and try" anyway, so that was the design approach I used. A circuit from another synthesizer was used as the starting point, and experimentation helped to determine the final values.

Fig. 4 shows the basic circuit. R1 and R2 together with C1 establish the main cutoff frequency, which must be somewhat less than the system reference frequency. C2 and C3 must be several times smaller than C1 to avoid instability and are included to add to the filtering action. R3 dampens the filter to control overall synthesizer system stability. My design method was to listen to VCO harmonics on an fm receiver and to make a sudden change in synthesizer frequency. R3 was then adjusted until the system demonstrated stable transient behavior.

Reference-frequency circuit. The reference frequency of a synthesizer is normally equal to the channel spacing. For this design, the spacing is 5 kHz and the reference frequency is 5/3 kHz, since the VCO operates at one-third the output frequency. For reasons of stability and accuracy, crystal control is usually considered a must.

After looking at several alternatives, I chose a crystal frequency of 2.56 MHz and a CD4060 CMOS oscillator/divider IC to generate the 5-kHz signal. A CD4027 dual J-K flip-flop was then used to divide by three to get 5/3 kHz.

System tests. When all of the preceding synthesizer circuits were hooked together, Murphy put in his first appearance. Switching from receive to transmit invariably causes the loop to drop out of lock, and output signal purity was awful. The first problem resulted from something I overlooked; transmit/receive switching caused the input frequency to the variable divider to jump by i-f/3, which could exceed the reliable counting speed of that IC. Adding a frequency-shifting circuit to the VCO to retune the tank for the receive and transmit ranges solved the problem.

A buffer amplifier was placed between the VCO and mixer because it was found that the crystal oscillator was the cause of spurs in the VCO output, and the oscillator signal was getting to the VCO through the mixer. Purity now measured better than 60 dB, so I figured the design was adequate. After finishing the rest of the circuits, doing printed circuit artwork, building the synthesizer and hooking it up to my rig, I learned that Murphy doesn't give up very easily! The oscillator/tripler of my rig's receiver degraded purity to only 55 dB. Since the crystal oscillator was still the source of unwanted signals, lowering the output of the crystal oscillator was the logical way to reduce

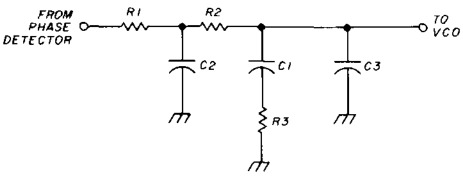


fig. 4. Schematic diagram of the lowpass filter that is inserted between the phase detector and the VCO.

the spurs. Unfortunately, this change resulted in insufficient drive to the variable divider, and new circuit boards were required to add the extra amplifier between the mixer and divider needed to bring the level back up. But the effort was worth it. Purity at the synthesizer output improved to about 75 dB, and at the receiver mixer 65 dB.

Output divider. To obtain the various output frequencies needed from this synthesizer, I made the logical choice of a divider stage driven by the VCO. Of the frequencies listed earlier, only f/18 presented any problems. The others could easily be derived by dividing the VCO frequency by 2, 4, or 8. Not providing an f/18 output did seem to be a compromise of the original requirements, but the improvement in circuit simplicity looked like a desirable trade-off, especially since only one rig that I knew of (Regency HR-2B) used f/18. Builders are still encouraged to consider this synthesizer design even though they may have to substitute for a small portion of the circuitry in order to get the precise frequencies.

operation

This section will briefly describe the function of the major circuit elements shown in **fig. 5**. The VCO is composed of Q3 and surrounding components. CR1 is the tuning varactor. Three buffer amplifiers and U6, the output divider, follow up the VCO. Buffer Q4 acts as a squaring amplifier to convert the sinewave VCO output to a squarewave suitable to drive NAND gate buffers U5D and U5B. U5D amplifies the f/3 VCO signal, and U5B isolates this output from U6 and its back-fed divider products. Any of the four divider outputs from U6, as well as f/3 from U5D, can be jumper-selected as the synthesizer receive and transmit output frequencies. NAND gates U5A and U5C actually supply these outputs to the transceiver.

To the left of the VCO are Q2 (the MOSFET isolation buffer) and Q1 (the VCO frequency-shifting transistor). During receive, Q1 is turned on and places C14 across the VCO tank. A TTL logic-level signal at the Q1 collector also serves to turn off the transmit output during receive.

Below Q1 are the two crystal oscillators, which together with mixer Q7 were added to simplify the variable divider design. Either Q5 or Q6 is turned on by application of supply voltage. LC tanks are in series with both crystals to allow slight adjustment of actual oscillator frequency.

The mixer output signal is amplified by Q8 and Q9 to an adequate level to drive U1, the variable divider. As shown, U1 looks deceptively simple; actually, it is very busy inside. The divide ratio is loaded from the switch inputs, U1 counts down N pulses to zero, produces a single output pulse, and then reloads the divide ratio to begin the cycle again.

All the frequency selector switches are shown in two groups below U1. Diode OR-gates between the transmit and receive switches isolate the two groups of switches and allow selection of transmit or receive operation by mere application of supply voltage. Toggle switches are used for both MHz and 5-kHz ranges, since they are all that is necessary and are much cheaper than thumbwheels.

To the right of U1 is the phase detector, U2. This IC provides a tuning voltage for the VCO at pin 13 that is filtered by the lowpass filter composed of R23, R24, R25, C27, C28, and C29. At pin 1 of U2 is the lock-detector output, which is normally high when the PLL is locked and goes low in a series of pulses when out of lock.

Q10 and Q11 amplify and stretch the lock detector pulses of U1 to produce a continuous logic-level signal that both lights the UNLOCKED indicator and shuts off the synthesizer transmit output. C26 serves to slightly delay the turn-on of Q10 so that slight disturbances of the loop don't shut down the transmitter.

ICs U3 and U4 generate the synthesizer reference frequency. U4 contains a crystal oscillator running at 2.56 MHz and a divide-by-512 circuit (in this application) to produce 5 kHz. U3 then divides this frequency by three to produce 5/3 kHz. C23 allows for exact adjustment of oscillator frequency.

In the power supply, three-terminal regulators U7 and U8 provide 5 and 8 volts respectively. Control gates Q12 and Q13 are driven by the transceiver push-to-talk line and select either receive or transmit operation of the synthesizer by providing logic supply voltages. The input LC filter, L8 and C49, serves to protect the synthesizer from transients and noises from the car's electrical system.

construction

The two synthesizer circuit boards may be assembled in any order, with the exception of the CMOS IC's which should be saved to the last to avoid damage from static electricity. (See **figs. 6** and **7**, respectively, for the circuit board pattern and parts placement diagram.)* Clip a ground wire from the soldering iron to the ground copper of the board when soldering these ICs.

Fig. 8 shows construction of the VCO coil.† Since this coil form was chosen for reasons of mechanical rigidity, you will need to cover the completed winding with Q-dope or airplane cement to ensure coil stability. Tighten the shield can to the base by making small impressions on at least two sides of the can into the plastic base with a center punch.

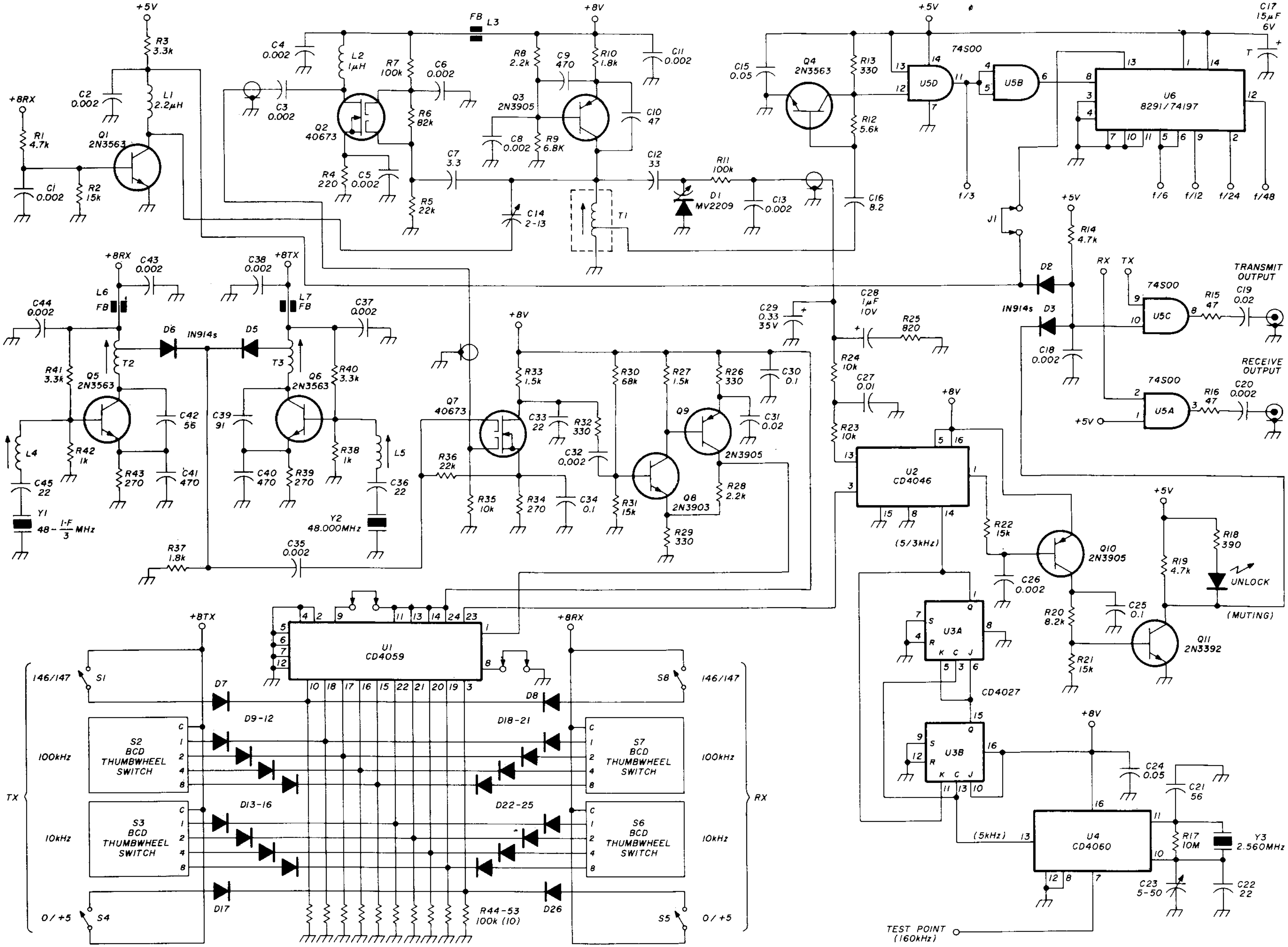
The plastic-molded coils used in the crystal oscillators may prove hard to find, and you can probably substitute most any good-quality coil forms of suitable size. Information on the number of turns is shown in the parts list.

L8, the supply filter choke can be made, or a suitable commercial part used. To make the choke, cut the heads off some small-diameter nails and tape them together to form a core roughly 30-mm (1 1/4-inch) long by 5-mm (3/16-inch) in diameter. Wind a coil of about one-hundred turns of no. 22-26 AWG (0.6-0.4 mm) wire over the core. Finish by covering with electrical tape. Form the leads to fit the circuit board and mount to the board with ordinary string or wire wrapped over the body of the coil.

Install jumpers to select the correct output frequencies for your rig. For receive, connect a small piece of insulated wire from pin 2 of U5 to one of the divider outputs. For transmit, the jumper goes from

*The circuit boards and many components to build the synthesizer are available from Radiokit, Box 429, Hollis, New Hampshire 03049.

†The VCO coil form is a standard 10-mm i-f transformer form. If you are unable to find such a part, it may be purchased from the author for $1.00.


+5V
+8V
+8RX
+8TX
FB
L3
R3 3.3k
L1 2.2μH
C2 0.002
R1 4.7k
C1 0.002
R2 15k
Q1 2N3563
C4 0.002
L2 1μH
R7 100k
C6 0.002
C3 0.002
Q2 40673
R6 82k
R4 220
C5 0.002
R5 22k
C7 3.3
C14 2-13
R8 2.2k
C9 470
R10 1.8k
C11 0.002
Q3 2N3905
C8 0.002
R9 6.8K
C10 47
C12 33
R11 100k
D1 MV2209
C13 0.002
T1
C15 0.05
Q4 2N3563
R13 330
R12 5.6k
C16 8.2
74S00
U5D
U5B
U6 8291/74197
C17 15μF 6V
f/3
f/6
f/12
f/24
f/48
J1
R14 4.7k
D2
D3
IN914s
C18 0.002
RX
TX
U5C
R15 47
C19 0.02
TRANSMIT OUTPUT
U5A
R16 47
C20 0.002
RECEIVE OUTPUT
C43 0.002
L6 FB
C44 0.002
D6
IN914s
D5
C38 0.002
L7 FB
C37 0.002
T2
T3
R41 3.3k
Q5 2N3563
Q6 2N3563
R40 3.3k
C42 56
C39 91
L4
R42 1k
R43 270
C41 470
C40 470
R39 270
R38 1k
L5
C45 22
C36 22
Y1 48 - 1-F/3 MHz
Y2 48.000MHz
R37 1.8k
C35 0.002
Q7 40673
R33 1.5k
C33 22
R32 330
C32 0.002
R36 22k
R35 10k
R34 270
C34 0.1
R30 68k
R31 15k
R27 1.5k
R26 330
C30 0.1
Q9
C31 0.02
2N3905
Q8 2N3903
R28 2.2k
R29 330
C29 0.33 35V
C28 1μF 10V
R25 820
R24 10k
C27 0.01
R23 10k
U2 CD4046
(5/3kHz)
R22 15k
Q10 2N3905
C26 0.002
R18 390
R19 4.7k
UNLOCK
(MUTING)
C25 0.1
R20 8.2k
R21 15k
Q11 2N3392
U3A
U3B
CD4027
C24 0.05
C21 56
U4 CD4060
(5kHz)
R17 10M
Y3 2.560MHz
C23 5-50
C22 22
TEST POINT (160kHz)
U1 CD4059
146/147
S1
S8
D7
D8
D9-12
D18-21
S2 BCD THUMBWHEEL SWITCH
S7 BCD THUMBWHEEL SWITCH
100kHz
TX
RX
D13-16
D22-25
S3 BCD THUMBWHEEL SWITCH
S6 BCD THUMBWHEEL SWITCH
10kHz
D17
D26
0/+5
S4
S5
R44-53 100k (10)

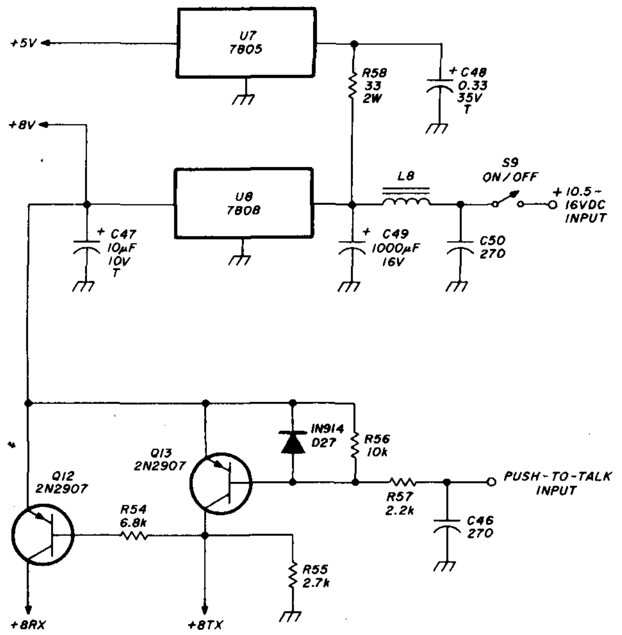


fig. 5. Schematic diagram of the CMOS synthesizer (opposite). The individual portions of this schematic are discussed in the text. All small-valued capacitors are NPO ceramics. Power supply is seen above; parts list is below.

L4 11 1/2 turns 6.5 mm (1/4 inch) diameter, slug-tuned, close-spaced molded plastic form

L5 10 1/2 turns, same as L4

L8 see text

T1 5 1/2 turns, tap at 2 turns, no. 32 AWG (see **fig. 8**).

T2 6 1/2 turns, tap at 1 3/4 turns, spaced 1 wire diameter, no. 26 AWG (0.4-mm) wire, plastic-molded 6.5 mm (1/4-inch) diameter form, J-iron core

T3 same as T2 except aluminum core

Y1 f = (48 – i-f/3) MHz, see text (44.4333 MHz, i-f = 10.7), same as Y2 except frequency

Y2 48.000 MHz, series mode, third overtone, 0.0025 per cent tolerance, HC-18/U case

Y3 2.5600 MHz, parallel mode, fundamental, 32-pF load, HC-6/U case with wire leads, 0.005 per cent tolerance

pin 9 of U5 to a divider output. Note that the receive frequencies actually contain an i-f offset and are really (f – i-f)/3, (f – i-f)/6, etc. If the f/3 receive option is used, install jumper J1 as shown in **fig. 7**. This will turn off U6 during receive and eliminate some low-level subharmonic spurs U6 produces. If f/3 is not used for receive, install jumper J2 instead to allow U6 to operate in both transmit and receive.

Temporarily install interconnecting wires between the two boards to allow circuit alignment. Connect the following: 5 volts, 8 volts, 8 volts RX, VCO tuning voltage (coax), and the VCO buffer output (coax). The synthesizer will operate on 146.000 MHz in this condition.

alignment

Alignment of the synthesizer requires the following equipment: a dc voltmeter (VTVM or high-input impedance), and a-m/fm radio (a portable set is fine), the diode detector probe shown in **fig. 9** , and a regulated power supply (preferably current limited). Other useful equipment includes a frequency counter (50 MHz, high-input impedance), a grid-dip meter, and a general-coverage receiver.

Connect the synthesizer to a 12-volt supply; it should draw approximately 125 mA. Next, check the 5- and 8-volt supplies, which should be within 5 per cent of the correct value. Test the 8V-RX and 8V-TX lines. With the push-to-talk line open, 8V-RX should read 8 volts and 8V-TX about a volt. Grounding the push-to-talk input should bring the 8V-TX up to 8 volts and drop 8V-RX to zero.

Next, some kind of check on the 2.56-MHz oscillator should be made. There are several possibilities, including connecting the diode probe to pin 7 of U4 (dc output voltage should be around 6.5 volts); measuring the same point with the counter (it should be 160.000 kHz, adjust with C23); and listening with the a-m/fm receiver (antenna near U4) at 640 or 1280 kHz, or listening with the communications receiver at 2.560 MHz. Alignment can be by the counter or by comparing one of the U4 divider products (1.28 MHz, 640 kHz, etc.) with a known frequency such as an a-m radio station.

The easiest method for testing the receive crystal oscillator is to hold a grid-dip meter near T2 as the slug is adjusted. An fm receiver tuned to the second harmonic of the oscillator can also be used, as well as the diode detector probe connected across R37. Adjust the slug of T2 for maximum output and then turn it toward the top of the coil until the dc voltmeter connected to the diode probe reads 0.25 volts. Ground the push-to-talk line, and then make the same adjustment and check on the transmit crystal oscillator and T3. If you are able to use an aluminum slug in T3, remember that, compared with an iron slug, it works backwards.

If the oscillators refuse to run, temporarily bypass the base to ground with a 0.001-to-0.002 µF capacitor. You can then find out what the free-running frequency of the oscillator is and make corrections in T2 or T3 or the value of C39 or C42. If you can adjust the oscillator in this condition to the crystal frequency, then crystal control should work, too.

If you have a high-impedance counter available, connect it across R37. Adjust the slugs of L4 and L5 to fine-tune the frequency of each oscillator. Without a counter, you may be able later to arrange some type of on-the-air check to adjust frequency.

Measure the VCO tuning voltage with your dc volt-

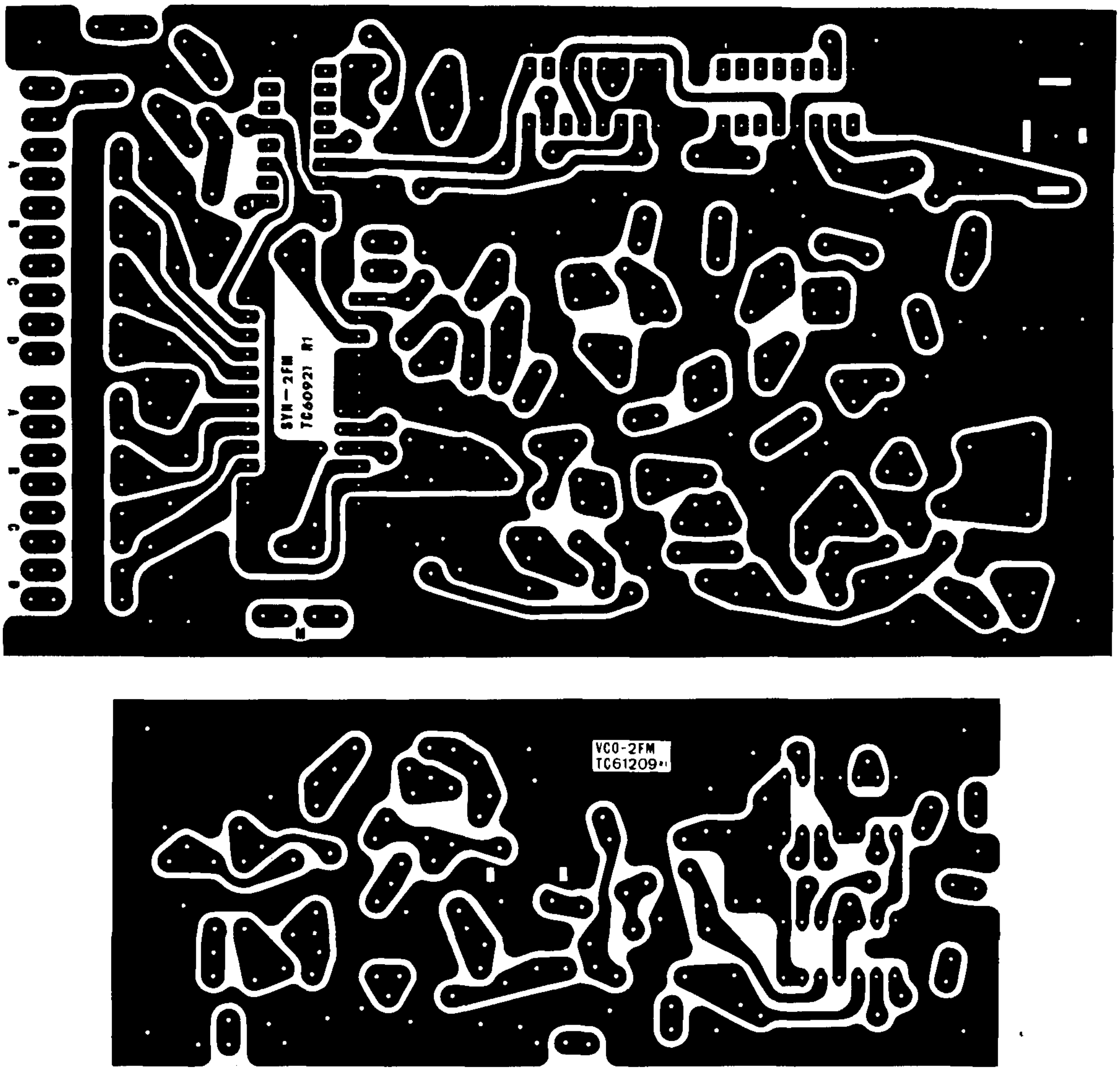


fig. 6. Foil pattern for the synthesizer board (above) and the VCO board (below).

meter and ground the push-to-talk input. If the voltage is 8 volts, turn the slug of T1 toward the top of the can. If the voltage is, instead, zero, turn the slug into the coil. Adjust for a final reading of 1.0 volt. At this point, you should be able to hear the VCO harmonic at about 97.3 MHz on the fm receiver.

If the tuning voltage cannot be adjusted, the problem may be one of several things. A high tuning voltage is caused by a high VCO coil inductance, and a low voltage by low inductance. If slug adjustment is insufficient, the coil turns may need to be changed; the turns can be spread apart or squeezed together. As an alternative, C10 can be changed slightly to get the right tuning range. A dead VCO, mixer, crystal oscillator, or mixer output amplifier will also cause the tuning voltage to go to 8 volts.

Once the VCO works correctly in transmit, remove the ground from the push-to-talk line. Adjust C14 for a tuning voltage of 1.0 volt and verify operation by listening to the VCO's second harmonic on the fm radio at 90.2 MHz.

final construction

The synthesizer boards may now be boxed to your preference. My experience has indicated two possible critical areas: the VCO board will very likely require a complete shield to keep transmitter rf away from the VCO. For the same reason, wires between the two circuit boards should be kept away from the power/control wires entering the box from the transceiver. Once the unit is assembled, I usually recommend a touch-up alignment of the VCO and crystal oscillators to compensate for any stray capacitance added by the case. When the frequency selector switches have been connected, you will have your first opportunity to check full operation of the syn-

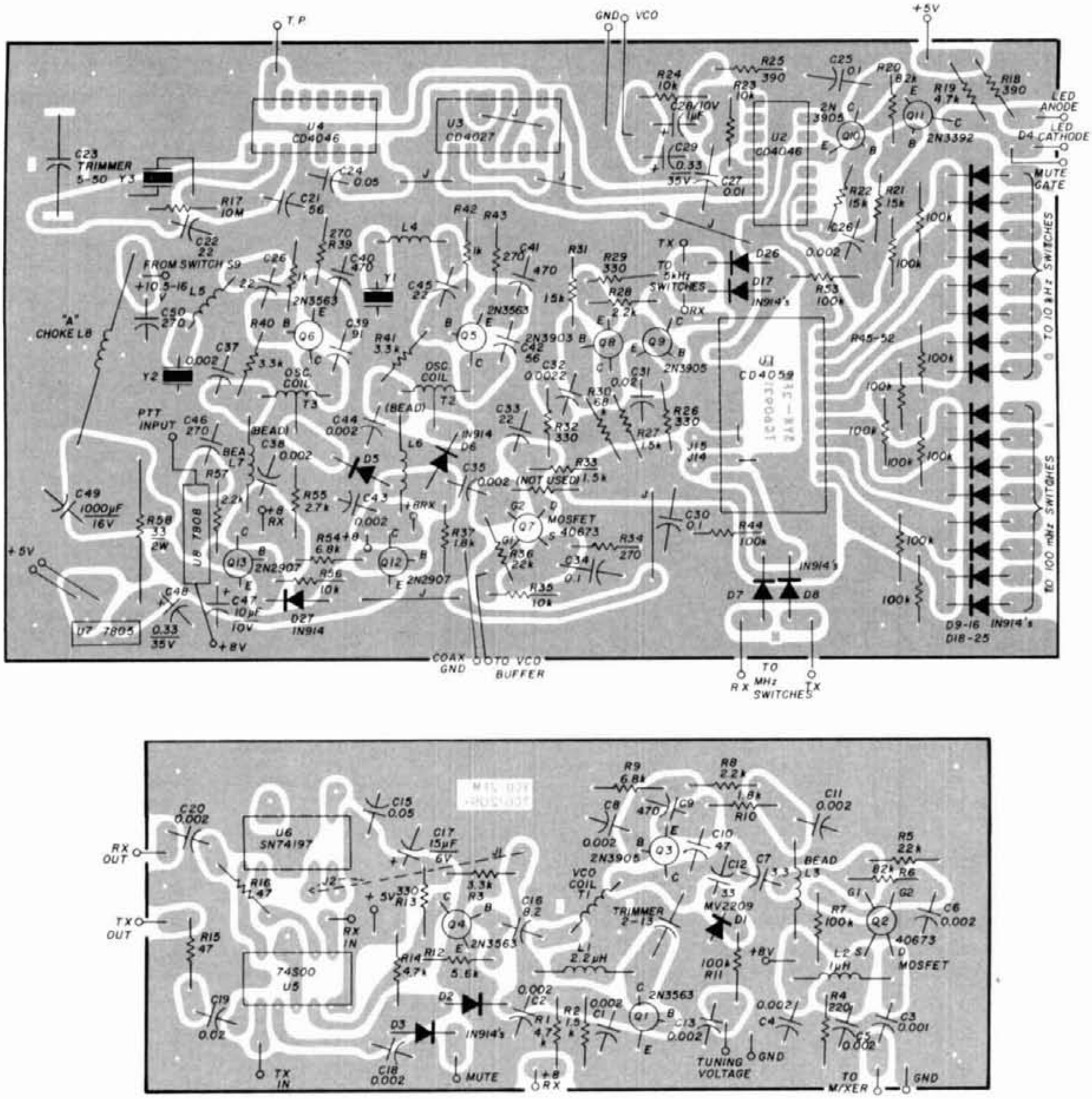


fig. 7. Component placement diagrams for the synthesizer board (above) and the VCO board (below).

thesizer by listening to it on the fm radio or your 2-meter rig.

connecting your rig

Fig. 10 shows the circuits I used to couple the synthesizer to my rig's receive and transmit oscillators. Install these right at the crystal sockets of your rig, drill holes, and mount two coax connectors on the rear of your rig. Connect up the entire system using coax cable. Make sure the inductor tunes to the transmit crystal frequency with the capacitors of your oscillator circuit, and adjust the resistor (470 ohms in **fig. 10**) to keep the transmit oscillator from running on its own (you will be mighty unpopular on 2 meters if it does).

Next, connect the power and push-to-talk lines from your rig to the synthesizer. Shielded cable is strongly recommended for this purpose.

The synthesizer and transmitter should now be thoroughly tested in the transmit mode, first on a dummy load and then on an antenna. Any rf that gets into the VCO can cause instability and flickering or illumination of the LED indicator. The dummy load check will determine if your rig is feeding rf back from its oscillators into the synthesizer. This condition may be corrected by insertion of a lowpass filter, having a cutoff frequency just above the synthesizer's output frequency, in one or both of the synthesizer output lines. The antenna test is somewhat more complicated in that certain antenna types, especially the gutter-mount and magnet-base variety can cause appreciable ground currents to flow on the coax. The solution in these cases is wire rerouting away from the synthesizer and possibly within the synthesizer, and use of additional shielded wires plus the VCO shield.

additional possibilities

The simple BCD programming of this synthesizer makes it easily adaptable to some interesting fre-

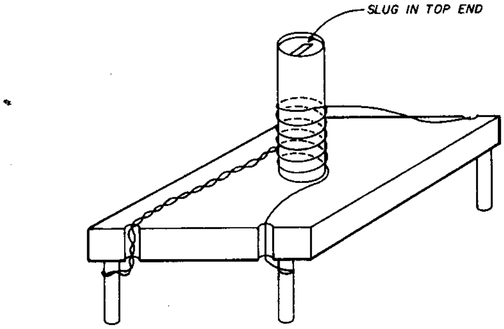


fig. 8. Detailed diagram of the VCO coil. The form is a standard 10-mm shielded slug-tuned form.

quency control methods. Replacing the switches with an up-down counter will allow scanning as well as LED frequency readout. A memory can be used to store favorite channel frequencies, or a microprocessor can be added for all kinds of control functions including scanning all channels, a group of channels, or those you preset. Your imagination is the limit.

When this synthesizer was developed, I intended to build at least one unit to cover 150 to 159.995 MHz to tune some of the vhf mobile channels, That is the

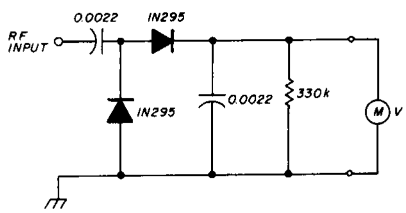


fig. 9. Schematic diagram of an rf diode probe suitable for tuning the synthesizer.

reason for the two jumpers beneath U1. Although I've not had time yet to try this, expanded coverage might appeal to you. This design will, therefore, permit operation over the expanded 2-meter band as proposed by the FCC.

final comments

I hope that this article has proven valuable to you. Should you have questions about the design or problems in construction, please feel free to write me, but do enclose an SASE.

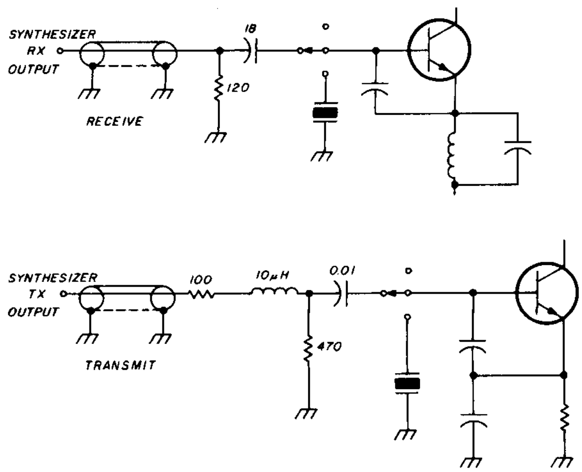
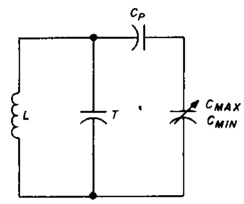


fig. 10. Diagrams of the interface circuits between the synthesizer and the rig's receive and transmit crystal oscillators.

I'd like to offer my thanks to those who helped me in this project: Dib, K9HLG, and Tom, W9IJ, for their encouragement, counsel, and interest in the project; to Russ, K9AYD, for his valuable ideas on logic and synthesizer design; and to Bill, WA9GUY, for his mechanical help and engraving of the front panel.

appendix



$$\alpha = \left(\frac{f_{max}}{f_{min}}\right)^2$$

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

$$b = \frac{(1-\alpha)\,[T(C_{max} + C_{min}) + C_{max}\,C_{min}]}{(1-\alpha)\,T + C_{max} - \alpha C_{min}}$$

$$C = \frac{(1-\alpha)\,T\,C_{max}\,C_{min}}{(1-\alpha)T + C_{max} - \alpha C_{min}}$$

$$C_T = T + \frac{C_p\,C_{max}}{C_p + C_{max}}$$

C_T = *total tank C at* f_{min}

$$L = \frac{1}{(2\pi f_{min})^2\,C_T}$$

$$= \frac{25330.34\,\mu H}{(f_{min})^2 C_T} \qquad f \text{ in MHz}, \quad C_T \text{ in pF}$$

ham radio

environmental aspects of antenna radiation

How to calculate approximate near-field radiation levels to meet existing environmental standards

At present there's a great deal of interest and controversy regarding non-ionizing electromagnetic radiation and its effect on the environment. This issue, of course, affects Amateur Radio. Suggestions have been made that all nonionizing electromagnetic radiation be eliminated from residential areas, or that such radiation be limited to levels that would make Amateur Radio operation impossible.

Some groups, in a wave of hysteria, are attempting to make *all radiation* illegal. As in most situations of this type, when one looks at the facts, the picture becomes clearer.

In a report by the U.S. General Accounting Office dated March 29, 1978,[1] it states that 10 mW/cm^2 is the maximum level to which a human should be exposed for 6 minutes per hour, and that 1 mW/cm^2 is the maximum continuous exposure. In other words, to be completely safe, one should stay at levels of 1 mW/cm^2 or less. These levels are recommended by the American National Standards Institute (ANSI) for frequencies between 10 MHz through the microwave region.

analysis

I have calculated the approximate separation distances between the radiation source (Amateur antennas) and humans to meet the recommended levels in reference 1. These data are shown in **figs. 1A** through **1D** for four Amateur antennas: half wave, quarter wave, eighth wave, and sixteenth wave. Parametric curves show the input power to the antenna at two field-strength levels, 1 mW/cm^2 and 10 mW/cm^2.

Looking at **fig. 1A**, one can see that if an Amateur operates on 7 MHz using a half-wave antenna with 100 watts input, the antenna must be at least 4.6 meters (15 feet) from any human to keep the field strength at 1 mW/cm^2 or less. At a power of 1 kW input to the antenna, the field at 4.6 meters (15 feet) increases to 10 mW/cm^2. Thus it's necessary to move the antenna a distance of 7.3 meters (24 feet) from any human to reduce the field to 1 mW/cm^2.

If the antenna has 10 dB gain in one direction, the equivalent antenna input would be 10,000 watts instead of 1000 watts. The field at 7.3 meters (24 feet) would increase to 10 mW/cm^2 in the direction of the antenna gain.

The apparent free-space field strength near any antenna can be approximated by:

$$p_{field} = \frac{755 L^2 P_{ant} K}{Z_{ant} \lambda^4} \ watts/meter^2 \qquad \textbf{(1)}$$

where L = length of antenna (meters)

P_{ant} = power input to antenna (watts at Z_{ant})

Z_{ant} = input impedance of antenna (ohms)

λ = wavelength in meters ($300/f_{MHz}$)

$K = \frac{(\alpha r)^4 + 3(\alpha r)^2 + 5}{(\alpha r)^6}$ (values in **table 1**)

$\alpha = 2\pi/\lambda$

r = distance from the antenna (meters)

f = frequency

After the apparent free-space field strength has been calculated in watts/meter^2 it can be converted to mW/cm^2 by multiplying the calculated value by 0.1. In other words, 100 W/m^2 is the same as 10 mW/cm^2. If the field strength in volts/meter is desired, the following expression can be used:

$$E_{field} = \sqrt{120\pi (P_{field}\ W/m^2)} \ volts/meter \qquad \textbf{(2)}$$

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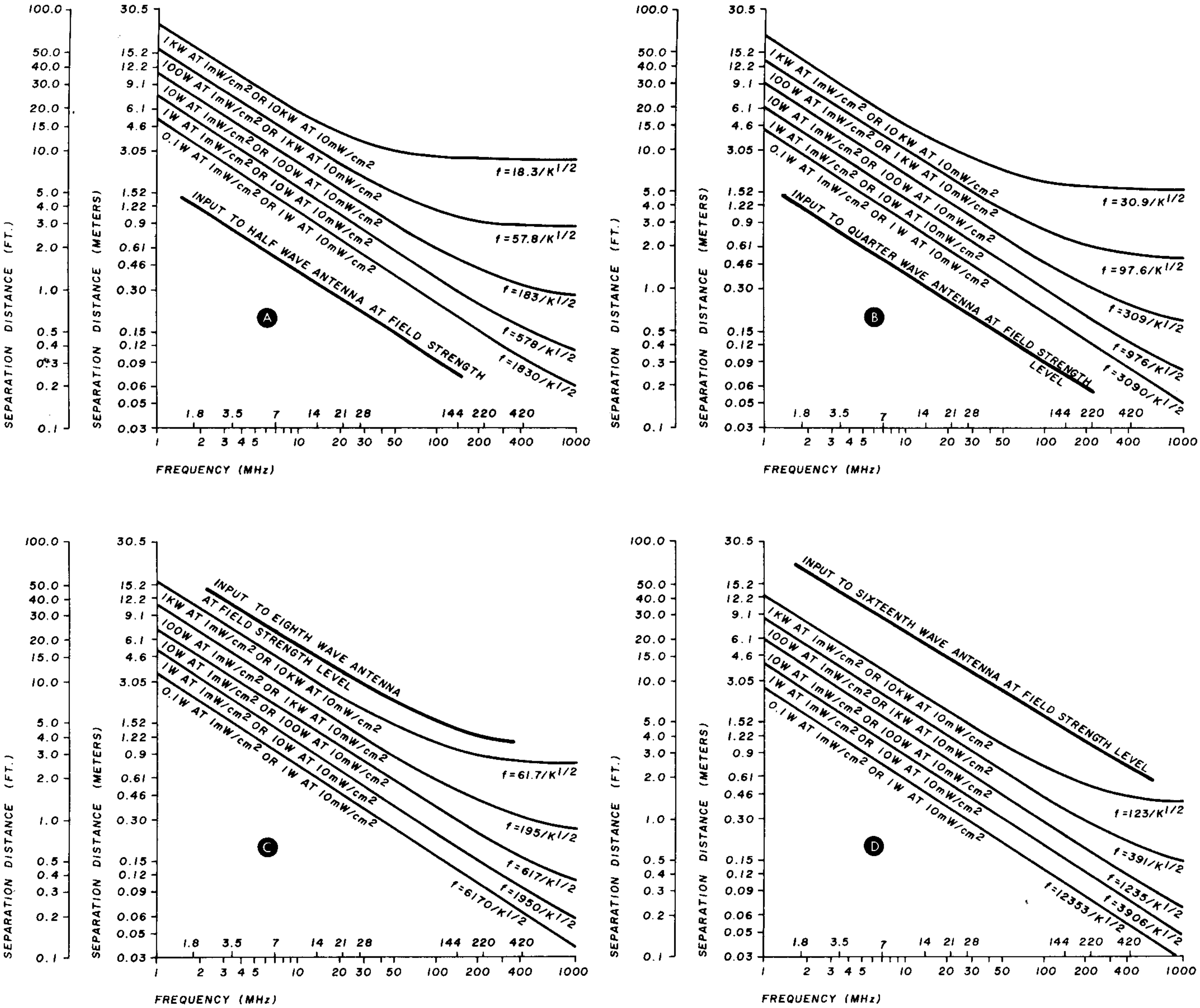


fig. 1. Separation distance as a function of frequency for recommended input power to four Amateur antennas: (A) one-half wavelength; (B) one-quarter wavelength; (C) one-eighth wavelength; and (D) one-sixteenth wavelength.

where 120π is the impedance of free space. Using this expression, 10 mW/cm^2 is the same as 194 volts/meter field intensity.

K values (**eq. 1**) are shown in **table 1** as a function of r/λ, the ratio of the distance from the antenna in meters to the wavelength in meters. Using this table and **eq. 1** for P_{field}, it's possible to calculate the fields at various distances from an antenna to obtain the apparent free-space field intensity. Otherwise, use **fig. 1** to make sure that your antenna is always at a separation distance with less than 1 mW/cm^2 field intensity. (A mathematical derivation of **eq. 1** is available from *ham radio* upon receipt of a self-addressed stamped envelope).

practical considerations

There should be no problem for most Amateurs in installing an antenna away from houses and areas occupied by humans, except for the 160- and 80-meter bands. In these cases it may be necessary to limit power input to the antenna if necessary distances can't be maintained. The real difficulty lies in the operation of handheld portables above 25 MHz. If adequate separation from the body is maintained, it will be difficult to talk into a handheld unit. You'll have to decide if the risk is worth the exposure.

Mobile operation above 25 MHz should be no problem if simple precautions are followed. **Tables 2** table 1. Proximity coefficient K as a function of the ratio of distance from antenna, r, to wavelength, λ.

ratio of distance from antenna, r, to wavelength, λ r/λ	proximity coefficient, K
0.01	81,000,000.000
0.015	7,170,000.000
0.02	1,282,000.000
0.04	20,620.000
0.06	1,890.000
0.08	364.700
0.10	103.000
0.15	12.300
0.20	3.110
0.25	1.186
0.30	0.640
0.35	0.388
0.40	0.252
0.45	0.187
0.50	0.135
0.60	0.089
0.70	0.0596
0.80	0.0452
0.90	0.0346
1.00	0.0272
1.20	0.01872
1.40	0.01368
1.60	0.0104
1.80	0.0080
2.00	0.00654
2.50	0.00416
5.00	0.001028

and **3** show a summary of approximate operating distances that should prevent overexposure for most Amateur installations.

I'd like to emphasize that, in this article, I make no attempt to account for the shielding effects of buildings or the susceptibility of humans to radiation at any given frequency. The field levels are simply calculated at each frequency shown. It may well be that 1 mW/cm^2 is more of a hazard at 420 MHz than at 1.8 MHz. Such matters will have to be explored by medical research.

The data presented here will allow Amateurs to estimate field-strength levels from the antennas described in a manner that will meet present recommended criteria. Furthermore, the data will provide ammunition with which to fight pressure groups who are trying to abolish Amateur Radio!

addendum

The effects of radiation from electronic equipment on the environment has become a hot issue of late. The FCC has issued a Notice of Inquiry (NOI), General Docket 79-144 (June 15, 1979) which states in paragraph 33:

"It may be desirable for the Commission to consider the need for applying to the subjects of its jurisdiction one of the existing safety criteria, such as the 10 milliwatt per square centimeter (10 mW/cm^2) short-term exposure limit used by ANSI and OSHA . . ."

Furthermore, ANSI is considering reducing this level to 1 mW/cm^2. What does all this mean to Amateur Radio? The answer is presented in the article above.

If you are concerned you'll want to file comments to the FCC/NOI mentioned above before the December 15, 1979 deadline.

hr report has been publishing material on this subject since early March, 1979. The following excerpts from *hr report** are for those wishing more background information:

PROHIBITION OF RADIO TRANSMISSIONS in residential areas is being considered by the Oregon State Senate. Senate Bill 423, sponsored by Senator Ted Hallock of Portland, proposes sharply restricting all electromagnetic emissions in residential areas.

In Testimony Favoring the bill Merrie Buel, government affairs coordinator for the Oregon Environmental Council, said that medical studies "have found that persons living next to electromagnetic sources often experience serious health effects, including rashes, headaches, dizziness and tingling sensations."

Power Transformers and transmission lines as well as radio and TV transmitters would be curtailed under the bill's provisions, though Senator Hallock and members of the Senate Committee on Environment and Energy have been discussing removing transmission lines from its coverage.

As Written the bill would become effective January 1, 1983, after which violations of the standards established for it would be a misdeameanor punishable by a $250 fine. However, Ms. Buel termed the $250 fine "merely a slap on the hand," stating that her group felt that "endangering

table 2. Approximate operating distances between an antenna and humans for 1 mW/cm^2 or less exposure.

	antenna length and minimum separation meters (ft.) with 100 W antenna input for 0.1 mW/cm^2 field or 1000 W antenna input for 1 mW/cm^2 field			
frequency (MHz)	half wave	quarter wave	eighth wave	sixteenth wave
1.8	16.8 (55)	13.7 (45)	11.0 (36)	8.5 (28)
3.5	11.0 (36)	9.1 (30)	7.0 (23)	5.5 (18)
7.0	7.3 (24)	5.8 (19)	4.6 (15)	3.7 (12)
14.0	4.9 (16)	4.0 (13)	3.0 (10)	2.4 (8)
21.0	4.0 (13)	3.0 (10)	2.4 (8)	1.8 (6)
28.0	3.7 (12)	2.7 (9)	2.1 (7)	1.5 (5)

**hr report* is published by Communications Technology, Inc., Greenville, New Hampshire 03048.

table 3. Approximate operating distances between portable/mobile antenna and humans.

Antenna length and minimum separation, cm (in.) with 10 W or 1 W antenna input for 10 mW/cm²* (divide power by 10 for 1 mW/cm² field)

frequency (MHz)	half wave 10W	half wave 1W	quarter wave 10W	quarter wave 1W	eighth wave 10W	eighth wave 1W	sixteenth wave 10W	sixteenth wave 1W
50	55.9 (22)	40.6 (16)	48.3 (19)	33.0 (13)	35.6 (14)	25.4 (10)	27.9 (11)	20.3 (8)
144	27.9 (11)	17.8 (7)	25.4 (10)	15.2 (6)	17.8 (7)	12.7 (5)	15.2 (6)	10.2 (4)
220	22.9 (9)	15.2 (6)	17.8 (7)	12.7 (5)	15.2 (6)	10.2 (4)	10.2 (4)	7.6 (3)
420	15.2 (6)	10.2 (4)	12.7 (5)	7.6 (3)	10.2 (4)	7.6 (3)	7.6 (3)	5.1 (2)

*Maximum exposure at 10 mW/cm² should be limited to 6 minutes/hr.

people's health should be considered a much more serious offense." Furthermore, she said, the OEC wants the bill to become law much sooner since, "we suggest that the sooner electromagnetic radiation is under control, the safer the public health." (*HRR* 245, March 16, 1979).

EFFECTS OF CB ANTENNA RADIATION on the bodies of nearby people is being investigated by the Department of Heath, Education and Welfare. The first study, published in a 24-page booklet titled "Measurement of Electromagnetic Fields in Close Proximity of CB Antennas," discusses bumper, trunk lid and rooftop-mounted mobile antennas as well as those on hand-held units. Near field radiation distribution of each type is presented graphically, in an attempt to determine what hazard, if any, radiation presents.

The Study Concludes: "The health implications (of CB antenna radiation) are not clear at this time. The Bureau of Radiological Health is continuing to investigate this matter." HEW is obviously quite concerned with the effects of RF on the population, and with Amateurs running 200 times the power of CBers on frequencies from 1.8 MHz through millimeter wavelengths, our operations are sure to come under careful scrutiny as well — if they haven't already. (*HRR* 247, March 30, 1979).

AMATEUR RADIO WAS ATTACKED as "one of the main non-ionizing radiation hazards in the United States" at an April 9-10 meeting of the Subcommittee on Public Health Aspects of Energy, in New York. The group is an arm of the New York Academy of Medicine's Committee on Public Health, reports K6YB, who has an article on the effects of Amateur RF radiation on family and neighbors coming out in *ham radio* magazine later this year. (*HRR* 253, May 18, 1979).

RF RADIATION HAZARDS are the subject of a new FCC Notice of Inquiry, General Docket 79-144, agreed to by the commissioners earlier this month. Although the Commission noted that promulgation of RF radiation health and safety standards is the responsibility of health and safety agencies, it also recognized that it would have to consider radiation exposure standards adopted by other Federal agencies in its licensing activities.

Full Text Of This Potentially very important NOI, which is reported to contain a number of questions on specific areas of concern, hasn't yet been released. With the environment currently a hot public issue, this NOI could easily become a crucial one for Amateur Radio as well as most other radio services.

Comment Date for Docket 79-144 is December 15, with Reply Comments due March 15 of next year. (*HRR* 258, June 22, 1979).

ANOTHER FCC PROPOSAL that could affect Amateur Radio is in General Docket 79-163, which proposes changes in the Commission's environmental impact rules. At present those rules offer some leeway with respect to prospective stations that could have a "Major Impact" on the environment when the actual impact would seem less significant. Under the proposed tighter restrictions, it appears formal impact statements would be required of many more applicants, probably including a number of Amateurs, and the Commission would then have to prepare and distribute a written environmental assessment for each such case.

In A Dissenting Statement to the proposed change, Commissioner Washburn makes the point that environmental impact is not the Commission's business and to make it so would add to their already heavy workload and thus increase licensing delays.

Comments On Docket 79-163 are due by August 1. (*HRR* 260, July 6, 1979).

HIGH LEVELS OF RF RADIATION have been detected by the FCC in its test of some popular personal computers. Tests of computers manufactured by Atari, Apple, Commodore, Heath, Southwest Technical, and Radio Shack have reportedly shown that, in most cases rf radiation levels far exceed allowable Class 1 TV limits.

With The Popularity of home computers sharply on the rise, the FCC plans to use the data it's collected to set up new rules governing all computers that could be used in the home. It will probably be several months before the FCC decides what action to take and files a notice of proposed rule-making. (*HRR* 261, page 2, July 13, 1979).

references

1. *Efforts by the Environmental Protection Agency to Protect the Public from Environmental Non-Ionizing Radiation Exposure*, U.S. General Accounting Office, Report CED 78-79, March 29, 1978.
2. Richard A. Tell, "Broadcast Radiation: How Safe is Safe?", *Spectrum*, IEEE, August, 1972, pages 43-51.
3. *Reference Data for Radio Engineers*, Chapter 25, "Antennas," Howard W. Sams & Co., Inc., Fifth Edition, pages 25-1 through 25-3.

ham radio

the Hellschreiber
a rediscovery

European Amateurs are using a teleprinting system made from World-War II surplus — will it replace RTTY?

The Hellschreiber is a teleprinting machine based on a principle entirely different from that of the RTTY teleprinter. The Hell sytem (named after its inventor, Dr. Rudolf Hell) could have been invented with the requirements of the Radio Amateur in mind, but strangely enough the Hell system has never been fully accepted by the Amateur fraternity. The reason may be that an enormous number of used RTTY machines flooded the market at low prices after World War II.

Hell and RTTY existed simultaneously for a long time for both military and commercial use. However, Hellschreibers have now disappeared, mainly as a result of the introduction of protected RTTY systems with automatic-request and error-correcting circuits. Most hams have probably never heard of the Hell system as a means of communications.

the Hellschreiber

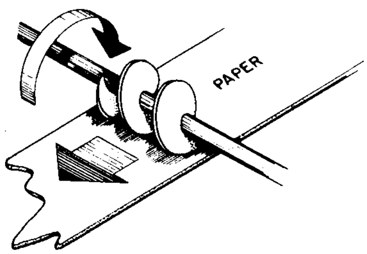
What is the Hellschreiber? In contrast to the RTTY machine, in which received pulses determine the character to be printed, the Hellschreiber uses the transmitted pulses to *directly* write images of characters on paper tape. Thus, Hell writing could be considered a simple form of facsimile, covering seven image lines per character, with seven elements per line.

Not only has this system of printing character images some very important advantages to offer, but the simple way in which the Hell teleprinter works is extraordinarily elegant. The thread of a fast-turning worm shaft wipes, with high speed, transversely across a slowly moving paper tape. This worm thread is wet with printing ink. Every time the paper is

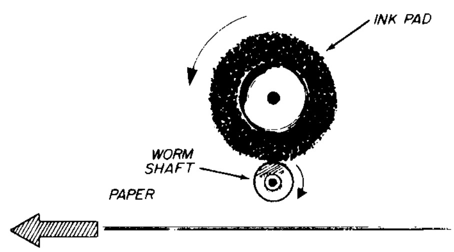
By Hans Evers, PA0CX (DJ0SA), Am Stockberg 15, D-5165 Huertgenwald, West Germany

How The Hellschreiber Works

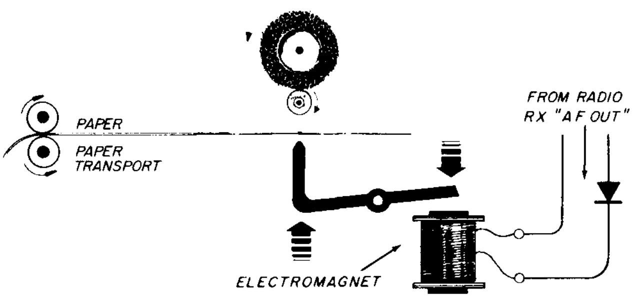
A. Imagine a fast turning worm shaft above a relatively slow-moving paper tape:



B. The thread on this worm shaft is kept wet with printing ink:

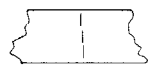


C. Under the paper is a mechanism that taps the paper against the worm shaft by means of an electromagnet:

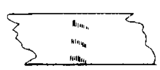


D. What is printed on the paper depends upon the rhythm and the length of time the electromagnet is actuated. For example, if the paper is just tapped, one gets:

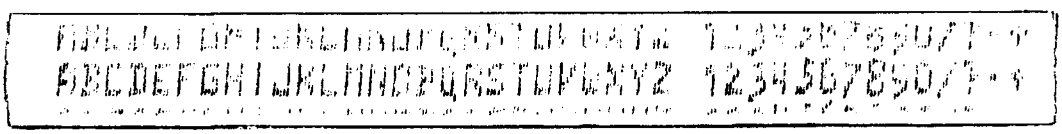
What you see are the little dots where the paper touched the fast-turning worm shaft. If the thread sweeps fast over the paper, and if the electromagnet pushes a bit longer, a little line is printed:



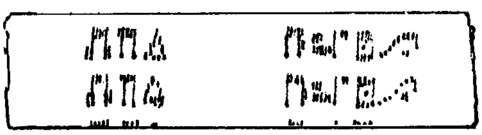
If the tape is tapped in rhythm with the revolutions of the worm shaft, a sequence of little dots is printed:



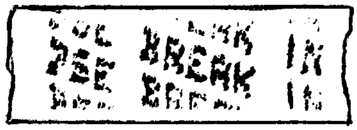
E. Thus, all sorts of simple images can be written; for instance, all the characters of the alphabet:



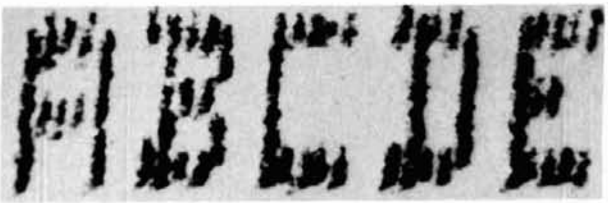
F. Or, if necessary, the characters of anybody else's alphabet, such as Greek, Arabic, or Chinese:



G. What happens if the worm-shaft speed is not quite correct? Nothing serious; the lines of the Hell text threaten to run off the paper tape:



This provides a simple method for determining the correct speed. If, for example, the lines show a tendency to drop, the motor speed must be increased until the lines run straight again. But whatever happens, the text remains legible.



Hell writing. This enlargement shows how each character takes the space of seven image lines. As a result of the relatively slow-moving tape, the characters hang slightly over.

tapped against the turning worm shaft, little lines are formed across the paper tape. Several of these lines together form a character.

The Hellschreiber of the World War II Wehrmacht type we're using runs somewhat slower than the RTTY machine: 2½ characters per second. Nevertheless, a respectable 25 words per minute is achieved. This CW terminology is not misplaced, as Hell and CW have much in common. In fact, given a certain bandwidth, the reliability of Hell communications approaches that of CW.

QRM proof?

During World War II the Hellschreiber proved its reliability. Users recognized that a Hellschreiber could be the only link between an isolated military unit and its headquarters. When all other means of communications failed, often the Hellschreiber managed to get the message through, even when only barbed wire and an earth connection were available as a signal path.

Amateur applications

Our Hell QSOs occur on 80 meters (over here, the official RTTY segment is between 3575-3625 kHz). It's difficult to think of a better part of the radio spectrum for putting the Hell system to the test because of the high QRM level in this portion of the band.

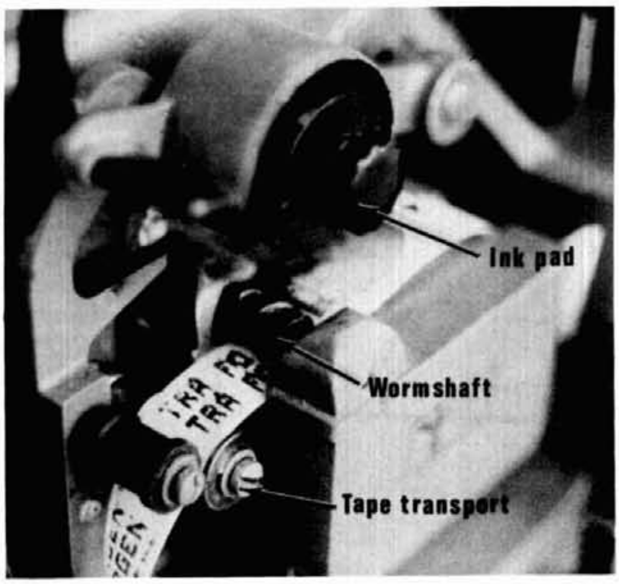
In this context I'd like to mention an interesting side effect. Our modest *prrt, prrt, prrt* Hell signals apparently tend to provoke fury among some hams, who seem to be convinced that the unusual sounds are caused by commercial stations. This turns our little Hell channel into the center of zero-beating and QRZ-blaring stations. This intentional interference does, however, provide us with an invaluable opportunity to test the communications system under highly adverse conditions and is, therefore, to some extent, not unwelcome.

Of course, the interfering transmitter determined to cause serious trouble by tuning carefully zero-beat with our Hell signals may eventually manage to temporarily destroy our communications, provided, of course, that the signal is stronger than ours. By maneuvering with tuning, bandwidth, and threshold level it's possible to get through. We might lose contact for a moment; however, contact is restored through the foggy QRM clouds on our printouts, and we pick up the text as soon as the characters become distinguishable again. This sort of working on the threshold is possible with Hell: The text, even under the worst conditions, is never subject to errors of a substitution-of-characters type. The character may, however, be difficult to read because of mutilation.

Hell *versus* RTTY

Under certain circumstances the communications reliability of Hell can be even better than that of CW. The received Hell signal is printed in its original form. At the moment of reception no decision has to be made such as, "Did I hear correctly?" Thus wrong decisions are avoided. The Hell printer enables the reader to decide later on, at his ease, what was actually sent by the distant station.

Some examples are shown of radio Hell-communications in which the printer obviously has great trouble in keeping the text intelligible because of a high noise level or heavy QRM. The examples contain considerably more information than can be deciphered on first sight. If you really take the trouble to read the text, you immediately realize to what the Hellschreiber owes its superior qualities: it calls in the services of a computer, *i.e.*, our human ability to recognize pictures in a chaos of little specks and lines.



Printing mechanism of the Hellschreiber. The ink pad (felt roll) has been lifted to show the worm shaft. The paper tape is slowly moved by the transport capstan. The electromagnet (not visible) taps the paper tape from below against the fast turning worm shaft.

The Feldfernschreiber. Hellschreiber of the German Wehrmacht (1938) as used on a large scale during World War II. It is with this type of machine that such remarkable results were obtained on the Amateur bands.

The Hell system is less sensitive to interference than RTTY because the Hellschreiber prints the interfering clutter as well as the desired text. This may sound paradoxical, but it becomes understandable if you realize that a teletype printer must translate its received signal into a character before it can decide which key must be pressed. It cannot count upon the services of a "computer." Thus, with RTTY, a single interfering rf spike may result in a wrong decision, turning out a character that has no resemblance whatever to the actual character transmitted. The unprotected teletype character can't warn the reader that it is in error; it can't even indicate that a certain amount of doubt existed during the moment of its selection!

The Hellschreiber, on the other hand, requires no such decisions. The machine just prints, complete with all the received interference. But (and this is the important distinction) although the interference may give the image of the characters an untidy appearance, the Hellschreiber is not capable of changing it. In other words, the Hellschreiber simply leaves to the boss the problem of sorting out the text from the rubbish and doesn't try to disguise the difficult reception conditions.

This is the explanation for the rather amazing fact that you may read Hell text from signals that are only barely audible through an overwhelming amount of QRM; indeed, that it's even possible to decipher Hell signals received *under* the noise level. No wonder we're highly enthusiastic about this fantastic system.

experience with Hell

For three years, almost every week, our little international Hell group (five Dutch, one German, one French, one British) make our regular Hell QSO of an hour or so, using one of the most crowded portions of the 80- and 40-meter bands. Our Hellschreibers are ex-Wehrmacht printers, some of them 40 years old and in fact valuable museum pieces.

As with CW and RTTY, the modest bandwidth requirements of Hell are a great advantage. They are determined by the shortest pulses contained in the signal, being 8.16 ms. This produces a speed of 122.5 baud, requiring a minimum bandwidth of 61 Hz. Even in an overcrowded band it's possible, with a sharp CW filter, to remove most of the QRM or, in case of telephony interference, to keep the bulk of the speech sidebands out of the picture.

Watching a Hellschreiber printer in operation, you can't help being impressed by its imperturbability: While the radio receiver produces the most frightening sort of QRM noises, the machine swallows it all. Quietly, apparently hardly disturbed by it all, it goes on spelling out its characters. Often the QRM is so bad that you need a Hellschreiber to establish that there's still a Hell signal in the air.

Transmitter section of the original Hellschreiber. The coded drum turns one revolution per character. Every time a key is pressed, one turn of the drum produces a series of pulses by the contact with one series of lamellas.

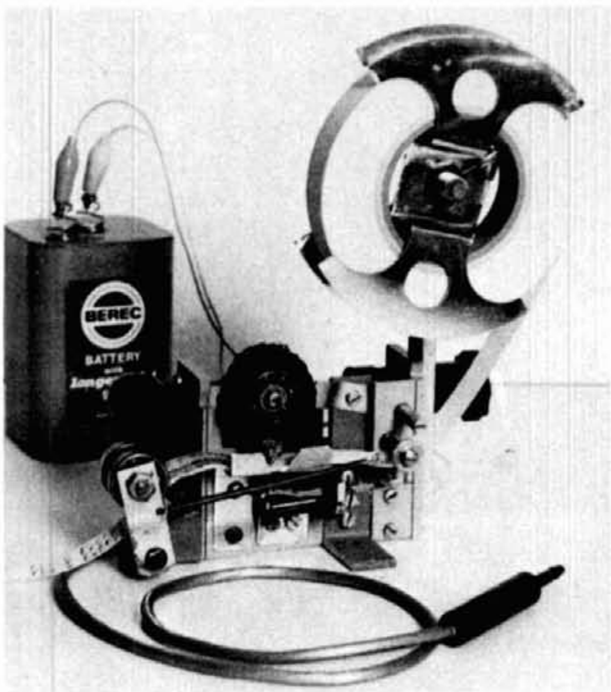
Between transmitter and receiver a certain amount of synchronization is needed, which requires a means of regulating receiver-motor speed. Not that this synchronization is very critical; contrary to what you might expect from a synchronous image-line system, the good old Hell machine is not so easily disturbed by the wrong motor speed. The only thing that might happen is that the written text might drop over the edge of the paper. The text remains legible, however, and, while continuing to read the text, you correct the motor speed by hand until the text prints correctly along the plane of the paper strip. It is this reliable, almost undisturbable, character of the Hellschreiber that makes it such a fine instrument for Amateur Radio communications.

The CW-like disposition of Hell signals permits break-in. Spaces don't produce signals (the tape just runs without printing), so it's possible to cut in between words of the distant station's text. You can even keep watch on the QRM situation between transmitted words.

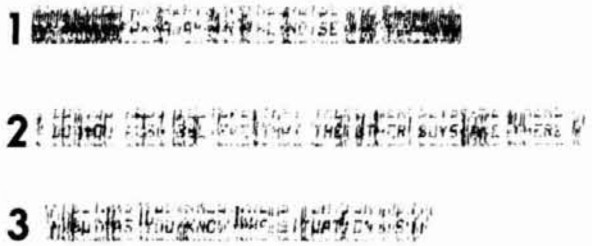
Hell is economical with transmitted energy. With considerable fewer marks than spaces in its signals, and without start and stop pulses, the average output is about 25 per cent of the maximum output. This low duty cycle permits increased transmitting power.

quo vadis?

It's possible to make a Hellschreiber yourself — something that can't be said for any ordinary teleprinter. The actual printer consists of only a simple mechanism. This is another advantage of the Hellschreiber. The receiving part is easy to build and may be a good starting point. After gaining some experience with receiving Hell QSOs, you can decide whether it is worthwhile building a Hell transmitter.



Home made Hellschreiber.



Reception of Hell signals under extreme conditions.

1. Very weak signal, drowning in the noise. On first sight it's unusable; however, our ability to recognize pictures in a chaos of little specks permits us to read the text into the noise.

2. Interference by a strong SSB telephony signal on the same channel. (Text: "Do you also believe that the other boys are there".)

3. Hell signal exactly zero-beat with equally strong 14-wpm CW signal. (Text: "but as you know the situation is".)

We have already built some mechanical Hell printers. Of course, electronics have advanced considerably since 1938, and the dimensions of our modern Hellschreiber can no longer be compared with those of that bulky German design. We now have small electric motors with solid-state speed regulation and we can use refinements such as coils with ferrite cores to pick Hell signals out of overwhelming QRM. Accurately defined Schmitt-triggers are available for separating signals of different levels.

You could even go as far as PA0WV, who has developed a microprocessor displaying received signals as a slowly moving line of characters, complete with interfering pulses (thus fully maintaining all qualities of the Hell system) on an oscilloscope screen.

The transmitter part, "pulse machine," is somewhat more complex to build. In the original Hellschreiber the transmit pulses were produced by a coded drum requiring some mechanical refinements. But a solid-state solution exists here. It was PA0WV again who built the first clock-plus-matrix system that can be hidden under a small keyboard, producing all characters in complete silence.

A converter isn't required for receiving Hell signals. The Hellschreiber can be plugged directly into the headphone jack of any radio receiver (or any telephone line, for that matter). The transmitter output plugs into the KEY jack of any CW transmitter, that's all.

ham radio

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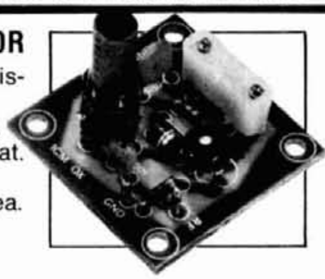

Crystal controlled transistor type. 3 to 20 MHz, OX-Lo, Cat. No. 035100. 20 to 60 MHz, OX-Hi, Cat. No. 035101.

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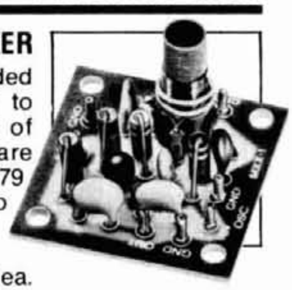

A single tuned circuit intended for signal conversion in the 3 to 170 MHz range. Harmonics of the OX or OF-1 oscillator are used for injection in the 60 to 179 MHz range. 3 to 20 MHz, Lo Kit, Cat. No. 035105. 20 to 170 MHz, Hi Kit, Cat. No. 035106.

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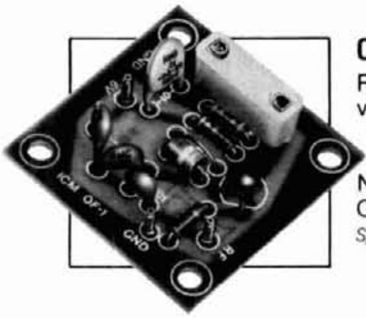

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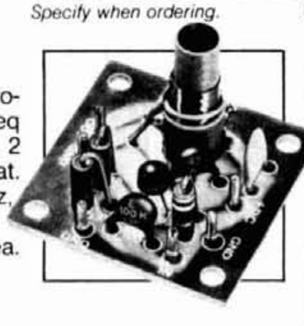

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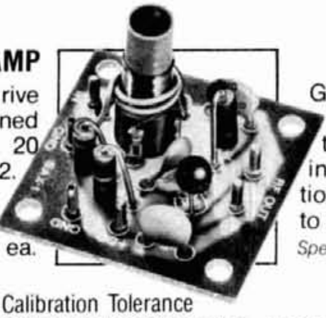

A small signal amplifier to drive the MXX-1 Mixer, Single tuned input and link output. 3 to 20 MHz, Lo Kit, Cat. No. 03512. 20 to 170 MHz, Hi Kit, Cat. No. 035103.

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log-periodic antenna design

· The LP is a useful antenna for Amateur applications — it provides constant gain and a low VSWR over a wide frequency range

The log-periodic (LP) array is a moderate-gain antenna useful for many Amateur applications. It has the desirable characteristics of constant gain and low VSWR over a wide frequency range. It's very forgiving of construction and design tolerances. Accordingly it doesn't require fancy test equipment or interminable pruning to achieve satisfactory operation. Minor errors may result in somewhat reduced gain but won't markedly affect the basic radiation pattern or front-to-back ratio. Once a design has been completed, it's seldom necessary to make adjustments after the antenna has been erected.

This article deals with the design of LP antennas using simple formulas that can be worked on any 4-function calculator. Also given is a simplified approach using only tables and elementary arithmetic, which allows you to design single or multiband LPs to fit into an available space. Examples shown are for wire antennas. For vhf arrays using tubing, appropriate changes should be made to obtain the effective element length. Robert Carrell presented an excellent paper, "The Design of Log-Periodic Dipole Antennas," which is in the IEEE International Convention Record for 1961. Data for the article here was, in a large part, derived from that paper.

The design gains shown here are approximate and may seem low to many readers. They are given as the free-space pattern gain with reference to a dipole (dBd). Many antenna designs are quoted as dB above isotropic (dBi) and sometimes include ground reflection gain over a perfect reflecting surface. Such approaches are misleading and can yield numbers anywhere from 2.2 to 8 dB higher over a dipole.

Amateur applications

The LP is particularly useful in split-band operation such as working DX on 40 or 75 meters, where the frequency separation of U.S. and foreign bands is frequently greater than the bandwidth of many other types of antennas, such as Yagis.

In many areas of the world, material such as telescoping aluminum tubing is difficult and expensive to obtain. The LP, either in a fixed configuration or rotatable in a design using forward V-shaped horizontal-wire elements supported by a shaped stress-line diamond, may be built from simple available materials: wire, bamboo, and nylon line. YV5DLT, Ansel Eckels, has built several of the latter configuration that have worked very well.

In all probability the number of Amateur bands will increase in the not too distant future; new bands have been proposed at 10.1, 18.1, and 25.25 MHz. An LP can be designed to cover 10-30 MHz, with performance and size making it comparable to many current triband beams. At least two Amateur manufacturers (KLM and Telrex) have a practical-size LP rotatable array that comes close to meeting this requirement. Alternatively, a multiband Yagi covering six bands would be quite a mechanical challenge. Unless extreme care is taken in trap design, it would be quite lossy.

For those who operate in the vhf bands, a single LP can yield good performance from 50 MHz through 432 MHz. After many years and many other design approaches, the LP has become the standard configuration for most TV antennas.

Don Bostrom, N6IC, in the DXpedition to Wallis Island in 1974, used a homemade LP with good suc-

By P. A. Scholz, W6PYK, and **George E. Smith, W4AEO.** Mr. Scholz's address is 12731 Jimeno Avenue, Granada Hills, California 91344. Mr. Smith can be reached at 1816 Brevard Place, Camden, South Carolina 29020

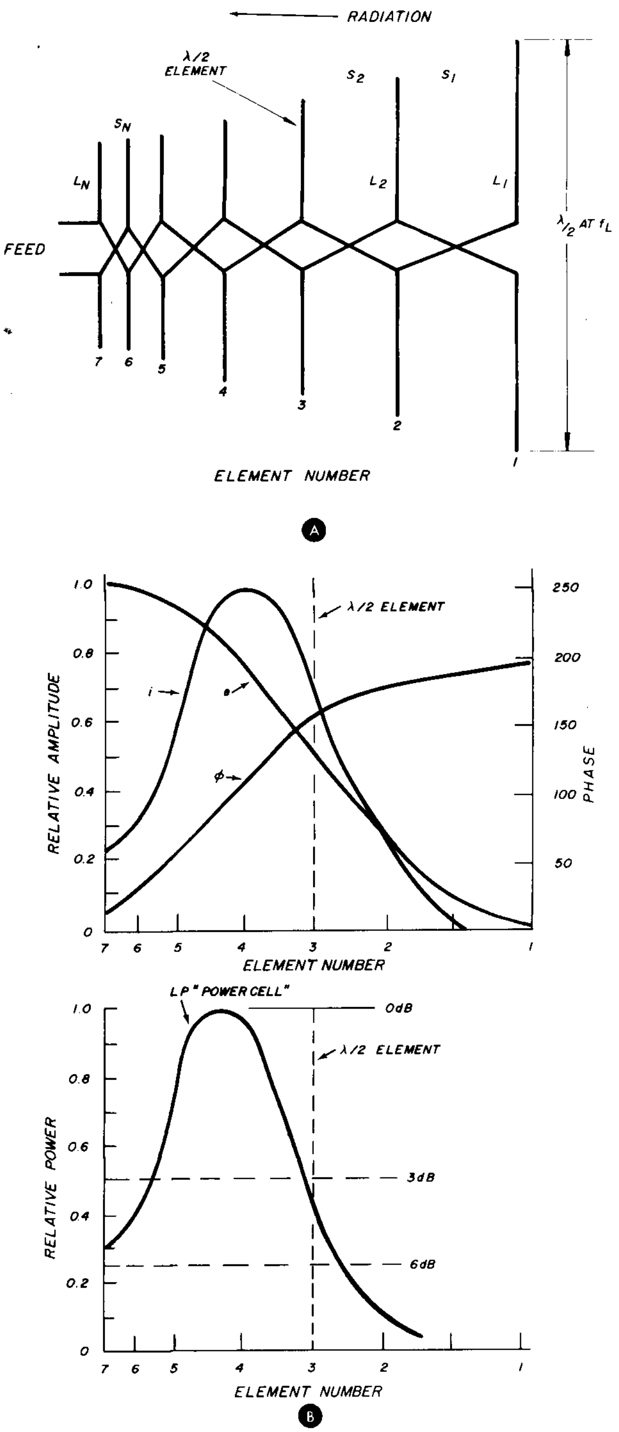


fig. 1. A 7-element LP antenna (A) with equations defining taper factor, τ, in terms of element length, L, and element-spacing relationships, S_1, S_2 . . . S_N. Characteristics are shown in (B). Upper curve: voltage, current, and phase at an arbitrary frequency higher than lowest desired operating frequency, f_L. Lower curve: approximate power distribution, illustrating the "LP cell," which occurs in the region where the elements are less than λ/2 long.

cess. It was a vertically polarized 5-element wire array for 10, 15, and 20 meters.

Compared with the Yagi, the LP will usually have somewhat lower gain. However, it does have a significant advantage in bandwidth and maintains its front-to-back characteristics over the entire design frequency. A properly designed LP working through a balun will allow a solid-state transmitter to operate efficiently over a wide frequency range without an antenna tuner. For example, a 7-element wire beam covering the low end of 80 meters to the high end of 40 meters, and with a good match, can be erected in a space roughly 43 meters square (140 feet square) and will have a gain of about 6 dB.

description

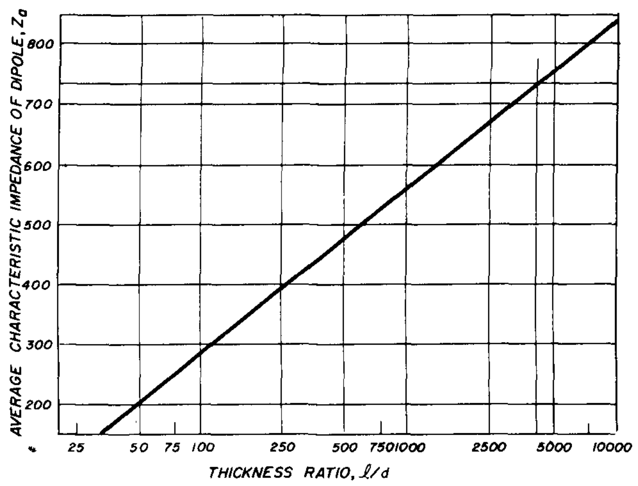
The basic LP consists of a number of dipoles arranged in a plane (**fig. 1A**). The element lengths, L, and the relative spacing, σ, are arranged in a geometric progression with a taper factor, τ. Each element is connected to the feeder in an alternating manner. The feedline is transposed between each set of elements as the easiest method with wire elements and as an intra-element feedline.

The array operates as a backward-wave antenna; radiation is in the direction of the feed. Propagation velocity is about 0.35. The free-space pattern in the plane at right angles to the elements is similar to a cardioid: egg-shaped in the radiation direction in the plane of the elements. The LP operates over a frequency band defined by the longest element, about λ/2 long at the lowest frequency, and the shortest element, about λ/4 at the highest frequency. The gain is constant over this frequency interval; therefore the beam width in E and H planes is constant in free space. Over real ground, the elevation beam maxima will change in angular position with frequency because the effective height in wavelengths will change.

Many configurations have been discussed in *ham radio* and other literature. These include the inverted T monopole, which is a vertically polarized ground-plane array; a configuration using inverted V elements; and truncated LPs, which use fewer than the optimum number of elements.

characteristics

Power to the elements is maximum in the region where the elements are somewhat less than λ/2 long. This is called the "LP cell" (**fig. 1B**). The cell at a particular frequency encompasses the longest element, λ/2, and a few shorter elements. The shortest



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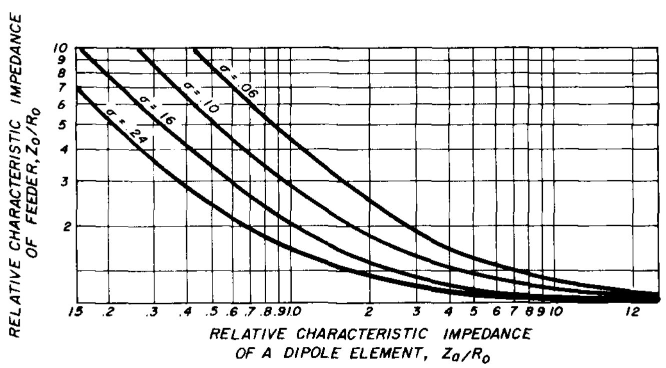

B

fig. 2. Data for the LP intra-array feedline design. The "natural" dipole impedance is selected from curve (A), which shows the average characteristic impedance of a dipole, $Z_a = 120\ (\ell n\ \ell/d - 2.25)$. **(B) shows relative feeder impedance as a function of dipole impedance with spacing factor,** σ, **as a parameter.**

element is approximately $\lambda/4$ long.

The array will operate as an antenna at frequencies lower than that defined by the longest element of $\lambda/2$; however, the front-to-back ratio will degrade rapidly, and the gain will be impaired.

For large taper factors (τ near unity) many elements are within the cell and the array has high gain. For small taper factors only a very few elements will be within a cell, and the gain will be much lower. **Table 1** shows how the gain for an optimum-gain design varies as a function of the frequency range and number of elements.

feed system

The LP at low frequencies is usually fed by coax through a balun. The LP feedpoint impedance, unlike that of other antennas, is a function of the natural dipole impedance and the spacing factor, σ. A convenient feedline impedance can be used with a balun at the antenna to obtain a low VSWR (**fig. 2**). Air dielectric should be used to prevent excessive phase shift within the array feed. Any other dielectric increases the spacing factor, σ, in a complex manner. From tests run by W4AEO, the driving point impedance for low-frequency arrays at $\lambda/4$ high is on the order of 225 ohms, with an intra-array feedline characteristic impedance of 450 ohms. Typically the intra-array feedline is formed from 14 AWG (1.6-mm) wire spaced for a feedline impedance of about 450 ohms. This, with a 4:1 balun, results in a close match to 52-ohm coax.

array gain considerations

For each value of taper factor, τ, there is a corresponding value of spacing factor, σ, which yields maximum gain (**fig. 3**). Smaller-than-optimum spacing factors may be used with consequent loss of gain but without pattern degradation. Larger-than-optimum values of spacing factor cause undesirable lobes. The optimum spacing factor, σ, is approximately 0.19 times the taper factor, τ.

Some commercial rotatable or space-saving arrays have spacing factors as low as 0.03. For example, one may build an array covering 7-30 MHz on a 10-meter (33-foot) boom (sixteen elements, $\tau = 0.895$; $\sigma = 0.03$). It would have about 5.5 dBd gain over this band. With the optimum spacing factor of 0.17, the gain would increase to 7.3 dBd, but the boom length would become 59.5 meters (195 feet). Thus by compromising gain for bandwidth you can build an effective, very broadband array with a practical boom length. This may be more desirable than stacking several potentially interacting Yagi arrays.

mathematical design

Here's a design approach for LP antennas using simple mathematics that can be worked on a 4-function calculator. It is presented to show how the basic design is evolved.

Definition of terms. All of the LP design terms needed for calculator implementation are as follows:

f_H = highest desired operating frequency

f_L = lowest desired operating frequency

$B = \dfrac{f_H}{f_L}$ = desired frequency ratio

B_s = structure bandwidth; ratio of length of longest-to-shortest element

B_{ar} = array bandwidth of active region (amount by which B_s is reduced to obtain usable bandwidth; $B = \left(\dfrac{B_s}{B_{ar}}\right)$

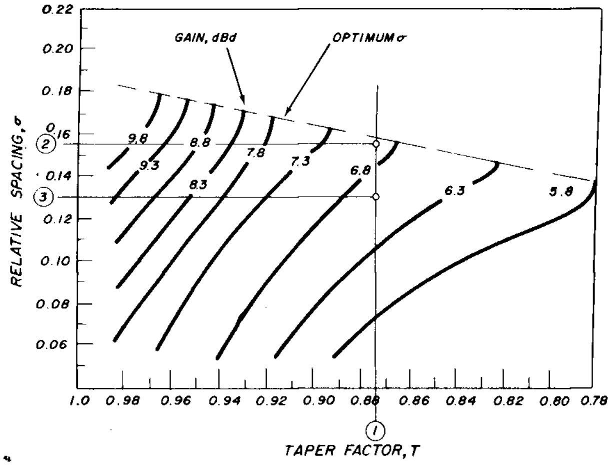

fig. 3. Relative spacing, σ, as a function of taper factor, τ, with contours of constant gain, dBd, as a parameter. Circled values are used in a simplified LP design approach as described in the text.

τ = taper factor

σ = spacing factor

λ_0 = free-space wavelength

$= \frac{300}{f_{MHz}}$ in meters; $\frac{984}{f_{MHz}}$ in feet

λ_L = free-space wavelength at lowest frequency

λ_1 = antenna wavelength using wire

$= \frac{286}{f_{MHz}}$ in meters; $\frac{936}{f_{MHz}}$ in feet

(λ_1 is 5 per cent less than the free-space number because of end effects.)

L = element length

= array length

S_1 = spacing of first two elements

dBd = gain above dipole, approximately 2.2 dB below isotropic gain (dBi)

design steps for log-periodic antenna

1. Select desired band ratio, $B = \frac{f_H}{f_L}$

2. Select an initial set of values for τ, the taper; and σ, the element spacing factor from **fig. 3**.
Spacing of first element $= \sigma \times \lambda_L$
Other elements spaced at $\tau \times$ previous S

3. Calculate array bandwidth, B_{ar}:
$B_{ar} = 1.1 + [30.8\,(1-\tau)\,\sigma]$

4. Calculate structure bandwidth, B_s:

$$B_s = B \times B_{ar}$$

5. Calculate array length:

$$\ell = \left(\frac{\left(1 - \frac{1}{B_s}\right)\left(\frac{4\sigma}{1-\tau}\right)}{4}\right)(\lambda_1)$$

6. Calculate number of elements, N:

$$N = 1 + \frac{\log B_s}{\log \frac{1}{\tau}}$$ and round to next largest number.

7. Calculate new length

$$\ell' = \frac{\text{rounded } N}{N} \times \ell$$

simplified design approach

Design of an LP array using the basic equations above is arduous, since the designer has little feel for the size of the array until he's made one or more iterations. Furthermore, most first tries won't equate to an integral number of elements. Accordingly, a computer program was written to yield initial data on key array characteristics and in terms of antennas with an integral number of elements, **table 1**. This table allows an antenna design based on desired parameter such as number of elements, length, and gain. The following example of antenna design uses the curves of **fig. 3** and **table 1**.

A three-band LP antenna is desired covering 14-29 MHz. Length is about 15 meters (50 feet). Proceed as follows:

1. $B = \frac{f_H}{f_L} = \frac{29}{14} = 2.1$ (desired frequency ratio)

2. Desired length is 15 meters (50 feet)

$$\lambda_L = \frac{300}{14} = 21.4 \text{ meters (70 feet)}$$

$$Length = \frac{15}{21.4} = 0.71\lambda$$

3. From **table 1** choose ($B = 2$), and $\ell/\lambda = 0.87$, or $\ell = 0.87\lambda$. The desired length is slightly shorter to stay on the correct, or lower, side of the optimum spacing factor, σ.

$$N = 10,\ \ell/\lambda = 0.87,\ \tau = 0.875,$$
$$\sigma/\lambda = 0.155,\ gain = 7\,dB$$

4. From **fig. 3** proceed as follows:

a. Modify σ of 0.155 by the ratio

$$\frac{0.71\ (desired)}{0.87\ (\textbf{table 1})} = 0.82 \times 0.155 = 0.13 = \sigma'$$

b. Draw a vertical line through $\tau = 0.875$ (see example in **fig. 3**).

c. Draw a horizontal line through $\sigma = 0.155$.

d. Optimum gain occurs at the intersection (7.0 dB).

e. Modify by drawing a horizontal line through $\sigma' = 0.13$ to intersect vertical through $\tau = 0.875$.

f. A new gain occurs, $\approx 6.7\ dB$ (down only 0.3 dB from optimum).

element parameters

$$f_L = MHz, \quad f_H = 29\,MHz,$$

$$\tau = 0.875, \quad \sigma' = 0.13$$

for a wire antenna, $\lambda/2$

$$L = \frac{143}{f_{MHz}} = 10.2 \text{ meters } (33.4 \text{ feet}) \text{ at } f_L$$

$$S_1 = \sigma' \bullet \lambda_L = 0.13 \times 21.4$$

$$= 2.8 \text{ meters } (9.2 \text{ feet})$$

table 1. Number of elements and array length with optimum taper and spacing.

	N	ℓ/λ	τ	σ/λ	B_s	gain over dipole dBd
1 one-band operation ($B = 1$)*						
	4	0.34	0.79	0.142	2.02	5.9
	5	0.52	0.88	0.155	1.67	7.0
	6	0.73	0.92	0.170	1.52	8.0
	7	0.91	0.927	0.175	1.39	8.9
	8	1.12	0.963	0.178	1.30	9.7
	9	1.33	0.972	0.180	1.26	10.2
	10	1.48	0.978	0.181	1.22	10.6
2 $B = 1.5$						
	6	0.49	0.810	0.142	2.9	6.1
	8	0.77	0.875	0.158	2.56	7.0
	12	1.35	0.93	0.172	2.21	8.2
	15	1.79	0.95	0.175	2.05	9.0
3 $B = 2$						
	7	0.52	0.795	0.142	4.0	5.9
	10	0.87	0.875	0.155	3.39	7.0
	14	1.37	0.917	0.170	3.07	8.0
	19	1.97	0.943	0.175	2.8	8.9
4 $B = 3$						
	9	0.59	0.8	0.142	5.93	5.9
	15	1.18	0.895	0.155	4.80	7.0
	22	1.90	0.932	0.170	4.38	8.0
5 $B = 4$						
	11	0.65	0.815	0.142	7.64	5.9
	17	1.32	0.89	0.160	6.57	7.0
	24	2.03	0.925	0.170	5.97	8.3

*B desired band ratio (harmonic number) $= \frac{f_H}{f_L}$ For an array covering 14, 21, 28 MHz, $B = \frac{f_H}{f_L} = \frac{28}{14} = 2$

$$S_2 = 2.8 \times \tau, \text{ etc.}$$

$$L_1 = 10.2 \text{ meters } (33.4 \text{ feet})$$

$$L_2 = 10.2 \times \tau, \text{ etc.}$$

N	L, in meters (ft)	S, in meters (ft)
1	10.2 (33.4)	2.8 (9.2)
2	8.9 (29.3)	2.4 (8.0)
3	7.8 (25.6)	2.1 (7.0)
4	6.8 (22.4)	1.9 (6.1)
5	6.0 (19.6)	1.6 (5.3)
6	5.2 (17.2)	1.4 (4.7)
7	4.6 (15.0)	1.3 (4.1)
8	4.0 (13.1)	1.1 (3.6)
9	3.5 (11.5)	0.9 (3.2)
10	3.0 (10.0)	
	array length	15.5 meters (51 ft)

conclusion

We have presented design details for a log-periodic antenna using simple mathematical formulas. We have also given a simplified approach to LP antenna design using tables and elementary arithmetic. The LP antenna certainly has a place in Amateur Radio. It has advantages of bandwidth which Yagi and quad antennas don't have. A well-designed LP will provide acceptable forward gain and front-to-back ratio over a wide band of frequencies. The LP can be designed with wire elements for lower frequencies — another advantage when aluminum tubing is hard to obtain. An LP antenna can be designed to cover 10-30 MHz with performance and size comparable to that of many current triband beams.

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appendix

intra-array feedline design

The characteristic impedance of the feedline is a function of the natural dipole impedance, Z_a, and the spacing factor, σ. The natural dipole impedance is related to the length-to-diameter ratio of the elements. The optimum intra-array feedline impedance is determined as follows:

a. Select the desired driving impedance, R_0

b. Determine the mean spacing factor $\sigma'' = \frac{\sigma}{\sqrt{\tau}}$

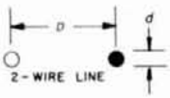
c. Determine the natural dipole impedance:

$Z_a = 120\,(\ln \frac{\ell}{d} - 2.25)$, or from **fig. A1**, where $\frac{\ell}{d}$ is length-to- diameter ratio

d. Calculate $\frac{Z_a}{R_0}$

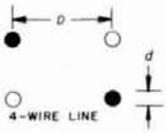
e. Determine $\frac{Z_0}{R_0}$, the intra-array feedline modifier, M.

$$\frac{Z_0}{R_0} = \frac{1}{8\sigma'' = \frac{Z_a}{R_0}} + \sqrt{\frac{1}{(8\sigma''\frac{Z_a}{R_0})^2} + 1};$$



or from **fig. A1**,

$$Z_0 = R_0 \times M$$
$$Z_0 = 276\log_{10}\frac{2D}{d} \text{ for 2-wire line or}$$
$$138\log_{10}\frac{\sqrt{2D}}{d} \text{ for 4-wire line,}$$



where D = center-to-center spacing, and d = wire diameter.

If a non-optimum intra-array feedline impedance is used, the antenna VSWR will vary considerably with frequency, in a somewhat periodic manner.

design example

Array parameters:

$\tau = 0.875 \quad \sigma = 0.13$

$R_0 = 200\ ohms$ (with 4:1 balun from 52 ohms)

Mean spacing factor

$$\sigma'' = \frac{\sigma}{\sqrt{\tau}} = \frac{0.13}{\sqrt{0.875}} = 0.139$$

Natural dipole impedance using no. 14 AWG (1.6-mm) wire:

$\frac{\ell}{d} = \frac{12 \times 21}{0.064} = 3938$; since Z_a varies as $\log_e$,

the mean $\frac{\ell}{d}$ is usually sufficiently accurate

From **fig. A1**, $Z_a = 72\ ohms$, and

$$\frac{Z_a}{R_0} = \frac{720}{200} = 3.6$$

Feedline impedance from **fig. A1**:

$$\frac{Z_a}{R_0} = 3.6\ \sigma'' = 0.139\ M = 1.3$$

$$\text{or } \frac{Z_0}{R_0} = \frac{1}{8\sigma''\frac{Z_a}{R_0}} + \sqrt{\frac{1}{(8\sigma''\frac{Z_a}{R_0}} + 1}$$

$$= \frac{1}{8 \times 0.139 \times 3.6} + \sqrt{\frac{1}{(8 \times 0.139 \times 3.6)^2} + 1} = 1.28$$

$$= 0.245 + 1.03 = 1.28$$

$$Z_0 = M R_0 = 1.28 \times 200 = 255\ ohms$$

ham radio

compact and clean
L-band local oscillators

A clean, L-band local-oscillator system featuring spurious rejection greater than 30 dB with simple test equipment

The recent popularization of microstripline construction for Amateur uhf equipment[1] has made it relatively simple for countless experimenters to build state-of-the-art multipliers, amplifiers, mixers, filters, and the like directly from magazine construction articles. A major exception, unfortunately, has been in the area of microwave local oscillator chains. Most will agree that the LO is the weak link in just about every microwave transmitter, receiver, or converter. Local oscillators generally require extensive tweaking on costly spectrum analyzers, even then often falling short of the required calibration tolerance, stability, and spectral purity. And, since a spurious or drifting LO can negate all the benefits of the very finest low-noise front end or ultra-linear power amplifier, it is evident that the LO requires a great deal of attention. I've been making an effort in recent months to take some of the mystique out of local oscillator design and construction. In this article I shall present the results of that effort — a high-stability, crystal-controlled LO for 1.1 through 1.6 GHz that is only 6.4×10 cm (2.5×4 inches) and employs microstripline construction (with absolutely *no* coils to wind). It can be built for about $70 in parts, and can be completely aligned with only a VOM and a diode detector.

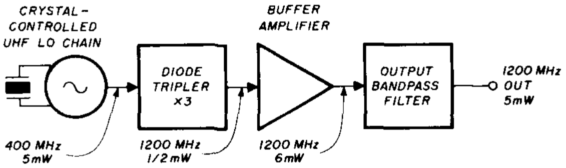


fig. 1. Block diagram of a module LO system for L band. Each block represents a physically separate stage, individually boxed and interconnected by short lengths of coaxial cable.

The basic approach that numerous experimenters (myself included) have used for L-band local-oscillator chains over the past several years is blocked out in **fig. 1**. I implemented this system modularly, with each of the blocks representing a separate, shielded box, the various modules being interconnected via coaxial cable. The basic 400-MHz oscillator is a version of my recently published uhf LO chain,[2] driving a diode tripler built on a microstrip bandpass filter board.[3] In order to make up the power lost in the passive multiplier, a buffer amplifier is employed.[4] The final microstrip filter keeps all spurious frequency components down better than 40 dB with respect to the desired output.

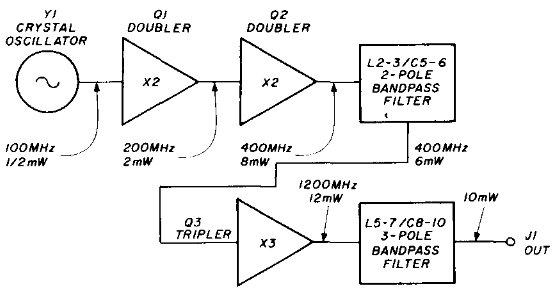


fig. 2. Block diagram of the microstrip local oscillator.

The modular LO chain has some major drawbacks. It is physically large, since each stage must occupy its own separate enclosure to achieve acceptable spurious rejection. The cost of jumper cables, coax connectors, and die-cast aluminum boxes can be

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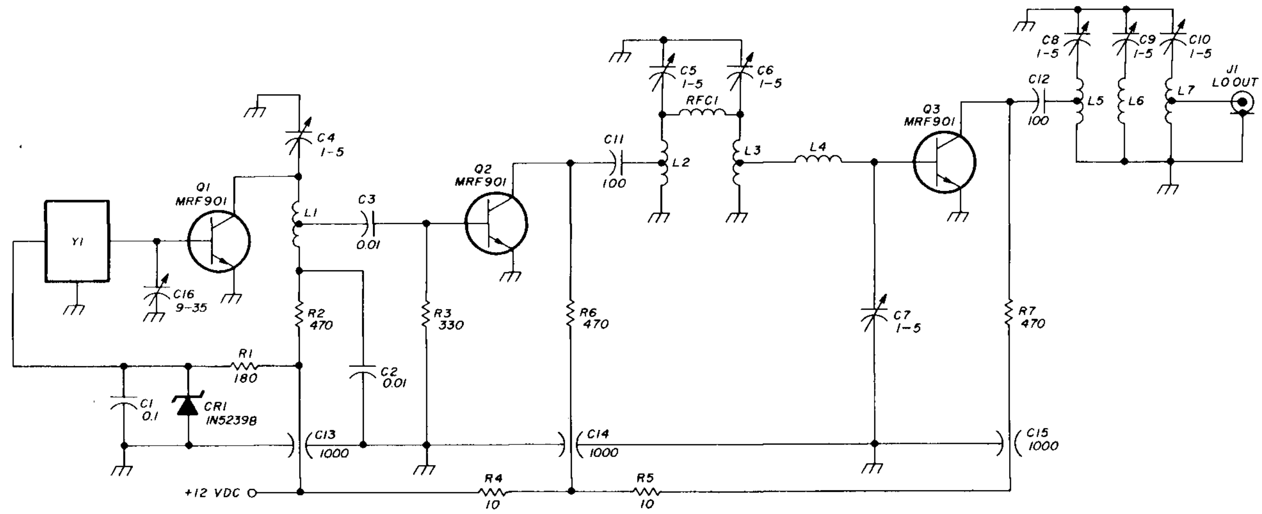

designation	description
C01-03	miniature ceramic disc capacitor, 0.01 μF
C04-10	ceramic piston trimmer capacitor, Triko 202-08M, 1-5 pF
C11-12	chip capacitor, ATC type 100B, 100 pF
C13-15	ceramic feedthrough capacitor, Erie 2404-000-X5UO-102P, 1000 pF
C16	ceramic trimmer capacitor, 9-35 pF
D01	zener, 9.1 V ± 5%, 400 mW. 1N52393
J01	SMA receptacle, E. F. Johnson 142-0298-001
L01-07	microstrip inductor (PC artwork LO-1200-601)
Q01-03	uhf silicon bipolar transistor, Motorola MRF-901
R01	carbon composition resistor, 180 ohm 10% 1/4 watt
R02, 06, 07	carbon composition resistor, 470 ohm 10% 1/4 watt
R03	carbon composition resistor, 330 ohm 10% 1/4 watt
R04-05	carbon composition resistor, 10 ohm 10% 1/4 watt
RFC01	0.33 μH molded choke, Nytronics mini-ductor
Y01	crystal oscillator assembly, International Crystal OE-5, 90 to 130 MHz (output frequency ÷ 12)

fig. 3. Schematic diagram of the L-band local oscillator. C1, C2, and C3 are miniature ceramic disc capacitors. C4 through C10 are Triko 202-08M, 1-5 pF ceramic piston trimmers. ATC type 100B chip capacitors are used for C11 and C12. All resistors are ¼ watt, 10 per cent tolerance. The rf choke is a Nytronics Mini-ductor. The crystal oscillator assembly should be in the frequency range of 90 to 130 MHz, depending upon the desired output frequency (LO output frequency divided by 12).

prohibitive. Plus, the use of a passive multiplier followed by a buffer amplifier is a crude and inefficient way to generate the required 5 to 10 mW of LO output.

I toyed with the idea of integrating the LO chain onto a single board, but became convinced that first it would be necessary to develop a reliable active multiplier circuit to take the place of the diode tripler and buffer amplifier modules. For a time I considered the push-pull tripler approach which Wade had used in his 1296-MHz LO,[5] but in studying the spectrum analyzer photos from his article, I noticed a few potential difficulties. Wade's output filter had the advantage of requiring no tuning whatever, but it afforded only about 20 dB of spurious rejection. His active multiplier, though far easier to tune than my diode triplers, appeared to offer about the same degree of conversion loss. Since I was seeking multiplier gain of not less than unity, I decided to try a single-ended active tripler followed by a tunable, 3-pole microstrip filter to keep the spurs down.

What finally produced acceptable results was an active parametric multiplier, a circuit technique I had employed in an earlier 1296-MHz converter.[6] The secret is to place a series tank circuit, which resonates at the desired output frequency, in the base lead of a standard class-C common-emitter multiplier. This throws the base into a negative-resistance region at the stage's *output* frequency, enhancing the gain of the desired frequency multiple relative to that of the other multiples. The result is an improvement in the multiplier's spurious rejection without resorting to harmonic-cancelling circuits like push-pull and push-push.

With the active multiplier and an output filter tacked onto the end of one of my 400-MHz LO chains, I found I was getting as much output, with as clean a signal as my modular LO had yielded. Plus, it took up only half the space, and without the various coaxial jumpers between stages. **Fig. 2** shows the block diagram of the new LO, and the complete schematic is shown in **fig. 3**.

circuit description

Refer to **fig. 3** and the accompanying parts list. The stages to the left of Q3 are essentially the same

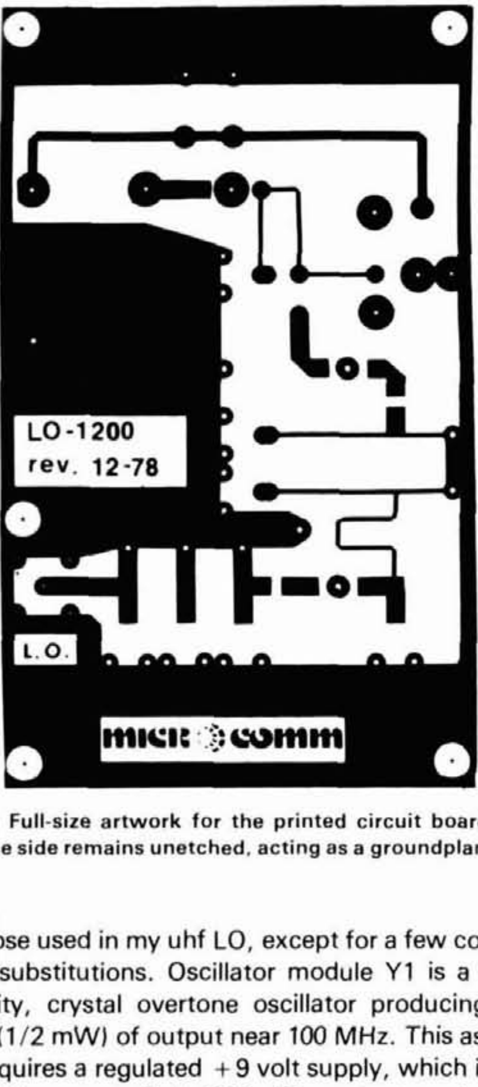

fig. 4. Full-size artwork for the printed circuit board. The reverse side remains unetched, acting as a groundplane.

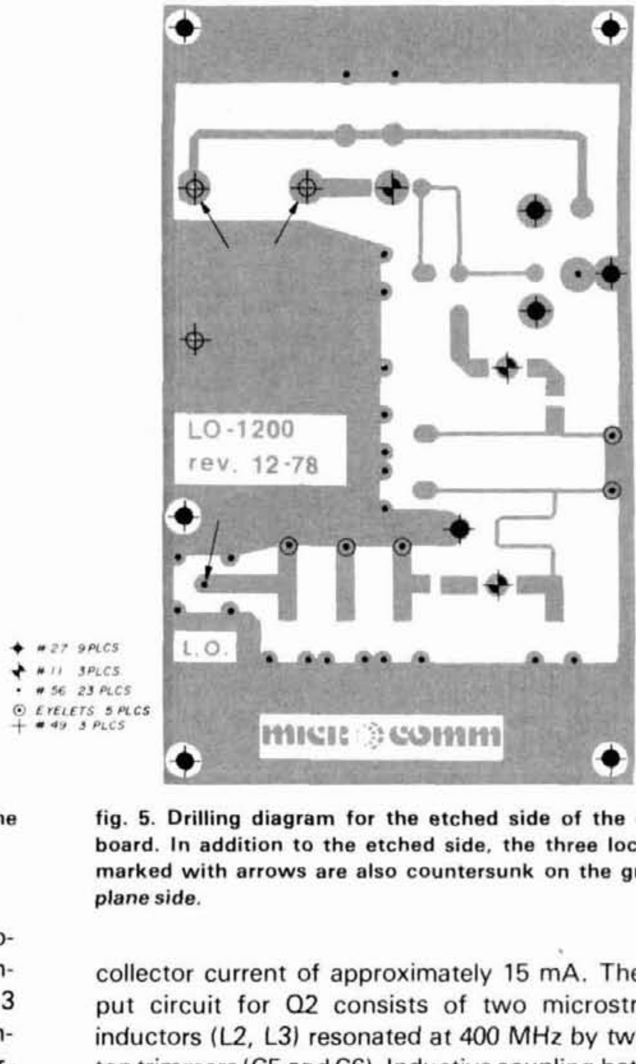

fig. 5. Drilling diagram for the etched side of the circuit board. In addition to the etched side, the three locations marked with arrows are also countersunk on the groundplane side.

as those used in my uhf LO, except for a few component substitutions. Oscillator module Y1 is a high-stability, crystal overtone oscillator producing −3 dBm (1/2 mW) of output near 100 MHz. This assembly requires a regulated +9 volt supply, which is furnished by zener diode CR1. The output port of Y1 exhibits dc continuity to ground, this continuity being essential in providing a bias return for the following stage.

At the input to Q1, common-emitter class-C double stage, C16 is used to resonate Y1's output inductive link, creating a double-tuned, interstage transformer between the oscillator stage and the first multiplier. The output circuit for Q1 consists of microstripline inductor L1 resonated by C4 at 200 MHz.

Collector current for Q1 is limited to 10 mA at R2. This resistor, along with C2 and C13, provide power supply decoupling for the first multiplier stage.

Q2 serves as a second class-C common-emitter doubler. Its input is fed via C3, which is tapped down on L1 for impedance matching. R3 provides base bias return. The collector is shunt-fed by R6, with a collector current of approximately 15 mA. The output circuit for Q2 consists of two microstripline inductors (L2, L3) resonated at 400 MHz by two piston trimmers (C5 and C6). Inductive coupling between filtering poles, provided by RFC1, suppresses higher-order harmonics at the output of the second doubler. The conversion gain of each doubler — exclusive of any filter losses — is on the order of +6 dB.

As mentioned previously, tripler Q3 operates as a parametric multiplier. The input is applied via a low-pass filter consisting of microstripline inductor L4 and piston trimmer C7. The series inductance of C7 is such that it self-resonates at the desired output frequency, maximizing gain at that particular frequency by driving the base impedance of Q3 negative. For this reason, use only the specified capacitor at C7. Shunt collector feed for the active tripler is via R7, with dc decoupling provided by R5 and C15. Collector current for Q3 is on the order of 15 mA, and the stage operates at approximately 3-dB gain.

The output of Q3 is capacitively coupled via C12 into a 3-pole output filter consisting of microstripline

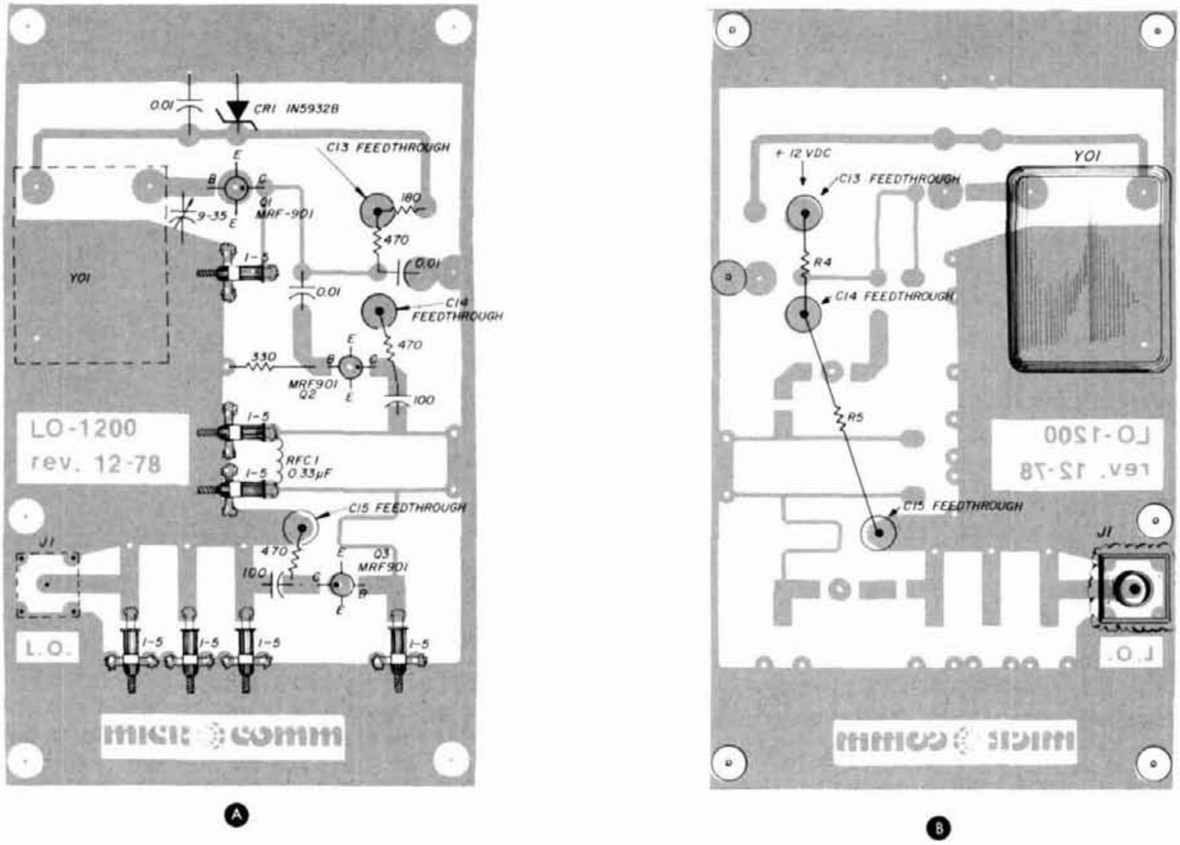

fig. 6. Component placement diagrams for the microstripline and groundplane sides.

inductors L5, L6, and L7, resonated at 1.2 GHz by piston trimmers. Coupling between filter poles is a result of the proximity of the piston trimmer stators; hence, spacing between filter poles is critical.

The final +10 dBm (10-mW) LO output is available on connector J1. This level is suitable for driving most transmit and receive diode balanced mixers.

construction

Assembling the LO is relatively simple, since all circuitry is mounted on a single printed circuit board and the bulk of the critical components are implemented as etched microstriplines. The circuit should be etched on fiberglass-epoxy circuit laminate 1.5 mm (0.063 inch) thick, clad on both sides with 1 ounce per square foot copper. One side of the board is etched in accordance with the artwork supplied in **fig. 4**, the other side remaining fully clad and serving as a groundplane. It is essential that the pattern of the printed circuit artwork be followed exactly (photo etching is recommended), since the dimensions of the microstriplines are critical and the placement of the circuits on the board determines the degree of spectral purity achieved. In fact, the layout of the board was changed several times during the development phases in order to optimize performance and ease of tuning.

After the board is etched, it should be drilled as in **fig. 5**. Be sure to remove a small portion of groundplane metallization from around the holes that will accommodate the center pin J1 and the output and power pins of oscillator Y1. Neglecting this crucial step will result in these pins being shorted to ground, which will obviously have a detrimental effect upon circuit performance! Note that five of the microstripline inductors (L2, 3, 5, 6, and 7) must be grounded through the board. This is best accomplished, as outlined in reference 3, with eyelets 0.5-mm in diameter set with a punch and soldered on both sides of the board.

When mounting components on the printed circuit board, you will find it helpful to refer to the photographs, the schematic diagram in **fig. 3**, the layout drawings shown in **fig. 6**, and perhaps to reference

2. I personally find it easiest to install J1 first, soldering the five pins to their respective pads on the microstripline side of the printed circuit board and then running a smooth bead of solder around the body of the connector, securing it electrically and mechanically to the groundplane side. Next, I install feedthrough capacitors C13, 14, and 15. Here I prepare a small solder preform (1 turn of multi-core solder wrapped around the body of the capacitor just under the flange), position the capacitor in its mounting hole, and apply heat to the flange from above. The solder preform will flow, filling the space between the flange and the groundplane. This technique prevents excess solder from accumulating on the groundplane side of the printed circuit board.

Installing the resistors and capacitors according to the layout diagram is relatively easy. Do *not* install power decoupling resistors R4 and R5 at this time; they will be added during the tune-up sequence. When installing the three transistors, note that the raised dot on the plastic package indicates the collector lead. The base lead emerges from the opposite side of the transistor package, with the two emitter leads appearing at right angles to the collector and base. Bend the two emitter leads of each transistor down sharply before installing the transistors in their holes. That way the emitter leads can protrude through to the groundplane side, where they will be bent over and soldered directly to ground.

When installing oscillator stage Y1, care should be taken to prevent traces on the oscillator's printed circuit board from shorting to the groundplane of the LO main circuit board. I recommend installing a thin insulating washer between the oscillator can and the groundplane.

The only additional advice I might offer in microstripline projects is that it is not unusual for component leads to be laid on and soldered directly to printed circuit board traces or pads, rather than running the component leads through holes in the board. For this reason, it is generally helpful to preform and pretrim the component leads prior to installation.

tune up and test

I cannot overemphasize the importance of employing a systematic, orderly approach in tuning up local-oscillator chains. Tuning for maximum smoke (a favorite Amateur pastime) is a surefire way to make one or more of the multiplier stages oscillate (see **fig. 7**). Further, since the LO chain was designed to provide maximum user flexibility, it may be built for a fairly wide range of frequencies and applications. As a consequence, each of the resonant circuits has a relatively wide tuning range, and it is entirely possible to tune up any one of the multiplier stages on the *wrong* multiple, if maximum apparent output power is the only criterion.

In fact, whether a microwave spectrum analyzer is available or not, it's a good idea to pre-align the various piston trimmers to the appropriate part of the spectrum. This is easier than it may sound. If the intended output frequency is below about 1.3 GHz, adjust all seven of the piston trimmer capacitors to *maximum* capacitance (screws all the way *in*). If the intended output frequency is above about 1.4 GHz, adjust all piston trimmers for *minimum* capacitance (screws almost all the way *out*). And, if the oscillator is intended to operate between about 1.3 and 1.4 GHz, pre-adjust all piston trimmers at about mid-range. Now, as you proceed with the alignment procedure, it should not be necessary to adjust any of these capacitors by more than a couple of turns. Keep this in mind, because if you find yourself adjusting the trimmers more than just a little, you're probably enhancing the *wrong* frequency component.

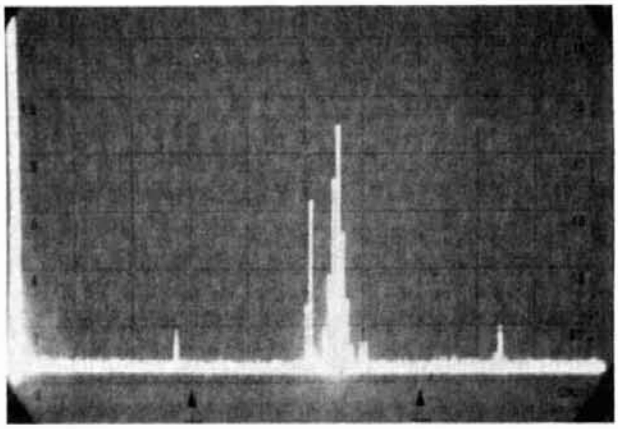

fig. 7. Aligning microwave LO chains for maximum output power often results in a spurious output spectrum. In this case, the LO was tuned for maximum output. The numerous spurs near the desired output are not harmonically related to the crystal frequency. This suggests that they are the result of one of the active stages going into self-oscillation. The horizontal scale is 0 to 2 GHz, with a vertical calibration of 10 dB/cm.

The approach I recommend for tuning this L-band LO requires no test equipment other than a VOM and a diode detector (or some other means of monitoring relative rf power). It is based upon the principle that, as a class-C multiplier stage is tuned, the signal level applied to the next stage (hence the next stage's collector current) will vary. By knowing what kind of variations to expect and by monitoring stage current closely, it is possible to tune the LO chain to produce an output spectrum such as that shown in **fig. 8**. But it is necessary to monitor the various stage currents separately, to be sure you're observing proper multiplier action and not oscillation.

With the piston trimmers pre-adjusted as outlined, the next task in aligning the L-band LO is to get the oscillator stage oscillating. On the side of the crystal oscillator can is a small access hole, behind which is found the ceramic trimmer capacitor which resonates the oscillator's collector tank circuit. This trimmer is pre-adjusted at the factory to ensure that the oscillator will start each time power is applied; it should not be adjusted at this time. Rather, it should be possible to optimize drive to the first multiplier stage merely by adjusting C16, which, you'll recall, resonates the oscillator stage's output coupling link.

Apply a well-regulated voltage between +12 and +13 volts to feedthrough capacitor C13. This powers both the oscillator stage and the first doubler. Monitor the current drawn by these two stages as C16 is adjusted through its range. Since the sum of the current drawn by the oscillator stage and its zener regulator will remain constant at between 16 and 22 mA (depending upon the power supply potential), any variation in current as C16 is adjusted represents the collector current of Q1. There is a point in C16's tuning range where the current at C13 will rise smoothly to about 10 mA above its minimum value (that is, 26 to 32 mA, total), and this is the point to adjust C16. Now, momentarily remove V_{cc} from C13. If the current returns to the previous value, all is well. If on the other hand Q1 appears not to be drawing any current (that is, total current at C13 decreases to between 16 and 22 mA), then the oscillator stage is not starting smoothly and it will be necessary to readjust Y1's trimmer. Do so *carefully*; it should be necessary to rotate the trimmer only about ten degrees one way or the other, and the current should rise again, indicating oscillation. Now, repeak C16 for the proper rise in current, and again remove and reapply power. The adjustments of C16 and the oscillator's trimmer capacitor are somewhat interactive, so repeat the above procedure until the oscillator starts reliably each time power is applied.

Once the adjustment of C16 and the oscillator trimmer is completed, do not under any circumstances change their settings while aligning the balance of the local-oscillator chain. I usually paint a dot of nail polish on C16 to lock it down and tape over the access hole in the side of Y1, lest I be tempted to backtrack and screw things up completely! Remember, the objective is to perform a reasonably clean LO alignment without the use of any costly test equipment, so don't jump sequence.

The easiest way to resonate the collector tank of the first multiplier stage is to monitor the current drawn by second doubler, Q2. Apply operating potential to feedthrough capacitors C13 and C14, and this time monitor the current drawn at C14. This current should peak smoothly at 10 to 12 mA while adjusting C4 no more than two or three turns from its preset position.

Adjusting the interstage circuitry between the second doubler and the parametric tripler is perhaps the trickiest part of aligning this LO because there are three separate trimmer capacitors and the adjust-

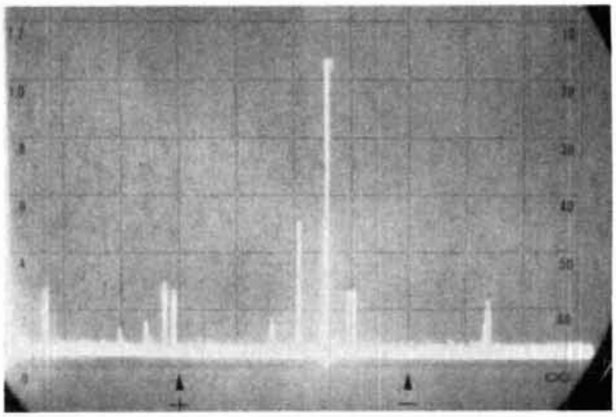

fig. 8. This photograph shows the output spectrum of the L-band LO system after being aligned as described in the article. Spurious rejection of close to 30 dB was achieved with the simple test equipment called for in the article.

ments are all interactive. Note that at this point the only clue you have to proper alignment is the collector current drawn by Q3, so watch it closely. Apply operating potential to all three feedthrough capacitors (C13, 14, and 15), this time monitoring current at C15. First, adjust C7, slightly, just to the point that a few milliamperes of current flow through C15. Now, carefully adjust both C5 and C6 to maximize this current. As before, a peak should occur before the trimmers have been adjusted very far from their preset positions. Once a peak has been found with C5 and C6 both set at approximately the same point, readjust C7 slightly. At this point, C4 (the collector tank of the first multiplier) may be adjusted ever so slightly to maximize current at C15. Now, back to C5 and C6 again, then C7 if necessary, and so on until the current at C15 settles in at about 15 mA. Note that when you're done, both C5 and C6 should have their tuning screws protruding by about the same amount.

All that remains is to align the output bandpass filter. An rf-diode detector can be connected to the output connector, the dc from the diode assembly being fed to a sensitive microammeter as an indication of relative rf output. Any other method of measuring relative rf power (bolometer bridge, calorimeter, or similar) may also be employed. The object is

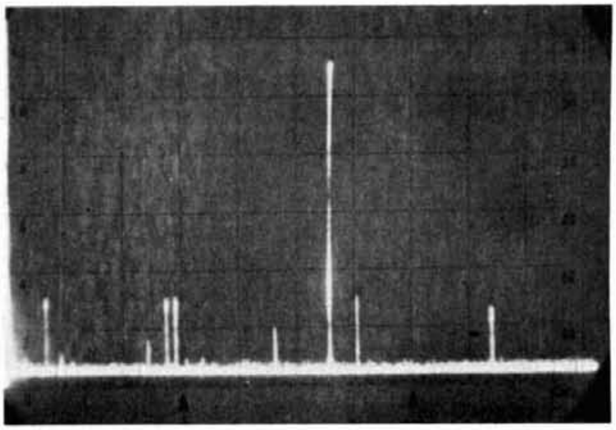

fig. 9. Example of the output spectrum after aligning the local-oscillator chain with a microwave spectrum analyzer. In this case, all spurs are down greater than 40 dB.

to adjust C8, C9, and C10 simultaneously for maximum indicated output (the three trimmer capacitors should all track relatively closely). Keep monitoring the current at C15. If it jumps abruptly, the tripler stage is self-oscillating. It can be tamed down by slightly adjusting C7 until current at C15 returns to its proper value.

With C8, C9, and C10 all peaked at approximately the same setting (not too far, you hope, from the preset position) and output power maximized, one last adjustment to C7 is in order. Adjust this capacitor for the maximum output level obtainable *without causing an increase in the current at C15.* At this point, you may be tempted to go back and repeak all the other trimmer capacitors in the circuit; resist that temptation. You can only disrupt what would in all likelihood be a very clean output spectrum, such as that shown in **fig. 8**.

Of course, if you are fortunate enough to have access to a microwave spectrum analyzer, adjusting the trimmer capacitors ever so slightly can indeed clean up the output spectrum still further (see **fig. 9**). But this should be attempted only after the current-sensitive tuning method has been completed and decoupling resistors R4 and R5 installed.

One final thought. Those super-purists fortunate enough to possess a complete laboratory of microwave test equipment will doubtless notice that any tuning adjustment can potentially affect output power, spectrum, *and* frequency. Thus, you may wish to simultaneously monitor all three parameters during alignment. **Fig. 10** is the lab setup I use in aligning these LO chains. The key to the success of this method is the resistive three-way power divider, which applies equal samples of the LO's output signal to the counter, spectrum analyzers, and power meter. The divider is built simply from four 27-ohm, 1/8-watt, carbon-composition resistors, arranged symmetrically in a small shielded box which supports four coaxial connectors.

parts procurement

Readers of my construction articles frequently write asking if I can supply a complete kit of parts for a given project. Unfortunately, I have neither the time nor the inclination to get into that business. But I am not heedless of the plight of the home constructor, and as much as possible like to help identify (or sometimes create) sources for some of the less-common components.

For example, I have in the past endeavored to make etched, drilled, and plated circuit boards available at cost, for the benefit of those experimenters who prefer not to fabricate their own. This project is no exception. I will supply the boards for $10 each, postpaid anywhere in the U.S. or Canada ($11 elsewhere).

In a previous article, I mentioned a source of supply for the Triko trimmer capacitors I employ in this and other modules. Unfortunately, I later discovered that the importer had a $50 minimum order requirement. Thus, I have recently obtained a quantity shipment of the piston trimmer capacitors used in the LO chain, and will gladly supply them to Amateurs in sets of seven pieces (the quantity needed for each LO chain) at $10.50 per set postpaid in the U.S. or Canada, $11.50 elsewhere.

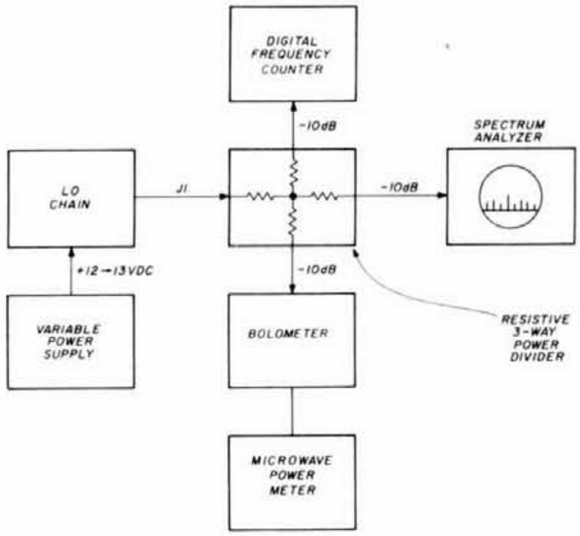

fig. 10. Test configuration for monitoring power, frequency, and output spectrum simultaneously during alignment of the LO chain.

The crystal oscillator assembly designated Y1 is available only from International Crystal Manufacturing Company. I recently spoke to Mr. Royden Free-land, himself a ham and microwave experimenter, and he assures me that this module will be sold to individual experimenters in single quantities. Be sure to allow six to eight weeks for delivery, as the units employ custom-ground crystals.

The MRF-901 transistors used for Q1, Q2, and Q3 are available from Motorola Semiconductor Company. When I first used this particular transistor in a 1296-MHz preamp a few years ago, the price was $9 each. Quantity production and improved yield brought the price down to $4.30 in 1977, and to an unbelievable $1.45 today. At that price, I'd recommend against trying to substitute any other transistor.

The rest of the components used in the LO are, for the most part, garden-variety. Though not necessarily available at your corner Radio Shack, they should nonetheless be obtainable by most experimenters after a bit of scrounging.

Of course, there are always those who need a microwave LO but prefer not to do the scrounging, building, tuning, and testing themselves. To such individuals I am able to offer a completely built, tuned, and tested Model LO-1200 Oscillator Module, packaged in an enameled die-cast aluminum box, operating at your specified frequency between 1150 and 1555 MHz, for $160 postpaid. Foreign orders please add $5 additional postage. This offer is extended to licensed Radio Amateurs only (state your call when ordering), and is restricted to units used for personal, noncommercial applications only. All orders for printed circuit boards, capacitors, or complete LOs must be prepaid in U.S. dollars, and all inquiries must be accompanied by a stamped, self-addressed envelope.

Frankly, I hope nobody takes me up on the above offer. I'd rather design gear than build it for others, and, besides, you're missing out on quite a feeling of accomplishment if you buy your gear ready-made. After all, it is the home constructor to whom this article is dedicated.

Happy building!

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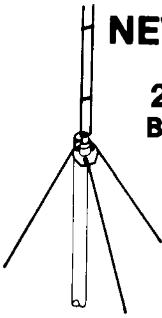

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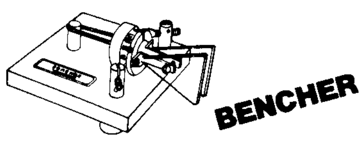

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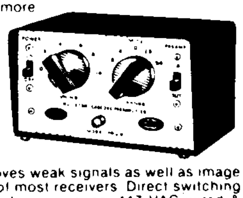

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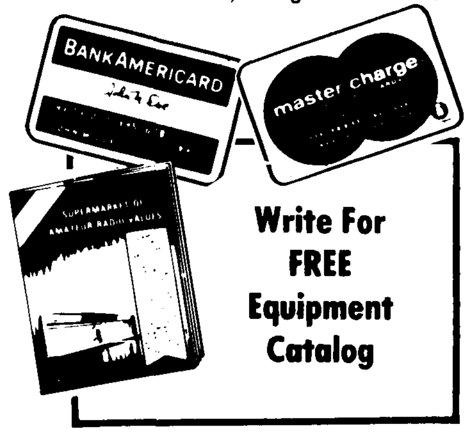

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vhf preamplifiers

A survey of today's devices and techniques for building a truly low-noise preamp for vhf reception

The vhf converters and preamplifiers of ten and fifteen years ago are inadequate by today's standards. This article presents the results of my experience with many low-noise preamps using modern devices and techniques, all of which have been built and tested as a part of a program to update my vhf station. The problems of wide dynamic range and intermodulation distortion are not addressed in this article.

analysis

Consider a typical vhf receiver (**fig. 1**). The receiver consists of X stages, where X is some number. Each stage has a noise figure, NF_X, and therefore a stage noise factor, F_X. Each stage has a gain, G_X, which may be an actual gain or it may be a loss as in the feedline, attenuator/filters, or even in the mixer. Noise figure, NF, is expressed in dB; noise factor, F, is nondimensional. NF and F are related by

$$NF = 10 \log F$$
$$F = \log^{-1}(NF/10) \qquad (1)$$

The most common error in analyzing system noise performance is to lump noise figures and noise factors — a simple but bold dB symbol after each noise figure helps to differentiate the noise figure numbers from noise factors.

It's well known[1-4] that the system noise factor, $F_{S,0}$, as presented by the receiver at the antenna, is influenced to some extent by the noise factors and gains of the following stages:

$$F_{S,0} = F_0 + \frac{(F_1 - 1)}{G_0} + \frac{(F_A - 1)}{(G_0G_1)} + \frac{(F_2 - 1)}{(G_0G_1G_A)} + \frac{(F_B - 1)}{(G_0G_1G_AG_2)} + \frac{(F_C - 1)}{(G_0G_1G_AG_2G_B)} + \frac{(F_3 - 1)}{(G_0G_1G_AG_2G_BG_C)} + \frac{(F_4 - 1)}{(G_0G_1G_AG_2G_BG_CG_3)} \qquad (2)$$

The "gain" of passive elements, such as the feedline and interstage networks (and often the mixer), is actually a loss. That is, a negative gain occurs, which is expressed in dB. When converted into a numerical ratio for use in **eq. 2** this "gain" will be less than unity.

Thus a feedline with a 2.2 dB loss has a ratio of 1.66, which means that about 39.7 per cent of the input power is lost in the feedline. The gain, in this case, is -2.2 *dB*, or $1/1.66 = 0.6$. Noise figure in this portion of the system is essentially equal to the loss, $NF_0 = G_0$, and noise factor $F_0 = \log^{-1}(-G_0/10)$. Therefore $F_0 = 1.66$ and $G_0 = 0.6$.

Similarly, a double-balanced mixer has a negative gain. A double-balanced mixer is a passive circuit to which rf signals and LO inputs are supplied. It uses nonlinear elements (diodes) to derive the i-f signal. However, the double-balanced mixer has a noise figure about 1-3 dB greater than the conversion loss. Thus a typical balanced mixer with a conversion loss of, say, 6 dB (*numerical loss* $= \log^{-1}(6/10) = 3.98$) has a numerical conversion gain $G_C = 1/3.98 = 0.25$ but may have a noise figure of 8 dB, or a noise factor of 6.3. The noise figure is greater than mixer conversion loss.

system noise factor: some simplifications

Several simplifications to **eq. 2** can be made. Each interstage network can be considered as a portion of the rf amplifier output circuit preceding the interstage circuit. The noise factor, F_4, of the subsystem following the first stage can be neglected if the first i-f has a gain, G_3, of at least 100 (20 dB).

The effects of the feedline (G_0 and F_0) can be removed if the first rf amplifier is placed directly at the antenna, and if the first rf amplifier has a gain, G_1, at least 10 dB greater than the losses in the fol-

By Geoffrey H. Krauss, WA2GFP, 16 Riviera Drive, Latham, New York 12110

lowing interstage network and feedline (G_A and G_0', **fig. 2A**).

Thus the receiving system (**fig. 2B**), for noise considerations, can be reduced to:

a) A low-noise preamp at the antenna, which may contain input and output filters as well as the feedline to the second rf amplifier, and which may have an overall stage noise factor, F_1' and stage gain, G_1'.

b) *The second rf amplifier with noise factor and gain* of F_2 and G_2.

c) The frequency conversion stage. For image and noise rejection it includes a bandpass filter as part of an interstage network. It has a noise factor, F_C', and a stage gain (or loss) of G_C'.

d) I-f amplifier, which has an input noise factor, F_3'.

The system noise factor, F_{S0}', is now

$$F_{S0}' = F_1' + \frac{(F_2 - 1)}{G_1'} + \frac{(F_C' - 1)}{G_1' G_2} + \frac{(F_3' - 1)}{G_1' G_2 G_C'} \quad (3)$$

noise factor and noise figure

The system noise factor establishes the overall receiving-system signal sensitivity; a low noise factor (and, therefore, a low noise figure) is desirable above 100 MHz. Occasions arise when low noise figures are also desirable below 100 MHz, as in an extremely quiet location for 50-MHz operation. Accordingly, an evaluation of low-noise amplifier stages for 30, 50, 144, 220, and 432 MHz is useful.

Many receiving systems may use more than one preamplifier. In the systems of **figs. 1** and **2**, first and second preamplifiers are used between antenna and mixer. Regardless of the number of preamplifier stages between antenna and mixer, the general rule is to first adjust the first stage (at the antenna) for minimum noise figure, and then adjust the remaining stages, between the first preamplifier stage and the mixer stage, for maximum power gain. However, as shown by the system noise-factor equations, even the remaining rf amplifier stages should have reasonably low noise figures.

As mentioned, the rule of thumb is that each stage should have at least 10 dB more gain than the noise figure (in dB) of the subsequent portion of the system. However, the noise figure of the subsequent portion is predominantly established by the noise figure in its first stage. Therefore, this stage should have a reasonable noise figure even if adjusted for maximum gain. Thus, if the noise factor presented by the conversion stage is $F_{S2}' = 10$ (or $NF_{S2}' = 10$ dB), the second amplifier stage should have a target gain, G_2, of $(10\ dB + NF_{S2}') = 20\ dB$ and a relatively low noise factor, F_2.

If, after the second rf amplifier stage is built and tested, the noise factor, F_{S1}', of the second rf amplifier/frequency-conversion/i-f system is measured, for example, as $F_{S1}' = 2$ ($NF_{S1}' = 3$ dB) the requirements for the first preamplifier stage are that the gain, G_1', be equal to $10\ dB + NF_{S1}'$ (= 13 dB), with the smallest noise factor, F_1', (and noise figure) achievable at that gain for minimizing the total system noise factor, F_{S0}'. The actual value of the system noise factor/noise figure will depend on the device selected for the preamplifier stage.

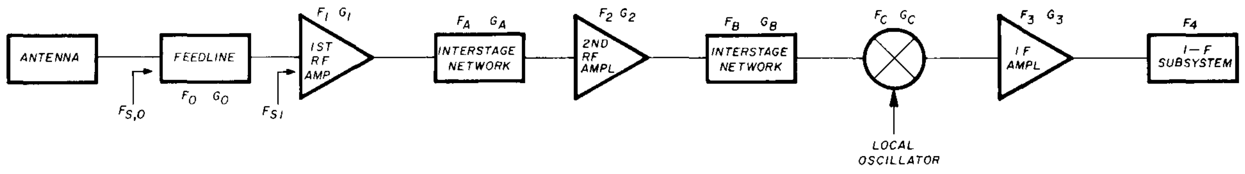


fig. 1. Typical vhf receiver used in the analysis.

background information on devices

What devices are available for low-noise vhf preamplifier use, and what performance can be achieved? To give a proper perspective, consider first a short history of low-noise vhf preamplification.

About twenty years ago, in 1958, the predominantly used vhf bands were 50 and 144 MHz. Little was done, except by a few adventurous Amateurs, with the higher frequency bands. The "low" noise figures then achievable were about 4 dB on 50 MHz, typically with cascade-connected triode vacuum tubes such as the twin triodes of the 6BZ7 variety. About 5 dB was obtained with single triode vacuum tubes such as the 6AM4 on 144 MHz.

Serious experimenters tried to obtain type 5842/417A triodes, which could be used to achieve 3-dB noise figures up to about 250 MHz and about 5-dB noise figures at 432 MHz. A better low-noise tube for 432 MHz was the gold-plated 416B, which was occasionally available as pullouts from microwave relay transmitters. The problems of using pullouts, particularly at the relatively high current levels required for low noise figure operation and the high cost and low availability of special-purpose tubes, deterred widespread Amateur use. Many 432-MHz and most 1296-MHz mixers were of the diode-mixer type, with no preamplification used in many 432-MHz and almost all 1296-MHz Amateur receivers.

Between about 1958 and 1963, vacuum-tube technology progressed to introduce the Nuvistor,® a miniature ceramic-metal tube best typified by the 6CW4, which had the high transconductance and low interelectrode capacitance necessary for high-gain, low-noise preamplification. Noise figures of 3 dB at 50, 144, and 220 MHz were possible, by 1963, in a host of commercially available vhf converters, transverters, and at least one complete receiver (the Clegg Interceptor, for 6 and 2 meters). Versions of the Nuvistor,® would even yield 3-5 dB noise figures at 432 MHz, and a few commercial converters and preamplifiers were offered.

With the advent of relatively inexpensive, relatively low-noise preamps, interest in 220 MHz increased, although not as rapidly as in 432-MHz operation. Interest in 432 MHz was greater because the output of a 2-meter transmitter could be tripled for operation on 432 MHz, whereas a completely new transmitter had to be built for 220 MHz.

Parametric amplifiers were noted for very low noise operation but weren't very popular because of the need for a pump oscillator above 1 GHz and for special components.

Solid-state devices. Early silicon transistors did not, in 1963, appear to have any advantage over the new vacuum tubes. Consequently they saw limited use. However, in another five years, by 1968, the pressure of the commercial home-entertainment market forced development of relatively inexpensive and relatively low-noise silicon transistors. Several transistor types became available at sufficiently low cost for Amateur use. The 2N3819 family was often used for i-f amplifiers. The 2N3823 was specified for rf amplification to more than 150 MHz. Devices such as the Motorola MPF-102 family, the 2N2857, and the 2N4416 appeared. Germanium transistors, notably the 2N1742 and the Philco T2028, were available with even better noise figures than most silicon devices. Transistorized preamplifiers and converters were built with noise figures of 2 dB on 10 and 6 meters, 3 dB on 2 meters and 220 MHz, and 4-5 dB on 432 MHz. The use of vacuum tubes in Amateur vhf receiving gear diminished and all but ceased.

By 1973 a host of solid-state devices were available for vhf receiver use. Silicon devices, with better performance vs temperature characteristics than germanium devices, had, in this period, also achieved better noise figures than their germanium counterparts.

Best remembered of that group were the devices available from KMC Transistors and from Fairchild Microwave Transistors. The KMC devices (now available under different device numbers from Microwave Associates) were capable of less than 2 dB noise figure at 144 and 220 MHz. With selected devices a 2-dB noise figure could be achieved at 432 MHz. If the preamplifier configuration and matching network were chosen and built with care, some of the FMT devices, such as the FMT 4575, reference 5, were capable of even better performance, but cost was generally prohibitive. The availability of the best FMT devices apparently ceased, as later batches of these devices were rumored to have poor noise figures, compared with the earlier batches, and a "lost recipe" was widely rumored as the cause. Solid-state low-noise vhf preamplifiers, were, however, firmly entrenched by this time.

Devices available today. As this is written, some twenty years after the 6BZ7 and 417A era, noise figures of 1 dB are easily obtained on the vhf Amateur bands. I built and tested forty-four preamplifiers to ascertain the performance that might be expected with a wide range of devices. Not all devices available to Amateurs were tested. Financial considerations limited this program to building units only with devices donated by various sources (including many engineers who wished to have a device evaluated and who arranged for these transistors to be made available to me). Results are shown in **table 1** by band and in order of increasing noise figure. **Table 1** also lists

a) The minimum noise figure, M:

$$M = 10 \log \left(1 + \frac{F-1}{1-\frac{1}{G}}\right) \qquad \textbf{(4)}$$

which is indicative of the minimum noise figure obtained with a cascade connection of several such stages.

b) The return loss, R, through the preamplifier from output to input, which is a measure of preamplifier stability and should be at least 8 dB greater than the forward gain, G, (reference 5).

c) A performance ratio factor, P, indicative of the noise figure vs preamplifier cost, which is changing between units of the same noise figure because of device price.

All preamplifiers in this article require a low-noise active device and a pair of matching networks. The input matching network matches the device input impedance to the antenna impedance (50 ohms). Device input impedance may be different, for mini-

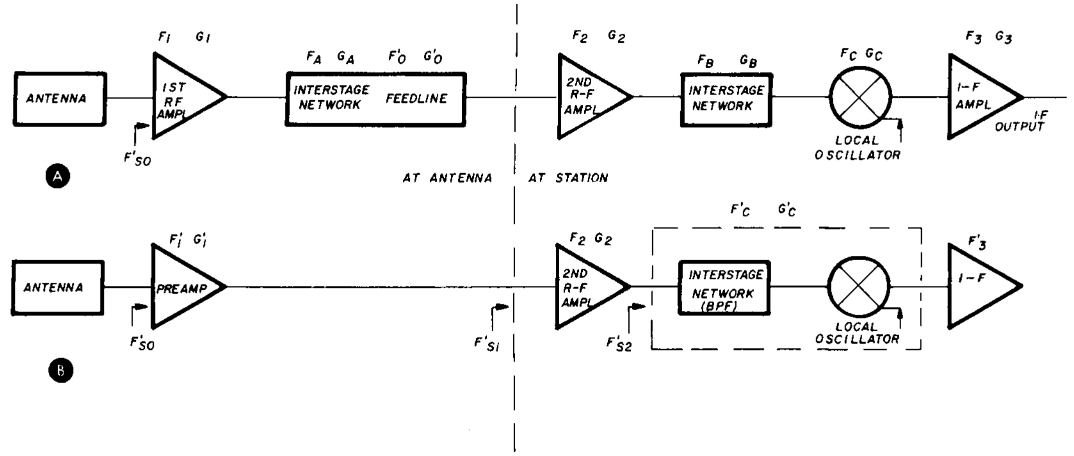


fig. 2. System block diagrams to illustrate how simplifications can be made to the system noise factor, $F_{S,O}$ (eq. 2). The system is divided into two parts: antenna and station. Sketches A and B are used in the text to explain the derivation of the simplified system noise factor, $F'_{S,O}$ (eq. 3).

mum noise-figure operation, from the data sheet values typically given for maximum gain operation. The output matching network matches the device output impedance to 50 ohms.

device choice

Choice of a suitable device should, in general, be governed by the manufacturer's data sheet. It not only lists expected noise figure and associated gain (at that noise figure) at some optimum bias level (generally stated in terms of V_{ce} and I_c for a bipolar transistor, or the equivalent V_{ds} and I_d for an FET device), but also gives

a) Rf parameters for determining stability and matching-network design.

b) A frequency range in which best operation of a device may be expected.

These data are advisory — they are typical values, which may be determined by testing a group of devices. They may not always be obtained for all devices of that type but are good starting points.

stability

If the selected device will operate in the desired frequency range, its stability is always the most important consideration. Nothing is achieved if a low-noise preamplifier is unstable with the input or output impedances found in the system. Instability may result in oscillation and may produce birdies, blocking, or other undesirable results.

Tests for stability are well covered elsewhere. However, two concepts should be kept in mind: First, the device should be stable over a wide frequency range, including the frequency of interest, and second, even potentially unstable devices can be used if care is taken in matching-network design. In this respect, the use of a collector resistance of relatively low value is often helpful and may be found in many of the low-noise preamplifiers discussed in later portions of this article.

bias voltage

The bias voltage and current figures on a device data sheet should always be considered as *nominal* values and should always be varied to achieve best noise figure, assuming that a noise figure test setup is available.

I've tuned several identical preamplifiers containing devices taken from the same shipment. I found

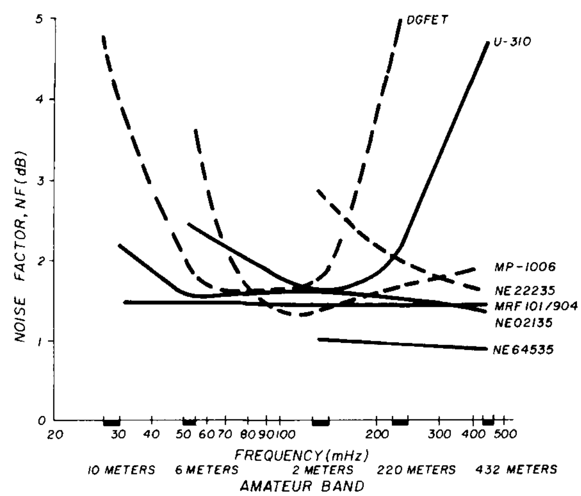
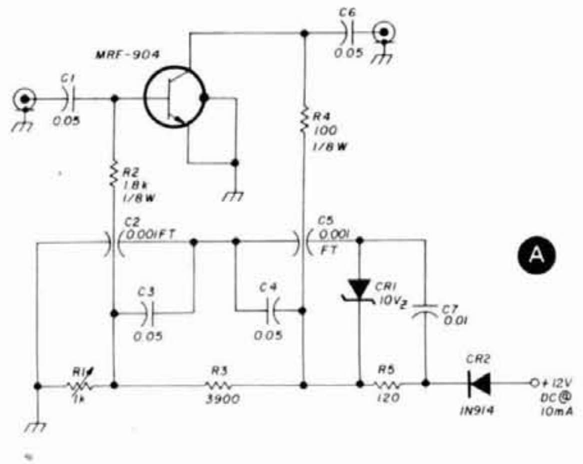
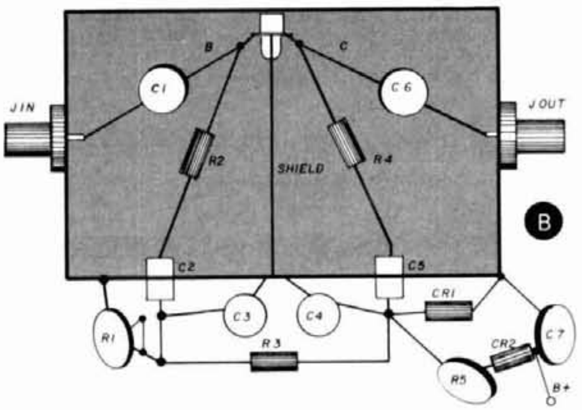


fig. 3. Noise figure as a function of frequency with several devices as a parameter. Note the apparent "useful frequency band" effect, particularly with the dgfet and the U310. Note also the lower frequency limitations shown by devices such as NE22235 and MP-1006.





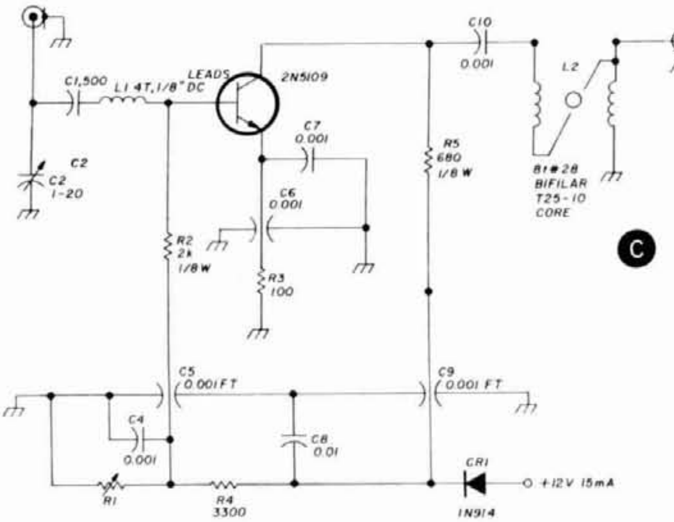


fig. 4. Preamplifiers for 30 and 50 MHz. (A) shows a circuit using the MRF-904, which has very wide bandwidth. (B) is a pictorial layout of the MRF-904 unit. In (C) a 2N5109 is used, which has a 1.7-dB noise factor. Many other devices are usable at this frequency (see text).

that, while almost identical noise figures and gains were achieved, each device had to be biased differently. For example I built and tested four units using NE64535s at 432 MHz; each unit was initially adjusted for data-sheet bias values of 8 volts V_{ce} and 7 mA I_c. All units were then adjusted for minimum noise figure in a test setup using a precision automatic noise figure indicator and noise source:

unit	ultimate NF (dB) (Y method, +0.3dB; −0.1dB)	V_{ce}	I_c
1	0.86	8.6	7.5
2	0.84	7.9	6.4
3	0.91	8.1	10.0
4	0.90	4.9	5.0

S-parameters

Most device data sheets give rf characteristics in terms of S-parameters for gain amplification. While S-parameter design may be new to many Amateurs, several very good articles have appeared.[9-13] The design procedures are relatively easy for devices that are unconditionally stable; *i.e.*, having a stability factor, K, greater than unity at the frequency of interest. However, if the device is only conditionally stable (K less than 1) and the input and output S-parameters (S_{11} and S_{22}) are less than 1, a low-inductance resistor (between 50-200 ohms) may often be used as a collector load to provide a stable preamplifier configuration.[14]

Many data sheets don't give noise-related information. Most particularly, the optimum source reflection coefficient, Γ_{os}, is often missing. This is the parameter that should govern the entire design of a low-noise preamplifier.[15-16] Given Γ_{os}, the input matching network is designed to transform the noise matched input impedance, derived from Γ_{os}, to the antenna impedance (generally 50 ohms). The output matching circuit may be designed once the device output impedance, for the device input connected to the optimum noise-match impedance, is known. Thus, both S and optimum-noise parameters should be known.

device frequency range

The range of device operation should be carefully determined from the data sheet. Many devices appear to be designed for a specific application and, in meeting those application requirements, have parameters that limit the useful frequency range. As an example, the dual-gate FET (DGFET) of **table 1** is an experimental device designed for a noise figure of less than 2 dB over the frequency range of the commercial fm broadcast band. A low noise figure is obtained near this band at 50 and 144 MHz, while the noise figure at 30 and 220 MHz is very high. These latter frequencies were of no interest to the device designer, so the device characteristics are relatively uncontrolled at these frequencies. Note also the noise figure (**fig. 3**) for the NE22235 and NE64535 devices, which are intended for use at 1-4 GHz and 0.5-4 GHz respectively. The noise figure curves show that these two devices were designed for optimum noise figure at frequencies above about 500 MHz. On the other hand, from the noise figure curves of **fig. 3**, devices such as the MRF901/904 were apparently designed for broadband application use from dc through the vhf range.

Device data sheets frequently give a circuit in which the device is tested, and which often makes a good starting point for design of an Amateur circuit. As the frequency increases, care must be taken in circuit layout. Short lead lengths and high-quality components (of types intended for vhf, uhf, or microwave use) are required.

band-by-band discussion

30-MHz. This band is primarily of interest for either OSCAR downlink or as an intermediate frequency for microwave equipment, such as the Microwave Associates Gunnplexers.® In i-f preamplifier application, the preamplifier design following the microwave mixer (generally a diode having 6-10 dB of conversion loss and noise figure) will be an important factor in establishing the overall receiver noise figure.

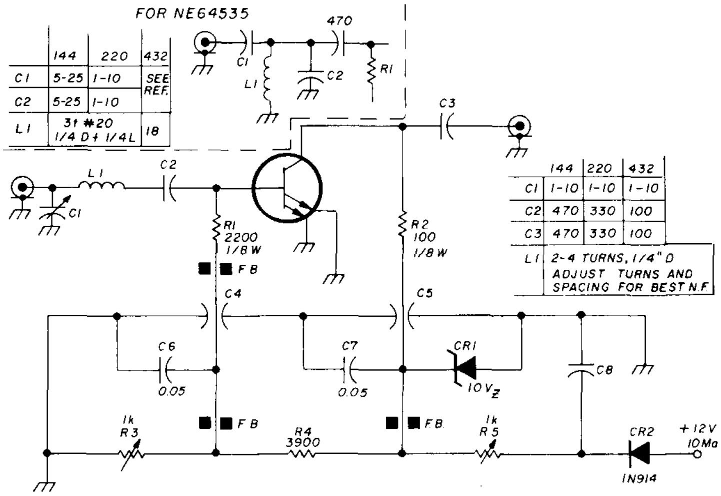
The MA42001-509, as used in W1HR's two-stage i-f amplifier,[17] certainly has the lowest noise figure of units thus far built and tested. However the cost (about $16.50 at the time of writing) of the two devices required for each such preamplifier must be balanced against the small loss in system sensitivity if you use a single-stage i-f preamplifier with a less-expensive 2N5109 or MRF901 device. Use of older devices, such as the 2N4416 and MPF102, is not advised; preamplifiers using these devices have been found to be only conditionally stable, even with heavily loaded output circuits.

The MRF-904 is of particular interest. Apparently it was designed from minimum noise figure when inserted into a 50-ohm system. As shown in **fig. 4A**, matching networks aren't required on either input or output of this transistor in a 50-ohm system. Coupling capacitors C1 and C6 provide dc isolation between input and output feedlines and the device bias circuit. This preamplifier has a very wide bandwidth, as no input or output filtering is used. Output filtering is usually provided by the receiver or converter. The need for, and degree of, input filtering is established by a particular use and location.

For the OSCAR down-link application almost any of the devices yielding less than a 2-dB noise figure should be acceptable, especially in front of a 10-meter receiver having a noise figure of 6-10 dB. I believe the antenna pattern and pointing accuracy are of greater concern than noise figure in this application.

50 MHz. The background noise at 50 MHz is high enough so that a system noise figure of less than about 2 dB is unnecessary. Most of the devices tested achieved this noise figure for a single-stage preamplifier. The noise contribution of succeeding stages (existing converter and the like) will dictate the preamplifier gain and also indicated that a first-*stage* noise figure somewhat lower than 2 dB is required if the 2-dB *system* noise figure is to be achieved.

For example, I presently use a 50-MHz converter with a single 6CW4 rf amplifier preceding a mixer (a remnant of earlier 1960s equipment not yet replaced).



	144	220	432
C1	5-25	1-10	SEE REF.
C2	5-25	1-10	
L1	3t #20 1/4 D + 1/4 L		18

	144	220	432
C1	1-10	1-10	1-10
C2	470	330	100
C3	470	330	100
L1	2-4 TURNS, 1/4" D ADJUST TURNS AND SPACING FOR BEST N.F.		

fig. 5. Preamp for 144 and 220 MHz using the MRF-901, NE02135 or NE22235. Design is similar to that by WB5LUA (reference 18).

The converters have input noise figures between 3.5 and 4.5 dB. If the highest expected converter noise figure (4.5 dB) is taken for worst-case analysis, and if a single-stage rf preamplifier with at least 14.5 dB (equal to 10 dB plus the maximum noise figure of the following stage, 4.5 dB) is to be used, then, by the noise-figure equation, $F_S = F_1 + (F_2 - 1)/G_1$, where the value of $F_2 = \log^{-1}(4.5/10)$, with $G_1 = \log^{-1}(14.5/10)$ and $F_S = \log^{-1}(2.0/10)$. Thus $F_S = 1.585$; $F_2 = 2.818$, and $G_1 = 28.184$. These values are substituted into the equation and give

$$1.585 = F_1 + (2.818 - 1)/28.184 = F_1 + 0.065$$

Therefore, $F_1 = 1.585 - 0.065 = 1.52$, and the preamplifier noise figure, NF_1, should be no greater than $10 \log(1.520) = 1.82$ dB.

Devices tested at 50 MHz that are usable for this particular applicatiion include MA42001 (0.95 dB NF, K = 18.92); NE02135 (1.55 dB NF, K = 19.08); MRF-901 (1.67 dB NF, K = 16.87); KD6003 (1.7 dB NF, K = 22.88); and 2N5109 (1.7 dB NF, K = 16.62). An example of a preamp using the 2N5109 is shown in **fig. 4B**.

Other factors, dependent on your needs, may now be considered, such as circuit simplicity, resistance to overload, and preamplifier bandpass characteristics.

Note that many of the low-noise preamplifier circuits are broadband; a high-*Q* input circuit will add undesired noise before the desired signals can be amplified by the device. Of course, if use is intended in an environment near other rf sources, then input filtering may be mandatory to prevent receiver overload. The amount of input bandpass filtering is determined on a case-by-case basis, although use of low-intermodulation-producing, and therefore overload-tolerant devices, such as the 2N5109 provides some leeway in achieving the desired result. I prefer to place a separate low-insertion-loss filter such as a helical resonator or interdigital filter, having steep skirts and high out-of-band loss, before the first preamplifier. Such filters have a varying impedance outside the passband of interest, so this reactance may cause a potentially unstable preamplifier to oscillate.

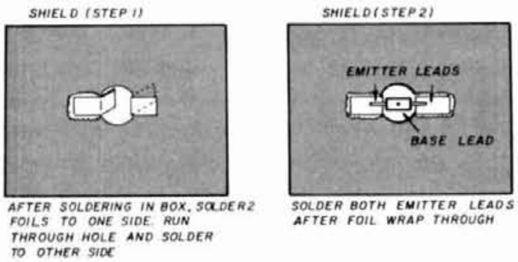


fig. 6. Details of preamp shield construction.

If a 50-MHz i-f system is used with higher frequency (microwave) equipment, the i-f amplifier noise figure may, depending on its total noise figure contribution, require a very low noise configuration using a device such as the MA42001. However, use of such a low-noise-figure preamplifier would usually be expected only when no additional rf amplification is inserted between the antenna and the higher-frequency mixer.

Devices are available for low-noise preamplifiers for the bands up to at least 3300 MHz. Image-rejection considerations at even higher bands, above 5650 MHz, dictate the use of i-fs higher than 50 MHz, so little is gained by using low-noise 50-MHz i-f amplifiers following a microwave mixer. Exceptions always occur — one is a 50-MHz i-f with the mixer portion of a 10-GHz Gunnplexer® transceiver in which a low-noise-figure device, such as the MA42001, would be advisable. Additional information for 50-MHz preamplifiers is shown in **figs. 4** through **7**.

144 and 220 MHz. The 2-meter band has a sufficiently low background noise to allow very low-noise-figure preamplifiers to be used to advantage. The frequencies of the 2- and 1¼ meter bands are relatively close (less than one octave apart). Similar device and circuit design and selection criteria are valid. Furthermore, the 2-meter band is a common i-f for 1296 and 2304 MHz converters, wherein low-noise-figure i-f amplifiers may be used to advantage. The same 2-meter preamplifier may be used between a 2-meter receiver and either a 2-meter antenna or the i-f output of a 1296- or 2304-MHz mixer.

For truly low noise figures, the NE64535, priced at about $17, provides a 1-dB noise figure in a relatively simple circuit. This device is potentially unstable at all Amateur bands of interest; therefore, the collector circuit (**fig. 5**) is heavily loaded with a 100-ohm, 1/8-watt resistor.

The design is similar to that of WB5LUA (reference 18), but with added shielding between the device base and collector terminals. This shielding must be properly designed, because with improper shielding, the preamplifier exhibits a strong tendency towards oscillation, as noted by WB5LUA in his article. While the 432-MHz preamplifier of that article did not require shielding, at the lower frequencies of 144 and 220 MHz, the greater forward gain of these devices makes full shielding mandatory.

The preamplifiers are built in a box (**fig. 7**) formed of double-sided PC board with a double-sided PC-board shield (**fig. 6**), into which a small hole is drilled. The hole diameter is slightly larger than the device package dimension between the opposed emitter leads. The edges of both copper-clad sides of the shield are soldered to the bottom and two side walls of the basic enclosure. Most important, at least two small strips of copper foil, as found in most hobby shops, are passed through the hole and soldered on each side of the shield.

The opposed pair of emitter leads are soldered to the copper foil straps on the *input* side of the shield after the collector lead is passed through the hole in the field (see **figs. 6** and **7**). Thus, the base and emitter leads are on the input-circuit side of the shield, and the collector lead extends through the shield to the output-circuit side of the box. Failure to use the copper foil straps under the emitter leads, or soldering the emitter leads to the output-circuit side of the shield, will invariably cause oscillation, even with a resistive collector load.

Note particularly the use of ferrite beads, low-inductance coaxial feed-through capacitors, and the 0.05 μF low-frequency bypass capacitors in parallel with the feed-throughs, to prevent oscillations at fretable 1. Test results obtained by the author on solid-state devices used in forty-four preamplifiers covering 30-432 MHz.

frequency (MHz)	device	mfg.	cost	NF (dB)	gain (dB)	R(dB)	M(dB)	P	BW (MHz)
30	MA42001	MA	11.50	1.05	18.0	−33	1.06	20.67	1
	2N5109	—	1.55	1.44	14.4	−39	1.49	14.23	BB
	MRF-901	M	2.10	1.47	26.2	−34	1.47	14.85	BB
	MPF-102	M	0.35	1.60	16.0	−16	1.63	13.61	1
	2N4416	—	0.50	1.62	13.1	−22	1.69	14.67	1
	MRF-904	M	1.25	1.65	23.0	−34	1.66	15.36	BB
	NE02135	NEC	4.00	2.15	16.7	−47	2.19	26.28	BB
	DGFET	—	1.00[1]	4.80	12.0	−58	4.99	44.91	1
50	MA42001	MA	11.50	0.95	15.3	−34	0.97	18.92	2
	2N4416	—	0.50	1.37	13.0	−15	1.43	12.16	2
	NE02135	NEC	4.00	1.55	15.0	−27	1.59	19.08	BB
	MRF-901	M	2.10	1.67	25.0	−32	1.67	16.87	BB
	KD6003	MA	5.00[2]	1.70	13.7	−28	1.76	22.88	T144
	2N5109	—	1.55	1.72	18.0	−29	1.74	16.62	T50-BB
	DGFET	—	1.00[1]	2.00	13.5	−46	2.07	18.63	2
	MPF-102	M	0.35	2.00	13.2	−27	2.08	17.37	2
	MRF-904	M	1.25	2.00	28.0	−34	2.00	18.50	BB
	U-310	S	4.00	2.50	10.3	−26	2.69	32.28	2
	MP-1006	AND	11.00	3.70	27.0	−29	3.70	70.30	BB
144	NE64535	NEC	17.00	1.00	22.0	−36	1.00	25.00	BB
	MP-1006	AND	11.00	1.37	16.6	−28	1.40	26.60	BB
	MRF-901	M	2.10	1.40	23.0	−28	1.41	14.24	BB
	MRF-904	M	1.25	1.41	17.0	−32	1.43	13.23	BB
	U-310	S	4.00	1.60	12.0	−27	1.69	20.28	2
	KD-6003	MA	5.00[2]	1.67	21.5	−25	1.68	21.84	BB
	DGFET	—	1.00[1]	1.75	17.0	−28	1.78	16.03	3
	NE02135	NEC	4.00	1.81	23.5	−31	1.82	21.84	BB
	2N4416	—	0.50	1.90	17.5	−20	1.93	16.40	2
	2N5109	—	1.55	2.45	12.5	−23	2.56	24.45	T50
	NE22235	NEC	17.00	2.80	17.0	−38	2.84	71.00	BB
	MPF-102	M	0.35	3.65	11.0	−24	3.86	32.23	2
	J-308	S	1.25	6.20	13.0	−18	6.37	58.93	(several measured)
220	NE-64535	NEC	17.00	0.96	19.0	−34	0.97	24.27	BB
	MRF-904	M	1.25	1.35	14.5	−28	1.39	12.86	BB
	MRF-901	M	2.10	1.40	18.1	−24	1.42	14.34	BB
	MP-1006	AND	11.00	1.66	15.1	−28	1.71	33.35	8
	NE-02135	NEC	4.00	1.87	20.8	−30	1.88	22.56	BB
	KD-6003	MA	5.00[2]	1.95	16.5	−23	1.99	26.87	T144
	2N4416	—	0.50	2.00	10.0	−20	2.17	18.48	4
	U-310	S	4.00	2.00	9.1	−28	2.22	26.70	4
	NE22235	NEC	17.00	2.10	29.3	−26	2.10	48.71	BB
	2N5109	—	1.55	4.60	7.9	−21	5.10	51.23	T50
	DGFET	—	1.00[1]	5.00	13.5	−22	5.14	46.26	4
432	NE24483	NEC	120.00	0.74	15.3	−25	0.76	114.00	GaAs fet
	NE64535	NEC	17.00	0.86	16.0	−24	0.88	22.00	BB
	MSC-H001	MSC	40.00[3]	(1.00)	(20.0)	—	(1.00)	—	GaAs fet
	NE02135	NEC	4.00	1.27	11.2	−27	1.36	16.37	BB
	MRF-904	M	1.25	1.38	11.0	−25	1.48	13.69	BB
	MRF-901	M	2.10	1.40	16.1	−22	1.43	14.44	BB
	NE22235	NEC	17.00	1.60	11.0	−26	1.71	48.25	BB
	MP-1006	AND	11.00	1.90	17.7	−28	1.93	37.64	55
	MA42161	MA	>15.00[4]	(1.90)	(16.0)	—	(1.94)	—	—
	MP-1004	AND	14.00	2.00	12.2	−34	2.11	47.48	BB
	MP-1001	AND	6.00	2.40	16.5	−23	2.46	34.44	BB
	KD-6003	MA	5.00[2]	4.20	7.7	−23	4.73	62.79	T144
	U-310	S	4.00	4.70	6.0	−30	5.57	66.84	10

notes

(1) Experimental device — not generally available.
(2) No longer available; MA42003 is a substitute. KD6003 circuit (reference 6) tuned to 145 MHz for all measurements.
(3) Not built. See reference 7 for data.
(4) Not built. See reference 8 for data.

legend

BB Broadband. Input noise matched at measured frequency.
M Minimum noise figure (dB). (See text.)
P Performance ratio factor. (See text.)
R Return loss through preamp (dB). (See text.)
Txxx Tuned at frequency indicated for all measurements.

manufacturers

AND AND Transistors, 770 Airport Blvd., Burlingame, California 94010.
M Motorola Semiconductors, 4221 East Raymond, Phoenix, Arizona 85040.
MA Microwave Associates (G.R. Whitehouse & Co., 17 Newbury Drive, Amherst, New Hampshire 03031).
MSC Microwave Semiconductor Corporation, (HAM-TRANS, P.O. Box 383, South Bound Brook, New Jersey 08880).
NEC Nippon Electric Company (California Electronic Labs, 1 Edwards Court, Burlingame, California 94010).
S Siliconix, 2201 Laurelwood Road, Santa Clara, California 95050.

quencies removed from the frequency to which the preamplifier is tuned. Also note the biasing arrangement. A zener between transistor collector and base is *not* recommended because zeners are themselves often used as noise sources. Thus, even if bypassed, they will inject noise.

As previously stated, arrangement is made for adjusting both the voltage and current at the device collector; noise figure is affected by these parameters. The zener limits the maximum V_{ce} to avoid device burnout; the actual V_{ce} value, adjusted for minimum noise figure, is less than the zener voltage.

The preamplifier may be initially tuned for maximum gain but should be finally tuned using a noise generator. A vacuum-tube noise generator, such as one using a 5722 tube, is discouraged: voltage spikes

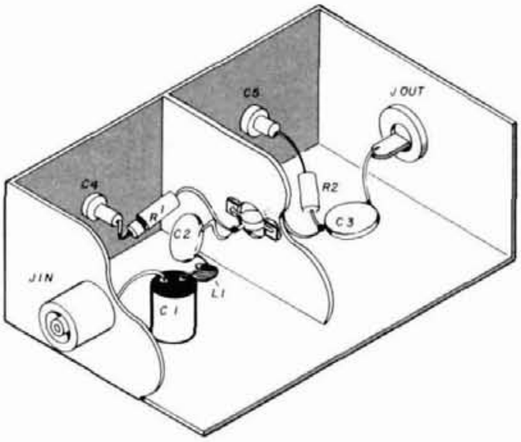


fig. 7. Isometric drawing showing internal construction.

of sufficient amplitude to destroy the transistor may be present.

A semiconductor noise generator is preferred, if available. If a precision noise-figure test set isn't available, you'll find that the units of this preamplifier are close to minimum noise figure when adjusted for maximum gain, at the manufacturer's suggested optimum bias level of 8 volts and 7 mA.

Slightly higher noise figures can be achieved with a variety of lower-cost devices, as listed in **table 1**. Of particular interest is the 2-meter preamplifier using the U310, which is unconditionally stable in a grounded-gate configuration.[19] It has tuned input and output circuits, achieving a relatively narrow bandpass characteristic. The narrow bandpass characteristic prevents preamplifier intermodulation and blockage problems caused by strong signals in the aircraft, business, and other adjacent bands.

432 MHz. This band may, to the purist, be considered above the vhf frequencies. However, it is the highest-frequency Amateur band at which point-to-point wiring techniques have been found to be generally usable. So it's included as the highest frequency band at which vhf-type preamplifiers may be easily built.

The listings of **table 1** illustrate that high-frequency versions of the 144- and 220-MHz preamplifiers previously discussed do, indeed, give performance unheard of twenty years ago. The best performance is, however, obtained by using microwave gallium arsenide field effect transistors (GaAs fet) devices such as the NE24483[20] or MSC-H001.[7] Because of the inordinately high cost and great susceptibility to damage of these devices, they are used mainly by moonbounce operators. The NE64535 or one of the MRF901 or NE02135 devices in a 432-MHz preamplifier if followed by a converter with a 3-dB input noise figure, will fulfill the needs of most operators of this band. The preamplifier must be installed at the antenna to realize this increase in sensitivity.

Values are shown for 432-MHz operation in the figures for 144-220 MHz preamplifiers having circuits directly extendable to 432-MHz. **Fig. 7** is a typical vhf preamplifier layout that can be adapted to most, if not all, of the preamplifiers discussed in this article.

further construction hints

I've tried both point-to-point and PC-board construction (**fig. 8**). Because full shielding of the preamplifier is desirable in addition to bias-lead filtering, PC technique is more costly and time consuming. Furthermore, if a double-sided PC board is used, with the unetched ground side as part of the shielding enclosure, the microstripline impedances of the circuit traces will have unexpected effects, particularly at 432 MHz. Unless the preamplifier is specifically designed to use microstripline (which will be considerably larger than a point-to-point wired preamplifier at 432 MHz), the low noise figure of modern-day transistors won't be realized.

Microstripline design generally assumes matching at least the resistive component of the input impedance to the data-sheet-specified device impedance. Because the device you use may have a slightly different impedance, optimum match can only be approached but never fully achieved. The illustrated, discrete-component, input-matching circuits allow adjustment for a range of resistive and reactive components of antenna and device impedances.

in summary

This article illustrates that truly low-noise vhf receiving preamplifiers can be built at the present time. Many of the preamplifiers built as part of this program have been tested not only on the lab bench

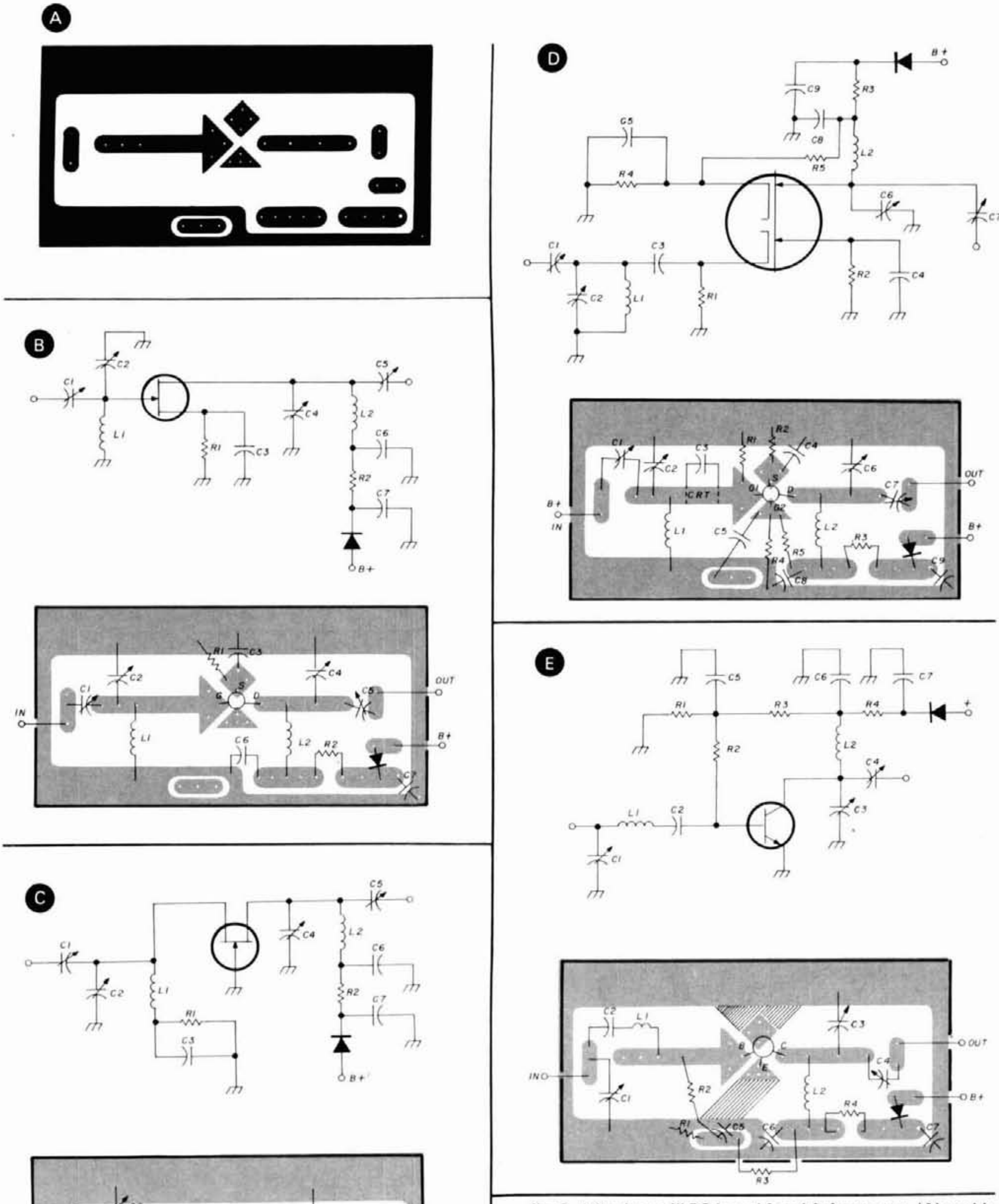


fig. 8. A "universal" PC board for vhf rf preamps, (A) usable to 225 MHz. Stripline effects above about 300 MHz may seriously affect operation of 432-MHz preamps if a two-sided board is used (reverse side as a ground plane). (B) through (E) show, respectively, circuits for a jfet, common source; jfet, common gate; dgfet, common source; and a bipolar transistor, common emitter. (Note that the arrow on the board points from input to output.)

but also under contest conditions by W2SZ/1 in various vhf and uhf contests over the past two years. Additional preamplifiers are constantly being built as new devices are received. And, depending on the results eventually derived from such new preamplifiers, an update may be forthcoming. Preamplifiers for 1296 and 2304 MHz are also being built and evaluated. An article covering these preamplifiers will be written when a sufficiently large number have been evaluated.

It should be kept in mind that all noise-figure measurements, especially those below about 2 dB, are not absolute but are *relative* indications of noise performance of the devices tested. The same test setup with circuits designed and built by one person may or may not result in the same answers obtained by others using identical circuits and procedures.

acknowlegments

I'd like to thank everyone who made devices available to me for construction and test of the preamps in this program. Special thanks are due the RPI Radio Club, W2SZ, for contest evaluation and, to Dick Frey, WB2BXP, for his help and encouragement.

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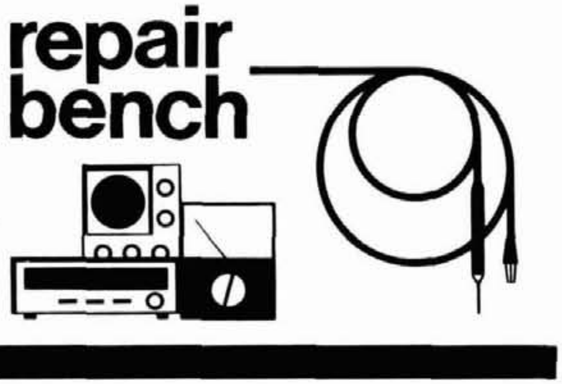
ham radio

More Details? CHECK — OFF Page 126



George Wilson, W1OLP

variable high-voltage supply

This workbench supply fulfills a need for test voltages in the 50-500 volt range. It can be built using mostly junkbox parts. It's particularly useful in reforming high-voltage electrolytic capacitors.

The title of this article originally was "The Care and Feeding of Electrolytic Capacitors." When it came time to take pencil in hand it appeared that the power supply needed to condition electrolytic capacitors has wider application and, hence, the more general (if less exciting) title. The supply described is meant to be used for light-duty experimental work. Its occasional utility on the workbench makes it a good investment of the construction time and cost for parts.

circuit description

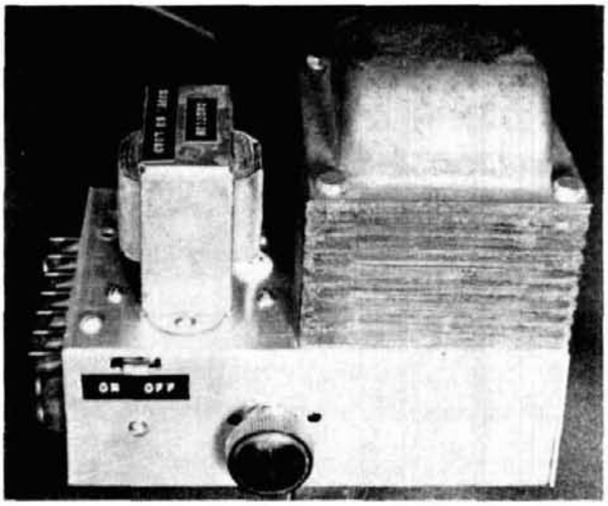
This is a *high-voltage* supply. Variable-voltage supplies for the range between zero and 50 volts are easily built with solid-state devices and have been described in the literature. The supply described here covers 50-500 Vdc. Output voltage adjustment to amounts less than 50 volts is touchy. So if you're interested in a power supply for lower voltages it's best to use one specifically designed for the purpose.

The circuit (**fig. 1**) is an ac voltage regulator (similar to those used in light dimmers and universal motor controllers) followed by a high-voltage transformer, rectifier and filter system. High-voltage silicon rectifiers are used. Vacuum-tube rectifiers would require a separate filament transformer and switch to provide constant voltage to the filaments. An LC filter is used to provide relatively ripple free dc. The choke also helps protect the 450-volt filter capacitor from current spikes. Note that this capacitor will be working close to its voltage rating at times. A 50-ohm 5-watt resistor can be substituted for the choke if you can't locate a suitable choke.

A separate ac switch is used rather than one connected to the voltage-control pot, which allows setting a voltage and turning the supply off and on without changing the set voltage. The bleeder resistor is primarily a safety device to help limit the voltage output (under light or no-load conditions) and to discharge the filter capacitor when the supply is turned off.

The component values in the ac voltage regulator circuit have been selected to provide relatively good voltage control over the 50-500 volt range. The triac circuit does not excel at the lower end of its control range, but its simplicity and low cost make it otherwise attractive.

(A) Author's experimental version of the variable high-voltage supply. The size and layout of yours will depend on the parts you can obtain. Note the separate on-off switch to allow voltage to be turned off and on without disturbing the variable setting. Terminal strips to mount small parts is recommended. A bottom plate will add safety. Don't omit the fuse. A low-value fuse is a good way to protect the circuit you're working on — a panel fuse mount will allow easy replacement.



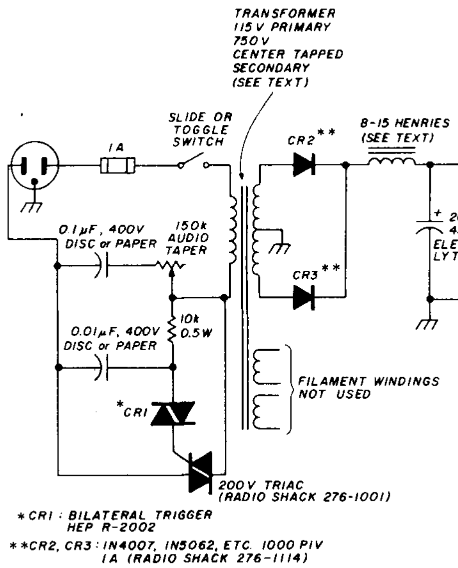


fig. 1. Variable high-voltage supply.

symbol	part	description
C1	capacitor	0.01 µF, 400 V disc or paper
C2	capacitor	0.01 µF, 400 V disc or paper
C3	capacitor	20 µF, 450 V electrolytic
CR1	bilateral trigger	HEP R-2002
CR2, CR3	diode	1N4007, 1N5062, etc. 1000 PIV 1 Ampere (Radio Shack 276-1114)
F	fuse	1 Ampere
L	choke	8-15 Henries (see text)
Q	triac	200-volt triac (Radio Shack 276-1001)
R1	potentiometer	150k audio taper
R2	resistor	10k 0.5 watt
R3, R4	resistor	100k 2 watt
T	transformer	115-volt primary, 750-volt center tapped secondary (see text)

Except for the transformer and choke, the components should be relatively easy to obtain. The best source for the transformer and choke is an old TV set or a flea market. The semiconductors may be Radio Shack or Motorola HEP devices. An audio taper pot is suggested for smooth low-voltage control; a linear control is acceptable. The maximum voltage may be limited by adding a resistor in series with the wiper end of the pot. Take care not to exceed the filter capacitor rating by more than a few volts.

uses

The supply can be used for experimental tube and transistor circuits calling for voltages within its range. Current output is limited by the transformer or the choke you can obtain. The supply is particularly useful in reconditioning (or reforming) electrolytic capacitors. If you're reactivating an old piece of equipment or using capacitors from the junkbox, it's always best (and frequently necessary) to reform the electrolytic capacitors before applying full voltage. If reforming is neglected, the capacitor may short (completely or partially) internally. This causes heating and promotes further shorting and eventual capacitor destruction. Reforming is accomplished by allowing the voltage across the capacitor to increase slowly to its rated value. This can be done by increasing the voltage slowly by means of the voltage control. Capacitor polarity *must* be observed: The positive side of the capacitor must go to the positive side of the supply.

A better method of reforming electrolytics is to allow them to reform through a high resistance. This may be done by connecting the capacitor to the supply through a 100k resistor (**fig. 2**). The resistor limits the current to less than 5 mA even if a shorted 450-volt capacitor is connected. A 2-watt resistor is recommended when reforming capacitors with working voltages above 300 volts. The current through the series resistor (and the power dissipated by it) will decrease rapidly if the capacitor can be reformed.

To use the setup in **fig. 2**, set the supply voltage (**A** to **C**) to the capacitor's rated voltage using a voltmeter. Then, with a high-impedance voltmeter connected to points **B** and **C**, measure the voltage across the capacitor. This voltage should increase slowly until the capacitor reforms completely. At this time the voltage across the supply (**A** to **C**) will be essentially equal to the voltage across the capacitor (**B** to **C**).

As in all high-voltage devices, be careful when

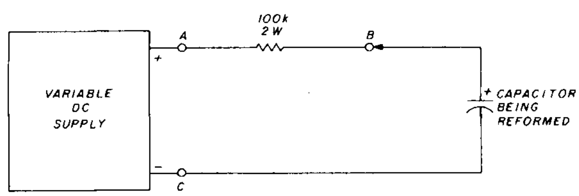


fig. 2. Circuit for reforming electrolytic capacitors. Procedure is described in the text.

assembling and using this device. *Dangerous voltages are present.* Make sure you use well-insulated clip leads in experimental setups and follow the old adage: "Keep one hand in your pocket when working with high voltages."

ham radio

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TR-7600

TR-7625

The TR-7600 and TR-7625 are Kenwood's popular synthesized 2-meter FM mobile transceivers. Combined with the RM-76 Microprocessor Control Unit, several memory and scanning capabilities are provided.

TR-7600/TR-7625 FEATURES:

- One memory channel.
- Mode switch for simplex or repeater operation. Repeater mode shifts the transmit frequency + 600 kHz or − 600 kHz or to the memory frequency.
- Full 5-kHz coverage from 144.000 to 147.995 MHz.
- Adaptable to any one MARS simplex or repeater channel between 143.7 and 148.3 (with modification kit).

KPS-7

The KPS-7 is a matching AC power supply for the TR-7600 and TR-7625. Output is 13.8 VDC at 7 A ICS (50% duty cycle).

RM-76

ADDED FEATURES WITH RM-76:

- Six memories.
- Automatic memory scan.
- Automatic scan up the band in 5-kHz steps, with selectable upper and lower frequency limits.
- Manual scan up or down the band in single or fast continuous 5-kHz steps.
- ± 1 MHz transmitter offset as well as ± 600 kHz and memory offset for repeater operation.
- MARS operation on 143.95 MHz simplex.
- Versatile digital display of transmit and receive frequencies, and operating functions.

TR-2400

The TR-2400 synthesized 2-meter hand-held transceiver features a large LCD frequency readout, 10 memories, scanning, and much more.

TR-2400 FEATURES:

- Large, illuminated LCD digital frequency readout. Readable in direct sunlight, and a lamp switch makes it readable in the dark. Shows receive and transmit frequencies and memory channels, and indicates "ON AIR", memory recall, battery status, and lamp switch on.
- 10 memories, with battery backup.
  - Automatic memory scan, for "busy" or "open" channels.
  - Mode switch for simplex, ± 600 kHz transmit repeater offset, and memory-frequency ("M 0") transmit repeater offset.
  - REVERSE momentary switch.
  - Built-in 16-button Touch-Tone generator.
  - Keyboard selection of 5-kHz channels from 144.00 to 147.995 MHz.
- Up/down manual scan and repeater or simplex operation from 143.900 to 148.495 MHz in single or fast continuous 5-kHz steps.
- Two lock switches to prevent accidental frequency change and accidental transmission.
- Subtone switch (subtone module not Kenwood supplied).
- More than 1.5 W RF output.
- High-impact plastic case and zinc die-cast frame.
- BNC antenna connector.
- Standard accessories included with the TR-2400 are a flexible rubberized antenna with BNC connector, ni-cad battery pack, and AC charger.

ST-1

OPTIONAL ACCESSORIES:

- Attractive leather case.
- Model ST-1 base stand, which provides 1.5-hour quick charge, trickle charge, and base-station operation with microphone connector and impedance-conversion circuit for using MC-30S microphone.
- Model BC-5 DC quick charger.

TOP CONTROLS

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any-state ni-cad charger

The advantage of constant-potential and constant-current charging techniques are incorporated into this circuit

Have you ever wanted to connect your hand-held to a charger and walk away, knowing the charger will charge the battery as quickly as possible, then automatically become a trickle charger to keep the battery charged until you need it again? Here's a circuit that does exactly that.

ni-cad battery charging

Before describing the charger, a brief review of ni-cad battery-charging technique is in order. There are two basic charging methods: constant potential and constant current.

Constant-potential charging. This is the most rapid method and requires no adjustment to the charger during the charge period. This method requires a charger capable of delivering high currents, because at the start of the charge, the battery has a very low internal impedance. This high-current charge tapers off exponentially as a function of charge time (**fig. 1**). The constant-charge voltage should be set at 1.55 times the number of cells in the battery.

A fully discharged ni-cad can be completely recharged by this method in one hour, although the charge should be continued for three hours or until the current stabilizes for one-and-a-half hours.[1]

The major disadvantage of the constant-potential charger is the high initial charge current. Not only does this require a voltage supply with high-current capability, but it's also possible to damage the battery being charged because of the high initial power being dissipated by the individual cells.

The usual method used to prevent cell damage with constant-potential charging is to monitor the temperature of a cell of the battery being charged and reduce the current to keep the temperature at some predetermined value. This is the system used with Motorola's "rapid charge" batteries and charger. The Motorola NLN-6900A rapid-charge battery has a thermistor that gives the charger battery temperature information.

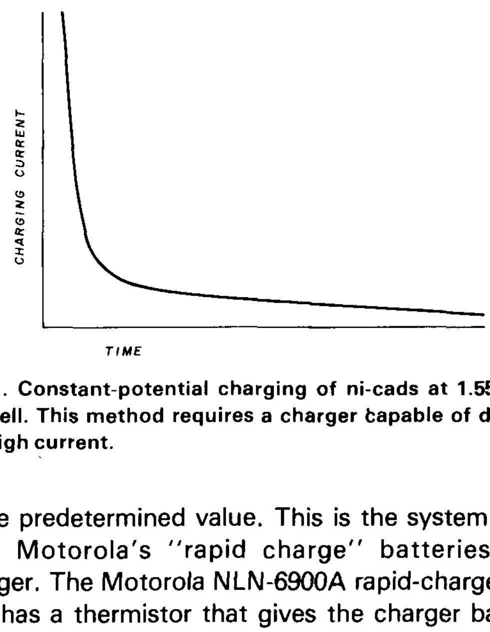

fig. 1. Constant-potential charging of ni-cads at 1.55 volts per cell. This method requires a charger capable of delivering high current.

Constant-current charging. The constant-current method requires a charging source of dc with a voltage of at least 1.8 times the number of cells in the battery. A simple constant-current charger is shown in **fig. 2**. To maintain constant current, the rheostat will require adjusting during the charge as the battery counter emf increases.

Practical values of charging current are usually

By Bill Bretz, WA6TBC, Hewlett Packard, 11300 Lomas Boulevard, N.E., Albuquerque, New Mexico 87123

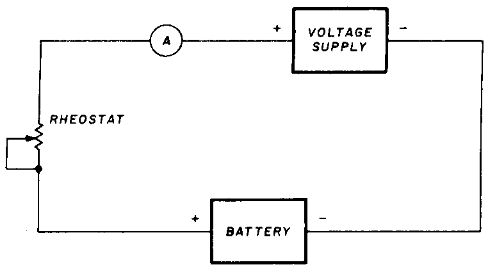

fig. 2. Simple constant-current charger. Constant current is maintained by adjusting the rheostat during the charge period.

selected to accomplish a full charge in three to seven hours. A 1-Ah battery might thus be charged at any current rate between 500 and 200 mA, since it is necessary to put more energy back into the battery than was taken out. A good working figure is to adjust the charging time so that the battery receives at least 30 per cent more energy than the discharge, measured in ampere hours. Cell voltage versus time is shown in **fig. 3**.

The major disadvantage of constant-current charging is that if the charge current is not terminated when the battery reaches full charge, the cell must dissipate power no longer needed for charging. This causes cell temperature to rise and also causes loss of electrolyte. The most common method of circumventing this problem is to charge the battery at a constant current of 10-20 per cent of its ampere-hour capacity for a minimum of twelve hours. Overcharging the battery at 10-20 per cent is not usually harmful (although not good) to the battery as long as there is ample electrolyte in the cell.

design

The charger shown in **fig. 4** is a constant-current, constant-potential battery charger that has the advantages of both the systems described above and the disadvantages of neither. With this charger, the current is initially constant at a reasonably high level until the battery nears full charge, at which time the current decreases and the charger becomes a constant-potential charger.

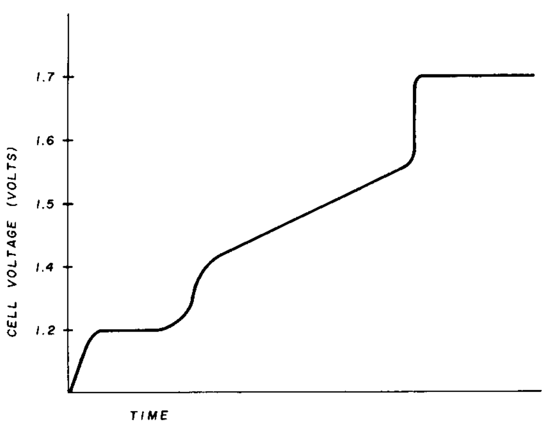

fig. 3. Cell voltage as a function of time using the constant-current charging method.

R3, R1, Q1, and CR1 form a constant-current source with R1 controlling the amount of current flowing in Q1 emitter. This is the major source of charging current for the battery. When the potential across the battery reaches the point where CR2 and Q2 turn on, Q2 starts pulling away Q1's supply of base current, which reduces the current from the current source, so that the potential across the battery is constant.

A practical charger for a 200-mAh, 12-volt nickel-cadmium battery is shown in **fig. 5**. This circuit charges the battery at 75 mA until the battery is charged, then reduces the current to a trickle rate. It

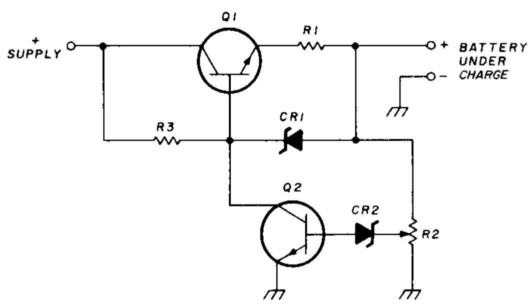

fig. 4. Basic design of the constant-current, constant-potential ni-cad battery charger, which has all the advantages of both techniques and the disadvantages of neither.

will completely recharge a dead battery in four hours and the battery can be left in the charger indefinitely. To set the shut-off point, connect a 270-ohm, 2-watt resistor across the charge terminals and adjust the pot for 15.5 volts across the resistor.

The circuit shown in **fig. 6** is built into the base of my HT-220 Omni charger. This circuit is built around the original Motorola power transformer and furnishes a constant-current charge of 200 mA (45 per cent of capacity). The charging lamp becomes a simple pilot lamp for the charger. The original "charge-trickle" switch now functions as an on-off switch for the charger. The pot is set for a voltage of 18.6 volts with a 240-ohm, 2-watt resistor across the charge terminals.

To modify this circuit to furnish different charge currents, the value of R1 may be determined by:

$$\frac{RL}{current\ desired} = 5\ volts$$

The constant-potential voltage can be determined by

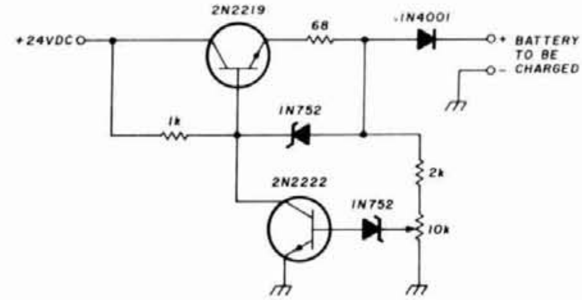

fig. 5. Practical charger for a 200-milliampere-hour, 12-volt ni-cad battery.

multiplying the number of cells by 1.55. The constant-potential voltage can be set by putting a load on the charger that pulls about one-half the desired maximum current and adjusting the pot for the desired voltage across this load. The supply voltage to the charger should be approximately two times the desired constant potential. This circuit will work for

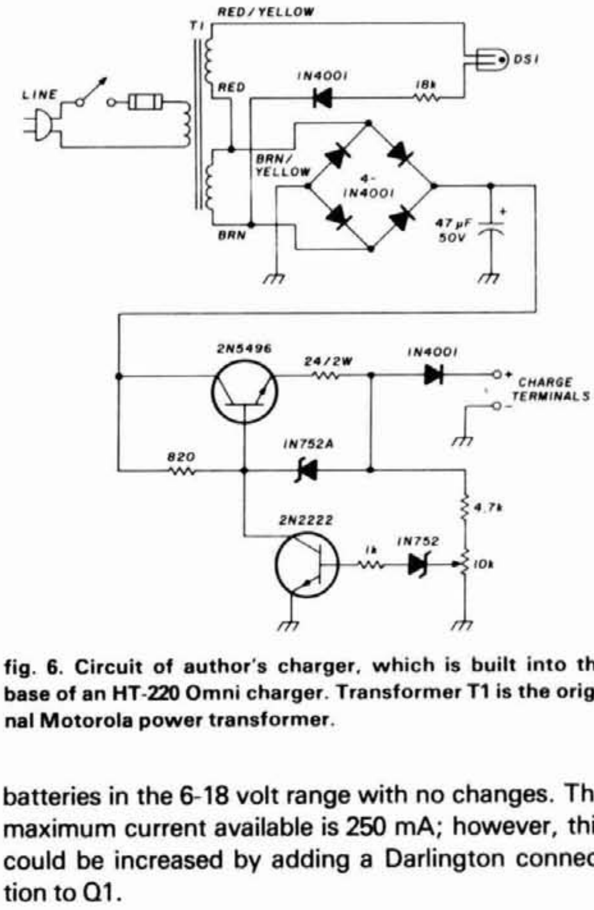

fig. 6. Circuit of author's charger, which is built into the base of an HT-220 Omni charger. Transformer T1 is the original Motorola power transformer.

batteries in the 6-18 volt range with no changes. The maximum current available is 250 mA; however, this could be increased by adding a Darlington connection to Q1.

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TS-1


TS-1JR


PE-2


SD-1

THE DAWNING

The age of tone control has come to Amateur Radio. What better way to utilize our ever diminishing resource of frequency spectrum? Sub-audible tone control allows several repeaters to share the same channel with minimal geographic separation. It allows protection from intermod and interference for repeaters, remote base stations, and autopatches. It even allows silent monitoring of our crowded simplex channels.

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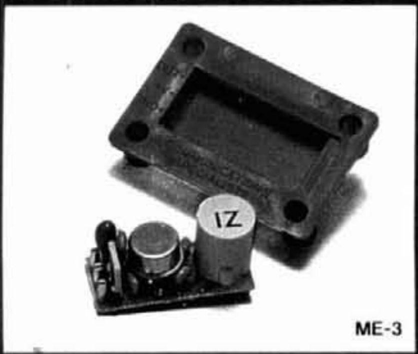
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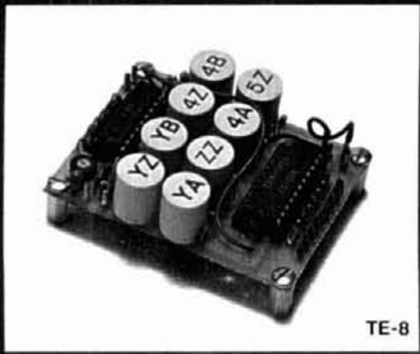
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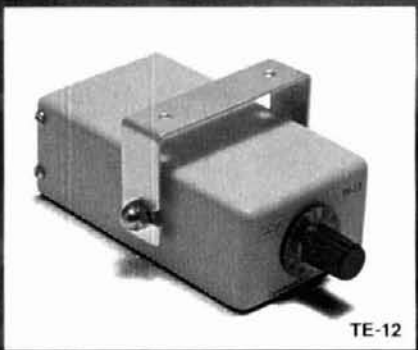
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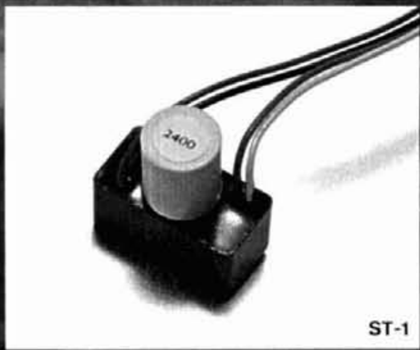
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SD-1 Two-Tone Sequential Decoder • Frequency range is 268.5 - 2109.4 Hz • Measures 1.2" x 1.67" x .65" • Momentary output for horn relay, latched output for call light and receiver muting built-in • **$59.95** with 2 K-2 elements.

TE-12 Twelve-Tone Sub-Audible or Burst-Tone Encoder • Frequency range is 67.0 - 263.0 Hz sub-audible or 1650 - 4200 Hz burst-tone • Measures 4.25" x 2.5" x 1.5" • **$79.95** with 12 K-1 elements.

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

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NEW products

Collins KWM-380 transceiver

Collins Telecommunications Products Division of Rockwell International Corporation introduced a new all-solid state Amateur transceiver in October. The new Collins KWM-380 transceiver is a self-contained Amateur station including an internal split frequency VFO function, a built-in ac dc power supply, and speaker. The fully synthesized KWM-380, with microprocessor controlled tuning, gives frequency stability and accuracy in four tuning rates, even down to 10 Hz. All frequencies are derived from a single master oscillator, allowing all bands and frequencies of 500 kHz to 30 MHz. Transmit power is 100 watts in upper sideband (USB), lower sideband (LSB), and CW modes within the 160- to 10-meter Amateur bands. In addition to providing four-speed tuning, the KWM-380's microprocessor controls the LED frequency display, band selection, and two-register memory for split-frequency operation without the need for an external VFO unit or separate receiver. The frequency display always shows the exact carrier frequency.

The KWM-380 has no bandswitch or tune-up controls; this enables the operator to transmit immediately upon dialing in a frequency. Even the transmit lowpass filters are automatically selected by relays. This feature provides split frequency operation on any two frequencies within the Amateur bands.

The front panel meter in the KWM-380 measures signal strength in receive; in transmit, it measures automatic level control (ALC), supply voltage (VC), forward power (FP), and reflected power (RF). Options for the new KWM-380 include a noise blanker for use in transceivers located in high-impulse rf noise environments, a choice of i-f filters for CW and RTTY, and related accessories. A speech processor will be available next year which will increase the average "talk-power" in SSB transmission.

Card Cage construction interconnected with ribbon cables allows the KWM-380 to be easily maintained. The cards can be removed for servicing while the radio is still operational, eliminating the need for extender boards.

The new Collins KWM-380 is available through any authorized Rockwell-Collins dealer/service agency.

new Larsen amateur antenna catalog

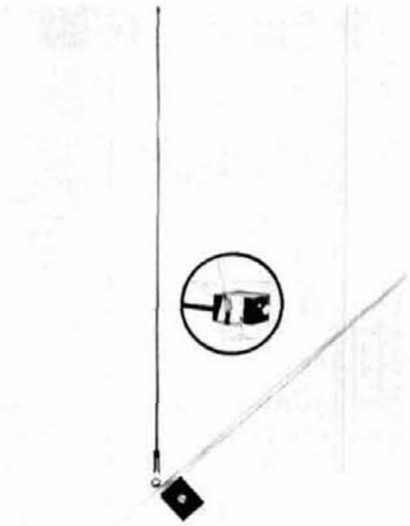
Larsen Electronics, Inc., one of the nation's leading mobile-antenna manufacturers, has just published a new catalog for its complete line of Kulrod (R) Amateur antennas.

In addition to mobile antennas, this 12-page brochure also includes base station antennas, Yagis, and antennas for hand-held and portable two-way radios. Available models cover the complete range of Amateur frequencies in the low, vhf, and uhf bands.

The catalog is conveniently arranged with antennas and corresponding mounts on the same page, so it's easy to order the right combination. There is also a complete price list for all equipment.

Copies of the Larsen Amateur Antenna Catalog are available at no charge from any Larsen dealer. They may also be obtained by writing to Larsen Electronics, attention John Beaman, P.O. Box 1686, Vancouver, Washington 98663.

Avanti no-holes 2-meter mobile antenna



From the research laboratory of Avanti Research and Development, Inc., of Addison, Illinois, comes a remarkable new concept in vhf mobile antennas.

It is an antenna that mounts on glass in minutes without tools, according to the manufacturer. No ground plane is required. There are no holes to drill.

A low-profile, one-inch, stainless steel mount holds the whip to the window by a new aerospace adhesive discovery that is stronger than a 1/4-20 metal bolt. Yet it can be easily removed, according to instructions, and is guaranteed by Avanti to hold securely under even abnormal weather conditions and excessive vehicular vibrations.

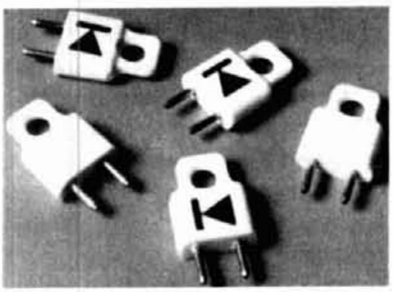
There are no external electrical connections to corrode, as the coax cable and capacitor box are mounted inside the vehicle. The new AH 151.3G antenna has been tested as 1 dB better than a conventional 5/8-wave, trunk-mount antenna, according to the manufacturer. It is also claimed to have a more uniform pattern than ground-plane type antennas.

Because this unique 1/2-wave design is mounted higher than a trunk-mount antenna, it offers a higher effective radiation point for maximum performance in all applications. The capacitive coupler forms a highly tuned circuit between the antenna and the radio to ensure maximum performance throughout the 2-meter band.

The stainless-steel whip and hardware connects to a chrome-plated casting. The tough ABS capacitive coupling box houses a tuning coil connected to the radio by a preassembled coaxial-cable assembly. All components have built-in Avanti quality.

For more information, contact Avanti Research and Development, Inc., 340 Stewart Avenue, Addison, Illinois 60101.

insulated diode-shorting plugs



New, insulated diode-shorting plugs that make custom programming panels and links simple, safe, and easy have been developed by the Cambridge Thermionic Corporation of Cambridge, Massachusetts. Cambion insulated diode-shorting plugs have the anode and cathode clearly marked on the bodies to reduce the possibility of polarity reversal. The diodes are molded into the insulation of the popular 0.040-inch diameter jumper plugs. Cambion insulated diode-shorting plugs with specific diode types are available manufactured-to-order.

The Cambridge firm offers a large selection of 0.040-inch diameter mating jacks suitable for mounting on many board and panel configurations to be used in conjunction with its new insulated diode-shorting plugs. Each diode plug, which is molded into a miniature handle-shaped holder, is designed for unidirectionally joining two points in any patch board with 0.040-inch jacks, printed circuit board, or connector having corresponding alignment and mating portions.

For spec sheets and complete information on these new Cambion insulated diode-shorting plugs, write Cambridge Thermionic Corporation, 445 Concord Avenue, Cambridge, Massachusetts 02138.

Nye Viking receives FCC phone-patch registration

Approval by the FCC for official registration under Part 68 of the FCC regulations has been granted to Wm. M. Nye Company, Bellevue, Washington, which allows the Nye Viking phone patch to be plugged directly into the telephone line without the need (or cost) of a telephone-company-supplied coupling device. However, users must still notify the telephone company that they are connecting the phone patch to the telephone line and must furnish to the company the official registration number and the ringer equivalence number, which are printed on an attached label.

Telephone patches may not be connected legally to party lines or pay telephone lines. Users are cautioned that they must comply with all other requirements of the FCC pertaining to Amateur Radio communications.

The Nye Viking phone patch comes in two models: No. 250-0046-001, without speaker, which provides connection to your own external speaker; and No. 250-0046-003, with built-in loud speaker, for use with most transceiver installations. Model -001 is priced at $45.50, and Model -003 at $55.00 throughout the U.S.A. These new units provide the finest interface connection to telephone lines.

Nye Viking phone patches manufactured before the official FCC approval and registration can be upgraded to approved status with the necessary changes, which include the 2-meter (7-foot) cord and plug to connect into the telephone company line socket. The charge for the complete modifications is $5 plus $1.50 for shipping and handling.

Units returned for modification should be carefully packed and contain your name and address, with check or money order covering modification charge and handling. Send by mail or UPS to Wm. M. Nye Company, Inc., 1614 130th Ave. N.E., Bellevue, Washington 98005. Expect a 2-3 week turnaround.

short circuit

updating vacuum-tube receivers

In W6HPH's article on "Updating Vacuum-Tube Receivers" in the December, 1978, issue of *ham radio*, the 2-volt zener diode (CR11 in **fig. 2**) should be replaced with a short circuit. In W6HPH's case, 2 volts was correct because Q12 happened to be a very low transconductance fet, but an average 2N3819 or MPF102 will be biased nearly to cutoff by 2 volts.

Also, in **fig. 2** the transistor in the upper right-hand corner should be marked Q2 — 2N2222, *not* Q12 — 2N3819; in **fig. 3** the trimmer capacitor to the left of L3 (L3 to ground) should be labeled C2. And in **fig. 4** the 5k trim-pot in the upper right-hand corner should be labeled R6; the bypass on pin 2 of Q17 is 0.001 μF.

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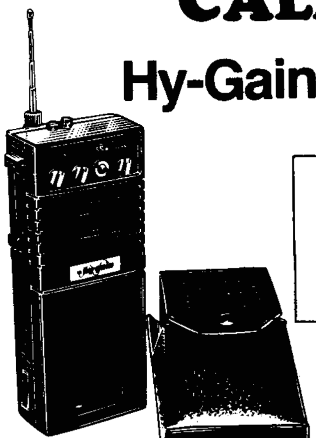

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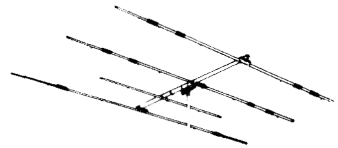

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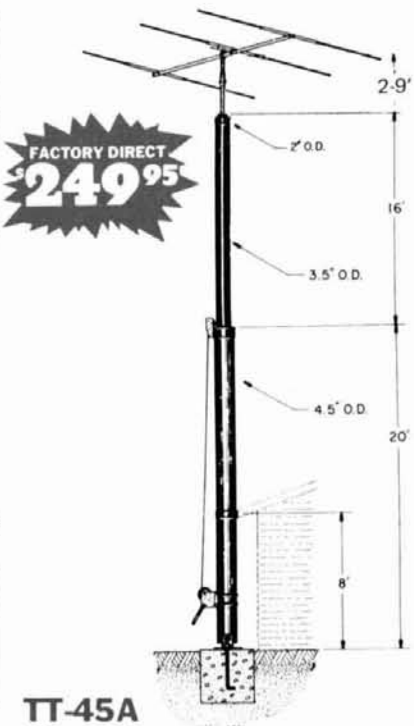
Mounting the House Bracket

The Hinged Base Plate allows tower to be tilted over for access to antenna and rotor from the ground.

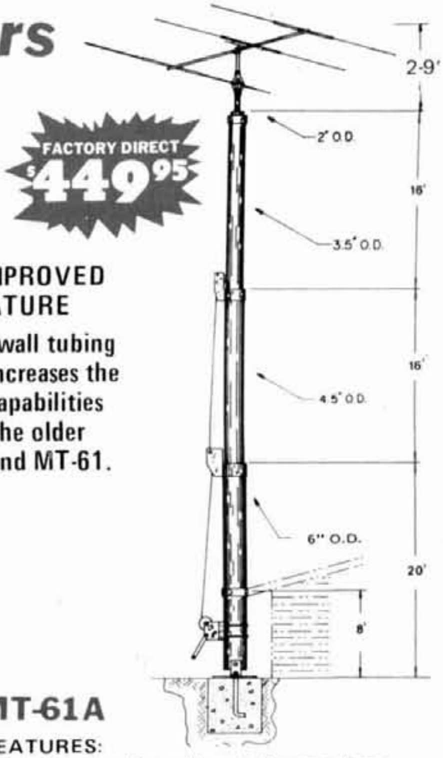


TT-45A

FEATURES:

- Maximum Height 45' (will handle 12 sq. ft. at 38') @ 50 mph
- 1200 lb. winch
- Totally freestanding with proper base
- Total Weight, 243 lbs.

The TT-45A is a freestanding tower, ideal for installations where guys cannot be used. If the tower is not being supported against the house, the proper base fixture accessory must be selected. (Requires 12"x12"x36" of concrete.)



NEW IMPROVED FEATURE

Heavier wall tubing greatly increases the stress capabilities over the older TT-45 and MT-61.

MT-61A

FEATURES:

- Is freestanding with use of proper base
- Maximum Height is 61' (will handle 12 sq. ft. at 53') @ 50 mph
- 1200 lb. brake winch
- 4200 lb. raising cable
- Total Weight, 400 lbs.

Recommended base accessory: RB-61A, FB-61A.

The MT-61A is our largest and tallest freestanding tower. By using the RB-61A rotating base fixture the MT-61A is ideally suited for the SY33 or SY-36. If you plan to mount the tower to your house, caution should be taken to make certain the eave is properly reinforced to handle the tower. If not, one of the base accessory fixtures should be used. (Requires 18"x18"x48" concrete.)

GENERAL FEATURES

All towers use high strength heavy galvanized steel tubing that conforms to ASTM specifications for years of maintenance-free service. The large diameters provide unexcelled strength. All welding is performed with state-of-the-art equipment. Top sections are 2" O.D. for proper antenna/rotor mounting. A 10' push-up mast is included in the top section of each tower. Hinge-over base plates are standard with each tower. The high loads of today's antennas make Wilson crank-ups a logical choice.

TILT-OVER BASES FOR TOWERS

FIXED BASE

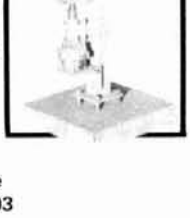
The FB Series was designed to provide an economical method of moving the tower away from the house. It will support the tower in a completely free-standing vertical position, while also having the capabilities of tilting the tower over to provide an easy access to the antenna. The rotor mounts at the top of the tower in the conventional manner, and will not rotate the complete tower. (Requires 3'x3'x5½' of concrete.)

FB-45A . . . $ 99.95
FB-61A . . . 129.95

ROTATING BASE

The RB Series was designed for the Amateur who wants the added convenience of being able to work on the rotor from the ground position. This series of bases will give that ease plus rotate the complete tower and antenna system by the use of a heavy duty thrust bearing at the base of the tower mounting position, while still being able to tilt the tower over when desiring to make changes on the antenna system. (Requires 3'x3'x6' of concrete.)

RB-45A . . . $139.95
RB-61A . . . 199.95



Tilting the tower over is a one-man task with the Wilson bases.

(Shown above is the RB-61A.)

(Rotor not included)

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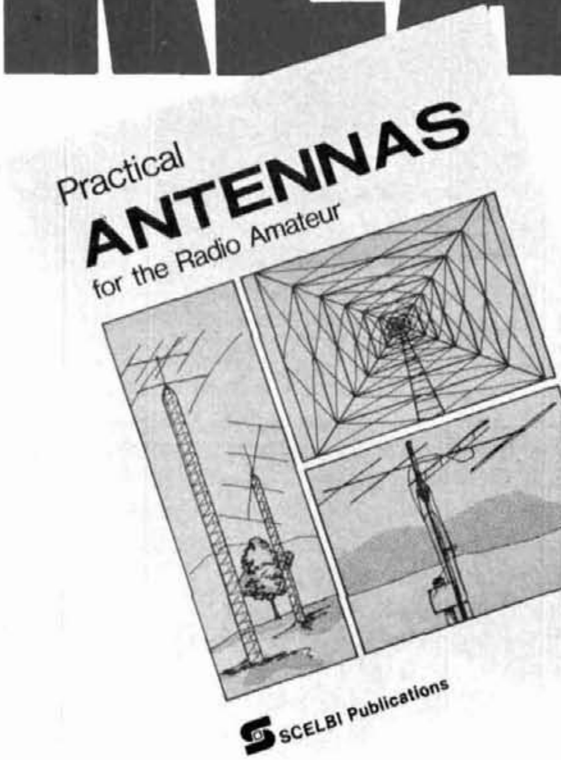

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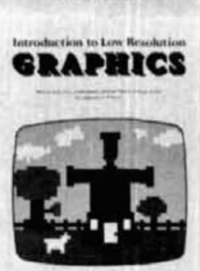

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REQUIRES 115 VAC AT LESS THAN 1/16 AMP.

COLLINS GRAY CABINET. WRINKLE PANEL - BRIGHT RED LED DIGITS (.33"). DECIMAL POINT IS THE PILOT LIGHT.

1.8-30 MHz

TWO SO-239 COAX CONNECTORS ARE AT THE REAR PANEL.

DIMENSIONS 3.5 x 5.5 x 7.5 INCHES.

WEIGHT IS 2 POUNDS.

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

WARRANTY ONE YEAR

SL-65 VSWR INDICATOR

- TWO DIGIT DISPLAY SHOWS VSWR TO AN ACCURACY OF .1 FOR VALUES FROM 1.0 AND 2.2. ACCURACY IS TO .2 FOR VALUES FROM 2.3 TO 3.4 AND TO .3 FROM 3.4 TO 4.0. FROM 4.1 TO 6.2 THE INDICATION MEANS THAT VSWR IS VERY HIGH.
- FOR VSWR VALUES NEAR 1.0, THE POWER RANGE FOR A VALID READING IS 20 - 2000 WATTS OUTPUT. FOR HIGHER VALUES THE UPPER POWER LIMIT FOR A FLICKER FREE VALID READING IS SOMEWHAT LESS (35 - 1000 WATTS FOR VSWR AT 2.0).
- DIVIDE THE ABOVE POWER LEVELS BY 100 TO OBTAIN THE PERFORMANCE OF THE SL-65A QRP MODEL.

SPECIAL AUDIO VERSION AVAILABLE FOR THE SIGHTLESS • WRITE FOR DETAILS •

SL-65 NET POWER INDICATOR

- THE POWER DISPLAYED IS THE DETECTED PEAK OF THE PEP FOR ANY MODULATION. THIS IS THE POWER THAT THE TRANSMITTER IS "TALKED" UP TO. DISPLAY DECAY TIME IS ABOUT ONE SECOND.
- THE POWER DISPLAYED IS THAT WHICH IS ACCEPTED BY THE ANTENNA (FORWARD LESS REFLECTED).
- POWER IS DISPLAYED ON THE SAME TWO DIGITS AS VSWR IN TWO AUTORANGED SCALES. 20 TO 500 WATTS AND 500 TO 2000 WATTS. TRIPOVER AT THE 500 WATT LEVEL IS AUTOMATIC EX: A READING OF 1.2 COULD MEAN 120 OR 1200 WATTS. YOU MUST KNOW WHICH RANGE YOU ARE IN.
- ACCURACY IS TO 10 WATTS IN THE LOWER RANGE AND 100 WATTS IN THE UPPER RANGE. DIVIDE POWER SPECS BY 100 FOR SL-65A.

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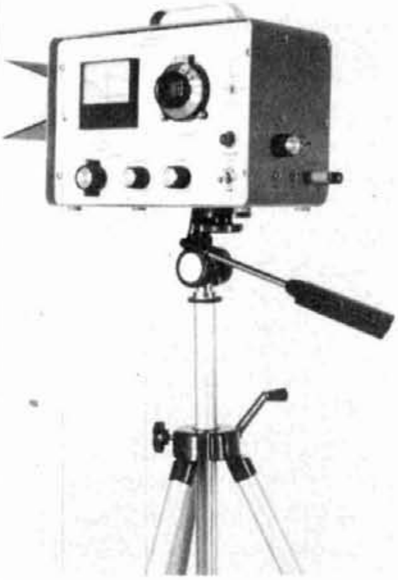

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VIDEO MODEL TVX10

- 10 GHz, 10 MW Gunn Transceiver
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- 45 MHz Low Noise Pre-Amp
- 45 MHz Post IF Amp/Demod
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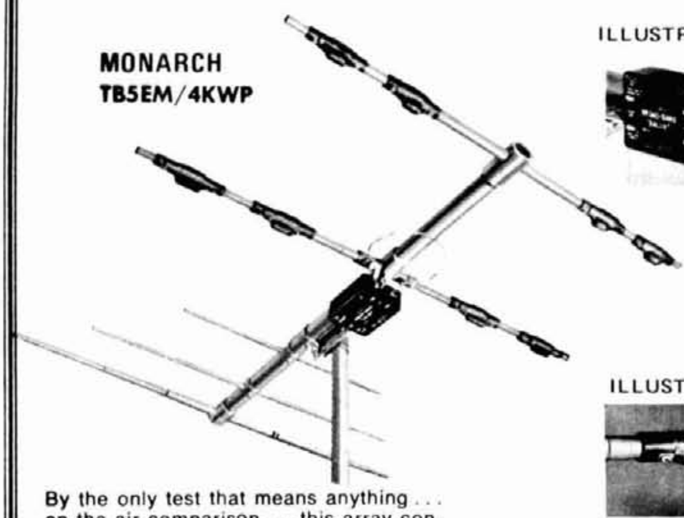

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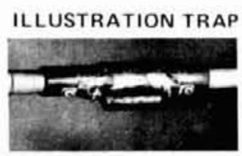

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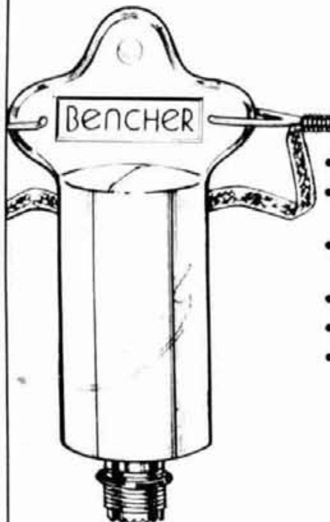

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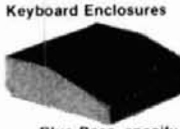

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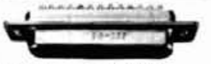

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M-CDP1802	User Manual	7.50
M-2650	User Manual	5.00

ROM'S

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

RAM'S

1101	256X1	Static	$1.49
1103	1024X1	Dynamic	.99
2101(8101)	256X4	Static	3.95
2102	1024X1	Static	1.75
21L02	1024X1	Static	1.95
2111(8111)	256X4	Static	3.95
2112	256X4	Static MOS	4.95
2114	1024X4	Static 450ns	7.95
2114L	1024X4	Static 450ns low power	10.95
2114-3	1024X4	Static 300ns	10.95
2114L-3	1024X4	Static 300ns low power	11.95
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5280/2107	4096X1	Dynamic	4.95
7489	16X4	Static	1.75
74S200	256X1	Static Tristate	4.95
93421	256X1	Static	2.95
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UPD416 (MK4116)	16K	Dynamic 16 pin 250ns	9.95
TMS4044-45NL	4K	Static	14.95
TMS4045	1024X4	Static	14.95
2117	16,384X1	Dynamic 350ns (house marked)	9.95
MM5262	2KX1	Dynamic	4/1.00

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1702A	2048	FAMOS	$5.95
2716INTEL	16K*	EPROM	59.95
TMS2516 (2716)	16K*	EPROM	49.95
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TMS2532	4KX8	EPROM	89.95
2708	8K	EPROM	10.95
2716 T.I.	16K**	EPROM	29.95
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5203	2048	FAMOS	14.95
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FEATURES:
- Full 8 bit latched output for micro-processor use
- 3 User Define keys with one being bi-stable operation
- Debounce circuit provided for all 19 keys
- LED readout to verify entries
- Easy interfacing with standard 16 pin 1C connector
- Only +5VDC required for operations

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JE600 .... $59.95
Hexadecimal Keypad only $14.95

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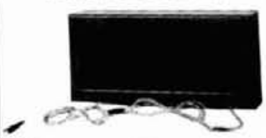

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FEATURES:
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- Utilizes a 2376 (40 pin) encoder read only memory chip
- Outputs directly compatible with TTL/DTL or MOS logic arrays
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- **Data Format** .... Asynchronous Serial (return to mark level required between each character).
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- **Transmit Channel Frequencies** .. Switch selectable: Low (normal) = 1070 space, 1270 mark; High = 025 space, 2225 mark.
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- **Transmit Level** .... –15 dbm nominal. Adjustable from –6 dbm to –20 dbm.
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- **Physical** .... All components mount on a single 5" by 9" printed circuit board. All components included.

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Expand your 4K TRS-80 System to 16K.
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- Documentation for conversion

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- No stripping or slitting required—just wrap

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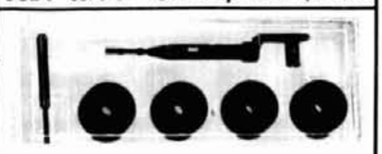

JUST WRAP™ Kit

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- JUST WRAP Tool
- Unwrapping Tool

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VV-1 .... $3.49

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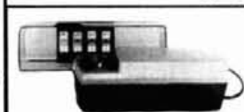

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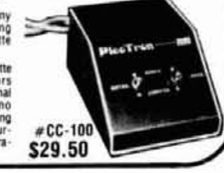

#CC-100 $29.50

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The MorseMatic is also available strictly as a non-memory keyer for only $79.95 (introductory price). For your free MorseMatic booklet explaining these and more features, or for information about ordering a MorseMatic, write or call Advanced Electronic Applications, Inc., P.O. Box 2160, Lynnwood, WA 98036. Phone 206/775-7373. (Dealer inquiries invited.)

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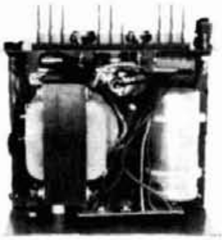

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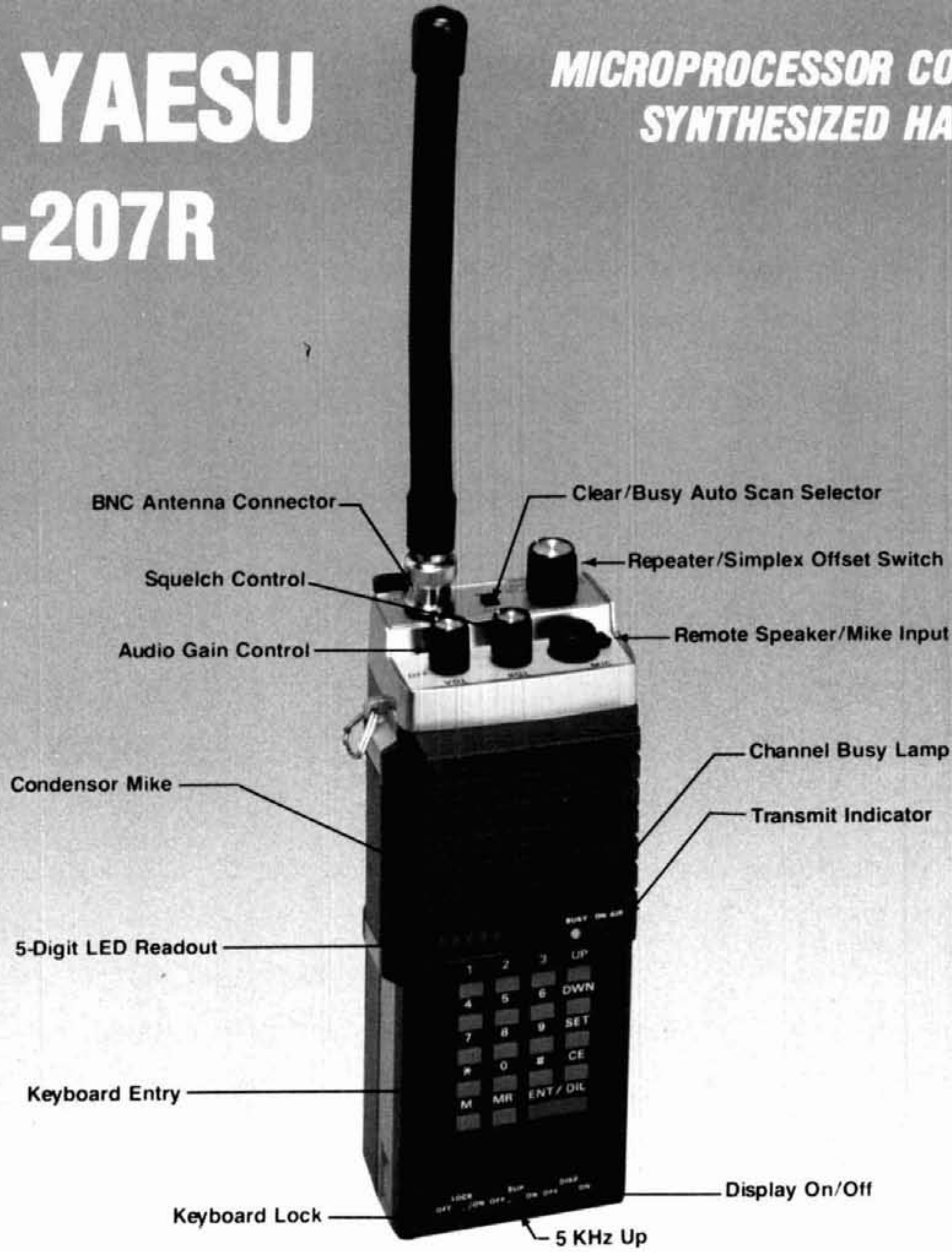

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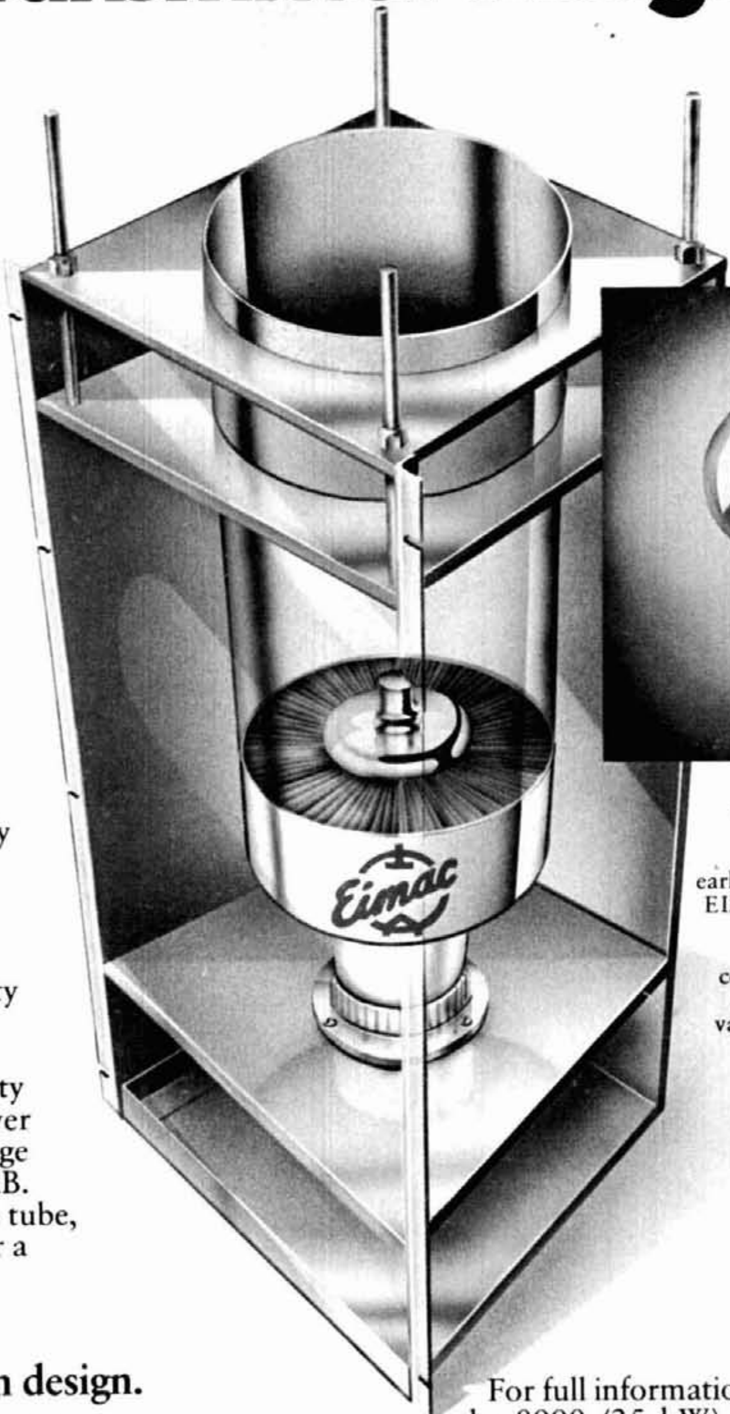

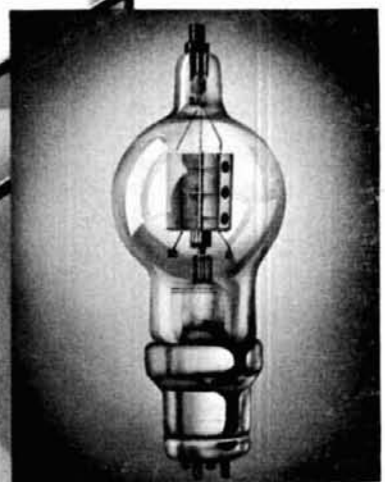

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