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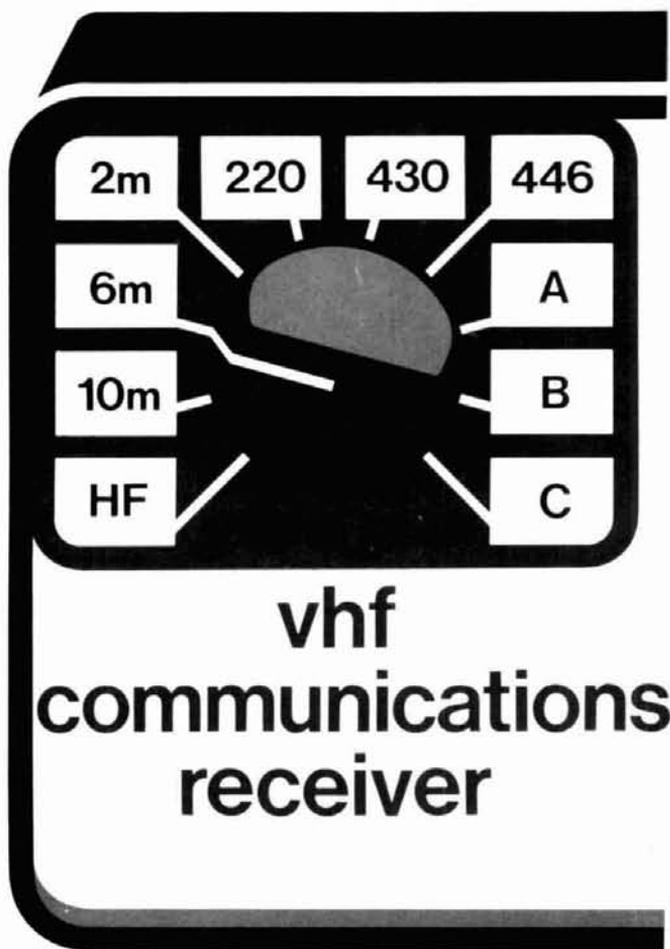
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JULY 1978

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- and much more . . .



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TEMPO FMH-2, FMH-5 & FMH-42 (UHF)

...portables you can depend on at a price you can afford

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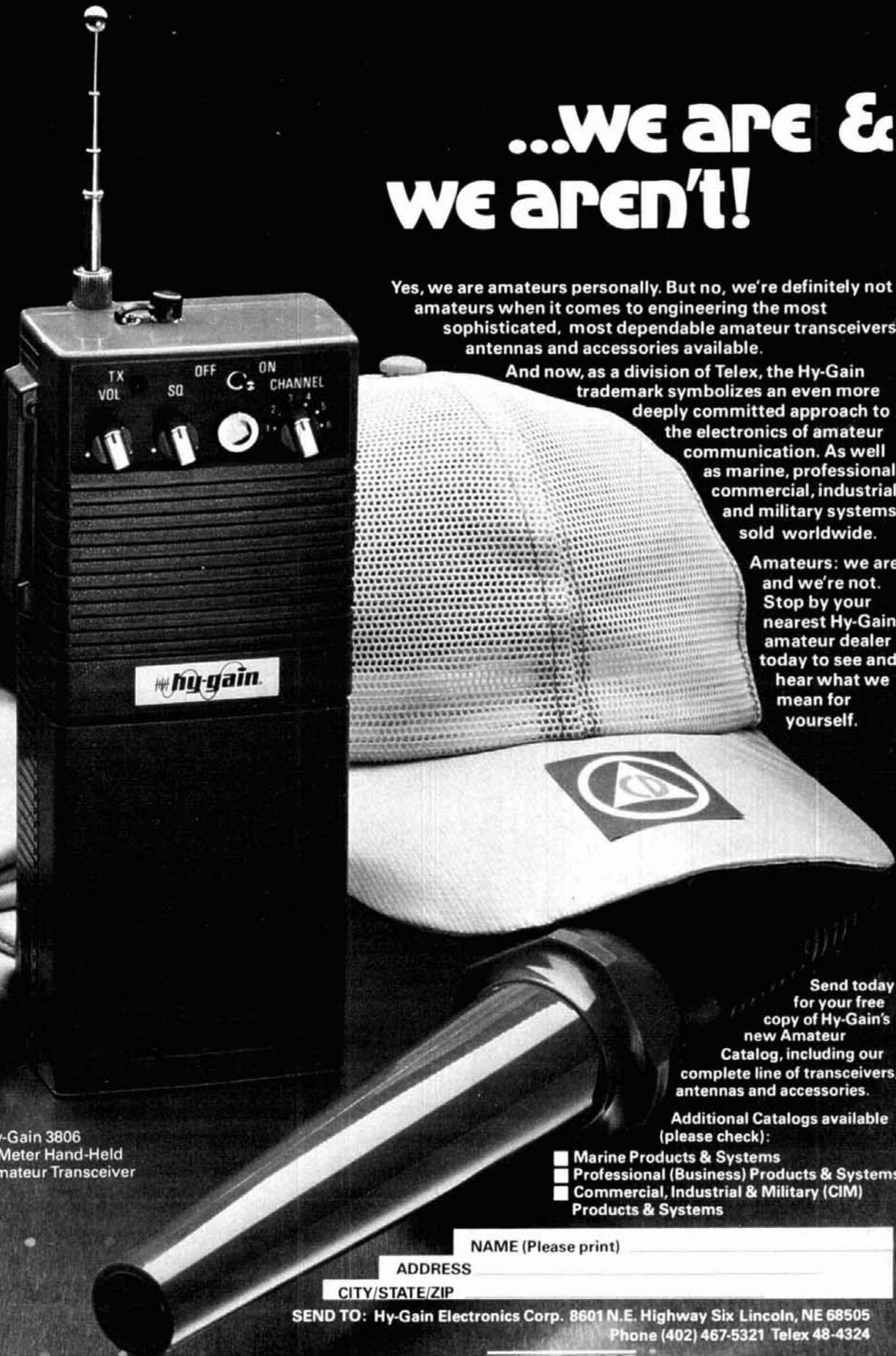
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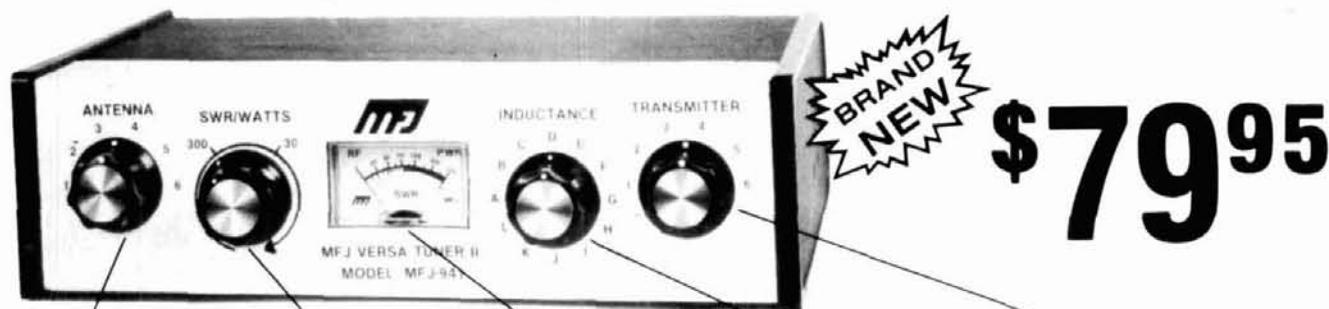
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Hy-Gain Electronics A Division of **TELEX** Communications

This NEW MFJ Versa Tuner II . . .

has SWR and dual range wattmeter, antenna switch, efficient airwound inductor, built in balun. Up to 300 watts RF output. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines.



Antenna matching capacitor. 208 pf. 1000 volt spacing.

Sets power range, 300 and 30 watts. Pull for SWR.

Meter reads SWR and RF watts in 2 ranges.

Efficient airwound induc-tor gives more watts out and less losses.

Transmitter matching capacitor. 208 pf. 1000 volt spacing.

Only MFJ gives you this MFJ-941 Versa Tuner II with all these features at this price:

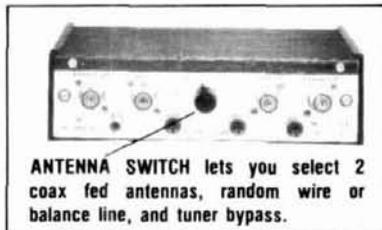
A SWR and dual range wattmeter (300 and 30 watts full scale) lets you measure RF power output for simplified tuning.

An antenna switch lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

A new efficient airwound inductor (12 positions) gives you less losses than a tapped toroid for more watts out.

A 1:4 balun for balance lines. 1000 volt capacitor spacing. Mounting brackets for mobile installations (not shown).

With the NEW MFJ Versa Tuner II you can run your full transceiver power output — up to 300 watts RF power output — and match your



ANTENNA SWITCH lets you select 2 coax fed antennas, random wire or balance line, and tuner bypass.

transmitter to **any** feedline from 160 thru 10 Meters whether you have coax cable, balance line, or random wire.

You can tune out the SWR on your dipole, inverted vee, random wire, vertical, mobile whip, beam, quad, or whatever you have.

You can even operate all bands with just

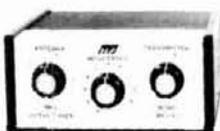
one existing antenna. No need to put up separate antennas for each band.

Increase the usable bandwidth of your mobile whip by tuning out the SWR from inside your car. Works great with all solid state rigs (like the Atlas) and with all tube type rigs.

It travels well, too. Its ultra compact size 8x2x6 inches fit easily in a small corner of your suitcase.

This beautiful little tuner is housed in a deluxe eggshell white Ten-Tec enclosure with walnut grain sides.

S0-239 coax connectors are provided for transmitter input and coax fed antennas. Quality five way binding posts are used for the balance line inputs (2), random wire input (1), and ground (1).



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MFJ-901 VERSA TUNER

New efficient air wound coil for more watts out.

Only MFJ uses an efficient air wound inductor (12 positions) in this class of tuners to give you more watts out and less losses than a tapped toroid. Matches everything from 160 thru 10 Meters: dipoles, inverted vees, random wires, verticals, mobile whips, beams, balance lines, coax lines. Up to 200 watts RF output. 1:4 balun for balance lines. Tune out the SWR of your mobile whip from inside your car. Works with all rigs. Ultra compact 5x2x6 inches. S0-239 connectors. 5 way binding posts. Ten Tec enclosure.



\$49.95

BRAND NEW

MFJ-900 ECONO TUNER

Same as MFJ-901 Versa Tuner, but does not have built-in balun for balance lines. Tunes coax lines and random lines.



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Operate 160 thru 10 Meters. Up to 200 watts RF output. Matches high and low impedances. 12 position inductor. S0-239 connectors. 2x3x4 inches. Matches 25 to 200 ohms at 1.8 MHz.



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JULY 1978

volume 11, number 7

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

subscription rates

United States: one year, \$12.00
two years, \$22.00; three years, \$30.00

Canada: one year, \$14.00
two years, \$26.00; three years, \$36.00

Europe, Japan, Africa:
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North America, South America, Australia
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foreign subscription agents

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Microfilm copies
are available from
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Ann Arbor, Michigan 48106
Order publication number 3076

Cassette tapes of selected articles
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Second-class postage
paid at Greenville, N. H. 03048
and at additional mailing offices
Publication number 233340



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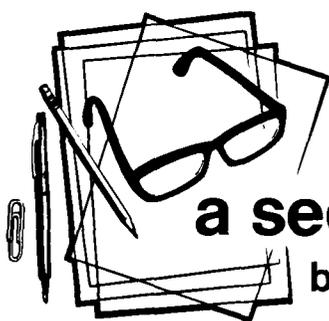
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a second look

by Jim Fisk

In our modern day world of solid-state electronic gadgets and centralized urban living, it's the rare amateur who hasn't been troubled at one time or another by interference complaints. As often as not the interference is caused by some other source, but if you have a tower in your backyard, you're a likely suspect and the first one to whom they turn when the local taxicabs (or whatever) tear up your neighbor's favorite television show or come booming through their quadraphonic stereo system.

As I have mentioned in this column several times in the past, the problem can be effectively cured only by proper design and construction of home-entertainment equipment at the manufacturing level. The consumer electronics business is highly competitive, however, so the manufacturers are reluctant to add filtering and lead bypassing that would increase the sales price of their equipment. For many years the manufacturers contended that less than 5% of home entertainment equipment operated in an rf environment which required special attention — but with the proliferation of two-way radio systems as well as higher power a-m and fm broadcasting stations and high-speed digital systems which can cause interference, I doubt that many consumers would agree.

Several bills have been introduced into Congress which would give the FCC authority to regulate the manufacture of home-entertainment devices to reduce their susceptibility to interference from nearby radio transmitters, but none have passed. Now Senator Goldwater is sponsoring a Bill which would require better RFI rejection; the Bill, S-864, has been referred to the Senate Subcommittee on Communications and hearings began in Washington on June 14th. Among those invited to testify were the ARRL, FCC, Institute of High Fidelity, and Heath. Although there's no chance that the Goldwater Bill will make it to the Senate floor during this session, the hearings will help pave the way for speedier action on future RFI legislation.

Consumers are becoming increasingly aware of the RFI problem, so the time is right for legislation such as that proposed by Senator Goldwater. Radio amateurs have known for a long time that the majority of RFI problems are not due to interference *per se*, but are caused by the interception of signals by devices which were not designed to operate in today's strong rf environment. The only way to eliminate 90% of the RFI problems is through legislation such as S-864 which would eventually require the manufacturers to correct those design deficiencies which lead to unnecessary interference.

Individual amateurs can help toward the eventual passage of a bill requiring better RFI rejection by letting their Senators know of their support for S-864, particularly if one of their Senators is a subcommittee member. In addition to Chairman Hollings (South Carolina), the members are Griffin (Michigan), Magnuson (Washington), Cannon (Nevada), Inouye (Hawaii), Ford (Kentucky), Durkin (New Hampshire), Zorinsky (Nebraska), Riegle (Michigan), Stevens (Alaska), Packwood (Oregon), Schmitt (New Mexico), and Danforth (Missouri). Letters to the Senators addressed to the United States Senate, Washington, D.C. 20510, will reach them promptly and may help considerably.

The letters do not have to be long, although background information on your (or your neighbors') RFI problems could be important. Even a note to the effect that you support S-864 would be a valuable contribution. Remember that previously introduced RFI legislation never made it through Congress — now that Senator Goldwater has started the ball rolling again, let's make sure it has enough momentum to become law. Now is the time to lend your support to this vital effort; write today and make your voice heard.

Jim Fisk, W1HR
editor-in-chief



ACTUAL SIZE PHOTO

Try on MultiMode Mobile

Multi-mode mobile is the newest trend in VHF. The next time you are lamenting the limitations of your present mobile rig, just imagine how it would size up against the **IC-245/SSB Mobile Maximizer**. In addition to the complete FM capability through repeaters and the SSB capabilities from a home station, the **IC-245/SSB** offers superb SSB mobile performance to squeeze the most out of simplex mobile contacts. Mobil CW? No problem with the **IC-245/SSB**: just push the button and hit the key, and you're on the best mode for longhaul mobile DX. Some of the features which make all of this possible are:

- **Single knob frequency selection:** The **IC-245/SSB** is synthesized with convenient single knob frequency selection over the entire 4 MHz. No more fussing with two or more knobs just to check what is going on around the band. One easy spin of the dial does it all.
- **Two VFO's built in:** The second VFO, which is an optional tack-on with most other transceivers, is an integral feature in every **IC-245/SSB**.
- **Variable offset:** Any offset from 10 KHz through 4 MHz, in multiples of 10 KHz, can be programmed with the LSI synthesizer.
- **Remote programming:** The LSI chip provides for input of a Touch Tone® like programming pad from an external source, such as the microprocessor controlled accessory which will also provide scan and other functions (available summer '78). Computer control from a PIA interface is also possible (data available on request).
- **FM stability on SSB and CW:** Synthesis of 100 Hz steps makes SSB as stable as FM. This extended range of operation is attracting many FM'ers who have been operating on the direct channels and have now discovered SSB.

So for your next mobile radio, go all out after all of it and get the maximum in multi-mode mobile with the **IC-245/SSB**. **P.S.** A microprocessor controller with memory, frequency setting and Touch Tone® dialing will be available soon.

Maximize the new repeater band: both the **IC-211** and the **IC-245/SSB** operate the new FCC repeater spectrum with no modification.

All ICOM radios significantly exceed FCC specifications limiting spurious emissions.

Specifications: Frequency Coverage: 144.00 to 148.00 MHz Modes: FM (F3), SSB (A3J), CW (A1) Supply Voltage: DC 13.8V ± 15% Size: 90mm (h) x 155mm(w) x 235mm(d) Weight: 6.8 Kg TX Output: F3, 10W; A3J, 10W (PLP); A1, 10W Spurious Radiation: -60 dB below Carrier Microphone Impedance: 600 Ohms Sensitivity: A3J & A1, 0.5 microvolt input gives 10 dB S/N; F3, 0.6 microvolt or less for 20 dB quieting S/N-D/N, at 1 microvolt input, 30 dB Spurious Response: -60 dB or better Synthesizer Frequency Range: 144.00 MHz to 148.00 MHz Synthesizer Step Size: 5 KHz for FM, 100 Hz or 5KHz for SSB

Specifications are subject to change without notice.

HF/VHF/UHF AMATEUR AND MARINE COMMUNICATION EQUIPMENT

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NEW 10, 18, AND 25 MHZ AMATEUR bands were all proposed in the FCC's WARC 79 eighth Notice of Inquiry released in early May. On the negative side, the Commission proposed shifting 50 kHz from the top end of 80, and the bottom 60 kHz of 160, to the broadcast service, but also proposed that all the HF Amateur bands (except for part of 160) be exclusively Amateur throughout the rest of the world. Specific FCC proposals for the Amateur Service:

160 Meters: 1800-1860 kHz, Region 3 only (shared); 1860-1900 exclusive Amateur worldwide; 1900-2000 shared
80 Meters: 3500-3950 kHz (loss of 50 kHz)
40 Meters: 6950-7250 kHz (shifted down 50 kHz)
30 Meters: 10.1-10.2 MHz (new)
20 Meters: 14.0-14.35 MHz (unchanged)
17 Meters: 18.068-18.168 MHz (new)
15 Meters: 20.95-21.45 MHz (50 kHz added at low end)
13 Meters: 25.11-25.21 MHz (new, moved from 25.76 proposed in the fifth NOI at the request of Radio Astronomy)
10 Meters: 28.0-29.7 MHz (unchanged)

Arguments By Amateurs were responsible for the new 10 and 18 MHz bands, the Commission noting that they were "so strongly requested and justified in the Service Working Group report while reducing the overall impact on the fixed services..."

These Proposed Bands are far from firm, however. There is a mention in the text that the Executive Branch will not support the 6950-7000 and 20950-21000 kHz slots or allow the Amateur Satellite Service to use 1250-1260 or 5660-5670 MHz (which in addition to 2390-2400 MHz, 76-81, 165-170, and 240-250 GHz, were included as a footnote with 435-438 MHz). The Amateur Satellite Service was also given "primary" status on the new 13-meter band.

CREDIT FOR PASSING THE CW exam can now be retained by an FCC license applicant even though he failed the written test; the Commissioners agreed to allow an applicant such credit for up to one year after he takes an exam. To receive the CW credit the applicant will be given an FCC Form 845 by the Field Office at the time of the exam. He turns it in when he returns for another try and receives credit for the CW portion of the exam. Form 845 will be honored only by the Field Office that issued it.

Ex-Novices, Whose Licenses expired less than a year ago, will be eligible to apply for reinstatement just like any other Amateure licensees. This provision is a result of the FCC's recent rules change making Novice licenses good for five years and renewable. Any former Novice, whose license expired after May 15, 1977, can submit a Form 610 (within one year of his original expiration date) and regain his license for a full five years — without retaking an exam.

General-Class Licensees are now permitted use of 50.0-50.1 MHz. The change brings General 6-meter privileges in line with those of Technicians.

AMATEUR RADIO'S RANKS GREW over 11 per cent in the past year, according to recent FCC figures. At the end of April, U.S. Amateur operator licenses numbered 337,959, compared to only 304,331 a year ago. Biggest growth was in the Novice Class, up 12,434 from last year at this time to total 58,016. Extra Class showed the next greatest percentage increase, up more than 20 per cent to 19,797, but every class showed healthy growth rates of 6 per cent or more.

W2BXA RECEIVED SATELLITE DXCC number 1 when he arrived at ARRL Headquarters with cards proving he'd worked 101 countries via the OSCARS! W6VPH/VP5, VP2EFZ, and a special hand-made FOØXA card from the recent Clipperton DXpedition pushed Ben over the top.

K5CM WORKED WAØLPK/KL7 FOR HIS 50TH 2-METER state in late April, becoming the second station to make WAS on 144 MHz.

In Addition, 2-Meter Worked All States was achieved by NØJA and K9HMB several weeks later, thanks to N6NB/7, who operated his portable moonbounce station from both sides of the Utah-Nevada border.

THE WINNER OF THE GRAND PRIZE in the 1978 Sweepstakes, a Drake UV-3, the world's only three-band vhf-fm transceiver, with ac power supply and encoding microphone, and an Atlas 350-XL high-frequency transceiver package which includes a deluxe power supply console with speaker, digital clock, phone patch, speech processor, and electronic keyer, is Shelton Boles, WA5KOK, of Cleveland, Texas. Other happy winners in this year's Sweepstakes are Murray Fisher, W7NSU, who won an Atlas 350-XL with ac power supply, and David Richards, K1YGF, who is a proud new owner of a Drake UV-3 with modules for 144, 220, and 440 MHz.

The evolution of the MLA

When the MLA-2500 was first introduced it was a new concept in high performance amplifiers. Low and sleek yet powerful enough for the military. Some wondered . . . needlessly.

A promise kept.

The MLA-2500 promised 2000 watts PEP input on SSB. A heavy duty power supply. Two Eimac 8875's. And as thousands of Amateurs across the world have proven, the MLA-2500 delivers!

Now DenTron is pleased to bring you **The new MLA-2500 B.** Inherently the same as the original MLA-2500, the B model includes all of the above specifications plus a few refinements. New high-low power switching for consistent efficiency at both the 1KW and 2KW power levels, and 160 - 15 meters.

Tested and proven.

What better test for an amplifier than the Clipperton DXpedition? Even after 32,000 QSO's, and an accidental dunk in the ocean, the same 3 MLA-2500's are still amplifying other rare DXpeditions around the world - listen for them.

Convinced? Isn't it time you owned the amplifier that powered Clipperton and thousands upon thousands of radio stations throughout the world?

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TS-820S

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You can always tell who's running a TS-820S. Its superb quality stands out from all the other rigs on the band... and when the QRM gets heavy, the TS-820S's adjustable RF speech processor, utilizing a 455-kHz circuit to provide quick-time-constant compression, will get the message through. RF negative feedback is applied from the final to the driver to improve linearity, and third-order products are at least -35 dB. Harmonic spurious emissions are less than -40 dB and other spurs are less than -60 dB. RF input power is 200 W PEP on SSB, 160 W DC on CW, and 100 W DC on FSK. Receiver sensitivity is better than 0.25 μ V for 10 dB S/N. The TS-820S is known for its superb receiver selectivity, and its famous IF shift easily eliminates heavy QRM. That's why the TS-820S is the DXer's choice.

See your local Authorized Kenwood Dealer today.



TS-820S VFO-820S SP-820

Kenwood's unbeatable combination. The VFO-820 solid state remote VFO adds greatly to the versatility of your TS-820S. It has its own RIT circuit and control switch and is a perfectly matched accessory. The SP-820 deluxe external matching speaker includes audio filters for added versatility on receive and two audio inputs.



The TS-520S... the most popular Amateur Radio transceiver in the world... provides a foundation for an expanding series of accessories designed to please any ham... from Novice to Amateur Extra.

TS-520S

The TS-520S transceiver provides full transmit and receive coverage of all Amateur bands from 160 through 10 meters. It also receives 15.0 (WWV) to 15.5 MHz and another 500-kHz range of your choice in the auxiliary band position. With the optional DG-5, you have a large digital frequency readout when transmitting and receiving, and the DG-5 also doubles as a 40-MHz frequency counter. The TS-520S includes a built-in AC power supply, and, with the addition of the optional DS-1A DC-DC converter, it can function as a mobile rig. It features a very effective noise blanker, RIT, eight-pole crystal filter, 25-kHz calibrator, front-panel carrier level control, semi-break-in CW with side-tone, built-in speaker, heater switch, 20-dB RF attenuator and easy phone-patch connection. RF input power is 200 W PEP on SSB and 160 W DC on CW. Carrier suppression is better than -40 dB and sideband suppression is better than -50 dB. Spurious radiation is less than -40 dB. Receiver sensitivity is 0.25 μ V for 10 dB (S+N)/N. Selectivity is 2.4 kHz at -6 dB/4.4 kHz at -60 dB and, with the optional CW-520 CW filter, is 0.5 kHz at -6 dB/1.5 kHz at -60 dB.

See your local Authorized Kenwood Dealer for more information, and a super deal!



A great station... at an affordable price! The TS-520S with its companion accessories... including two new units. The AT-200 antenna tuner provides a versatile tool in any station. The other is the TV-520S, Kenwood's 2 meter transverter for SSB and CW operation from 146 to 148 MHz.

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...pacesetter in amateur radio

The 599D "Twins" are offered to the discriminating Amateur who appreciates the advantages of operating a separate transmitter and receiver.

R-599D/T-599D

The R-599D receiver and T-599D transmitter provide greater flexibility with more features than found in a transceiver.

The R-599D receiver is all solid-state, covering all Amateur bands from 160 through all of 10 meters, as well as auxiliary band and WWV (10 MHz). With optional converters it also receives 6 meters and 2 meters. Modes include LSB, USB, CW, AM, and FM. A 2.2-kHz eight-pole filter is built-in for SSB, as well as a 500-Hz eight-pole CW filter and a 5.0-kHz six-pole AM filter. An optional 14.0-kHz six-pole FM filter is available. Also featured are an AGC control (slow/fast/off), 25-kHz calibrator, RIT, noise blanker, ANL (AM), squelch, monitor, VFO selector, and RF gain control which does not affect S-meter reading.

The T-599D transmitter is solid-state except for the driver and final tubes. It covers the 80 through 10-meter Amateur bands, on LSB, USB, CW, and AM. An AC power supply is built-in. Also included are VOX, anti-VOX, PTT, semi-break-in CW with sidetone, ALC, transmitter terminal.

Enjoy split frequency control in four separate/transceiver combinations with the 599D "Twins". See your local Authorized Kenwood Dealer for more information.



R-300

The R-300 all-band communications receiver covers the following ranges: (A) 170-410 kHz; (B) 525-1,250 kHz; (C) 1.25-3.0 MHz; (D) 3.0-7.5 MHz; (E) 7.5-18.0 MHz; and (F) 18.0-30.0 MHz. It receives AM, SSB, and CW. The receiver features large, easy-to-read drum dials. Bandspread is calibrated for 10 foreign-broadcast shortwave bands, and a replacement bandspread calibration is available for the 80-10-meter Amateur bands. Included is a three-way power supply (AC/batteries/external DC). Wide and narrow ceramic filters are employed for high selectivity. Also included is a 500-kHz calibrator.

IT'S NEW... IT'S UNIQUE... AND IT'S TRULY USEFUL. IT'S KENWOOD'S SM-220 STATION MONITOR. THE SM-220'S UNEXCELLED VERSATILITY ALLOWS YOU TO MONITOR YOUR TRANSMISSIONS, MONITOR INCOMING SIGNALS, AND MONITOR THE AMOUNT AND STRENGTH OF BAND ACTIVITY* AND PERFORMS AS A GENERAL-PURPOSE 10 MHz OSCILLOSCOPE, AS WELL.

Kenwood offers this totally unique unit as a perfect compliment to your TS-820S or TS-520S station.** The SM-220, based on a wideband oscilloscope (2 Hz to 10 MHz), permits you to monitor your transmitted signals, thus assuring optimum linearity and maximum performance. With the addition of the BS-5 or BS-8 Pan Display option you will be able to determine visually the location and strength of adjacent signals without tuning your receiver off frequency. The choice of options allows you to adapt the SM-220 to either the TS-820 series or TS-520 series.

The SM-220 has a built-in two-tone audio generator with full provisions for tuning your exciter and linear amplifier (160 m through 2 m).

All this costs little more than a general-purpose oscilloscope. And, of course, it's pure Kenwood quality.

*With BS-5 or BS-8 option

**For other models check with appropriate manufacturer for compatibility.

SM-220



Function: Selects operation mode. OSC/RTTY: General testing of station equipment, experimental design of new equipment, or troubleshooting; display of receiver IF output allows you to give "signal quality reports".

Power ON indicator: Power switch.

Intensity: Controls brightness of scope display.

Band Scope: (Pan Display) With BS-5 or BS-8 option, allows you to "see" the signals on both sides of your operating frequency without tuning your receiver off frequency. Useful for determining "band conditions" - band crowding, source of interference from adjacent stations... a visual display of what you would hear if you tuned across the band, without having to touch your receiver's dial.

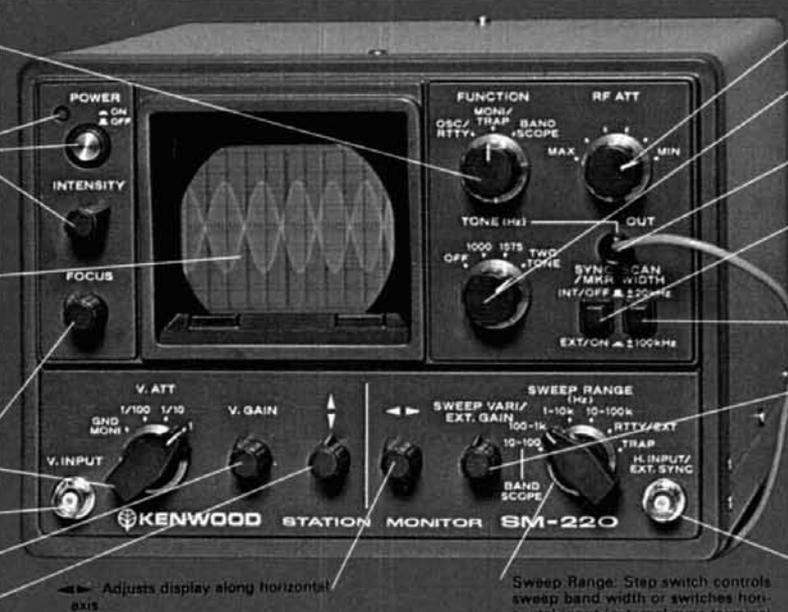
Focus: Controls sharpness of scope display.

Vertical Attenuator: Precision step attenuator (gain control) switch adjusts vertical input level.

Vertical Input: Accepts IF input, RTTY input or oscilloscope input.

Vertical Gain: Potentiometer to linearly adjust vertical input level.

Adjusts display along vertical axis.



RF Attenuator: Level control used in MONI/TRAP mode.

Tone: Step switch selects Wien bridge tone generators: 1000 Hz, 1575 Hz or both tones simultaneously.

Out: Output of the audio generator can be connected to the transceiver's microphone input for "two-tone test". Also for trapezoidal test of transceiver linear amplifier.

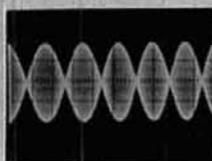
Synchronization Marker: Selects internal or external sync (similar to horizontal hold on TV). Turns On or Off the built-in marker which shows operator where his receiver is actually tuned.

Scan Width: Selects width of "window" or receive band display when using the Pan Display option. (100 kHz or 20 kHz).

Variable sweep control/External gain: Controls (1) sweep speed of display in any sweep range, (2) optional Pan Display (Band Scope) speed of display, (3) level of horizontal input/external synchronization input when sweep range is in RTTY Ext or Trap.

Horizontal Input/External Sync: Accepts either (1) RTTY input for tuning, (2) external sync input for test (oscilloscope functions), (3) external oscillator for Lissajous display.

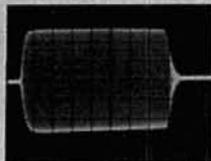
Sweep Range: Step switch controls sweep band width or switches horizontal input/external sync terminal "ON".



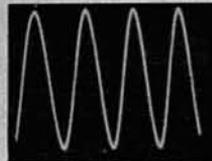
Two-Tone Wave Envelope For "performance" tune-ups or checking proper transceiver operation.



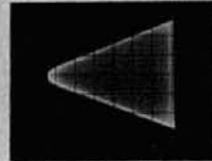
Pan Display Use to check source of interference during "QSO" without moving off-frequency. Also determines location and strength of adjacent frequencies. (Requires BS-5 or BS-8 option)



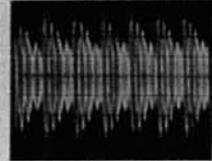
Keyed Waveform Shows detail of CW keying. Use to monitor the quality of your CW note. (Photo shows ideal waveform produced by TS-820S.)



Oscilloscope Operation (1 kHz) Oscillator function allows Sine, square wave, Lissajous patterns for testing or design work.



Trapezoid (TS-820S w/ TL-922) Shows linearity of power amplifier. Used primarily for testing.



Wave Envelope shows full SSB voice modulation, with processor on (full compression) and "clean signal" at full power.

 **KENWOOD**
...pacesetter in amateur radio

STILL THE SAME FINE, TIME PROVEN RIG. BUT NOW WITH THE SIMPLE ADDITION OF A PLUG-IN CRYSTAL, THE TS-700SP WILL BE ABLE TO UTILIZE THE NEW REPEATER SUB-BAND (144.5 to 145.5 MHz) STILL FEATURES ALL OF THE FINE ATTRIBUTES OF THE TS-700S: A DIGITAL FREQUENCY DISPLAY, RECEIVER PRE-AMP, VOX, SEMI-BREAK IN, AND CW SIDETONE. OF COURSE, IT'S ALL MODE, 144-148 MHz, VFO CONTROLLED ... AND KENWOOD QUALITY THROUGHOUT.

TS-700SP



TS-600

The luxury all-mode transceiver for 6 meters. All solid-state. SSB, FM, AM, and CW.

It's easy to work VHF DX on 6 meters with the TS-600 all-mode transceiver. The 10-watt, solid-state rig covers 50-54 MHz, with built-in VFO and 20 fixed channels. The main tuning dial is calibrated every 1 kHz for precise tuning. The built-in AC/DC power supply allows base and mobile operation. Other features include a noise-blanker circuit and RIT (receiver incremental tuning).

Features: 4 MHz band coverage (144 to 148 MHz) • Automatic repeater offset capability on all FCC authorized repeater subbands including 144.5 - 145.5 MHz • Simply dial receive frequency and radio does the rest... simplex, repeater, or reverse. Same features on any of 11 crystal positions • Transmit/Receive capability on 44 channels with 11 crystals • Operates all modes: SSB (upper and lower), FM, AM and CW • Digital readout with "Kenwood Blue" digits • Receiver pre-amp • Built-in VOX • Semi break-in on CW • CW sidetone • All solid-state • AC and DC capability. 10 watts RF output on SSB, FM, CW • 3 watts on AM • 1 watt FM low-power switch • 0.25 μ V for 10 dB (S+N)/N SSB/CW sensitivity • 0.4 μ V for 20 dB quieting FM sensitivity.

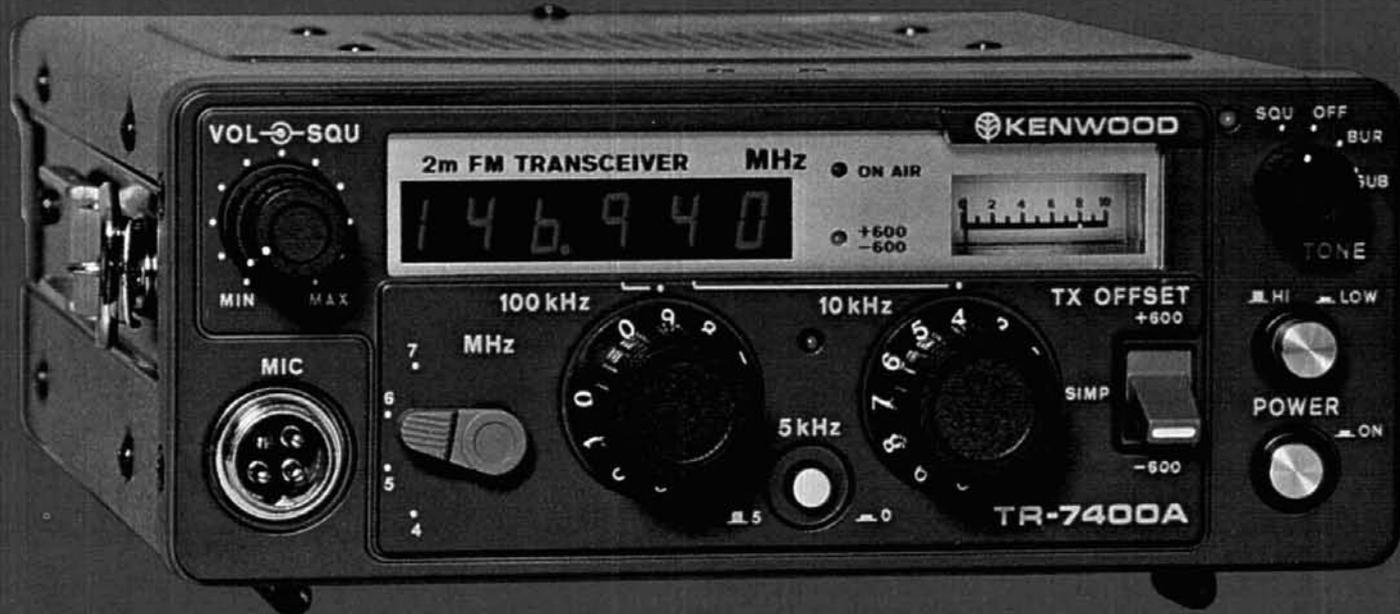
10 watts RF output on SSB, FM, CW • 3 watts on AM • 1 watt FM low-power switch • 0.25 μ V for 10 dB (S+N)/N SSB/CW sensitivity • 0.4 μ V for 20 dB quieting FM sensitivity.

The fully-synthesized TR-7400A 2-meter FM transceiver operates on 800 channels and features repeater offset over the entire 144-148-MHz range, dual frequency readout, six-digit display, and subaudible tone encoder and decoder. RF output is at least 25 watts!

The TR-7400A 2-meter FM transceiver provides fully synthesized operation, including 600-kHz repeater offsets, over the entire 144-148-MHz range. It can operate on any of 800 channels, spaced 5 kHz apart. RF output is at least 25 W, and typically 30 W. A low power position produces 5-15 W (adjustable). Included is a dual frequency readout with large six-digit LED display plus a dial readout. The subaudible CTCSS signaling feature may be used on transmit and receive, or transmit only. Optional tone-burst modules are available. Receiver sensitivity is better than 0.4 μ V for 20 dB quieting. Large, high-Q, helical resonators minimize interference from outside the band. A two-pole 10.7-MHz monolithic crystal filter provides excellent selectivity. Optional active filters are available for 15-kHz "split" operation. Intermodulation distortion is down more than 66 dB, spurious rejection is better than -60 dB, and image rejection is better than -70 dB.

See your local Authorized Kenwood Dealer today, for a demonstration of the fantastic TR-7400A.

TR-7400A



TR-8300

FM transceiver for 70-cm Amateur band. 23 crystal-controlled channels (three supplied). Transmitter output is 10 watts.

The TR-8300 450-MHz FM mobile transceiver provides 10 watts output (switchable to 1 watt) on 23 crystal-controlled channels (three pairs of crystals supplied). The transmitter covers 445 to 450 MHz, and the

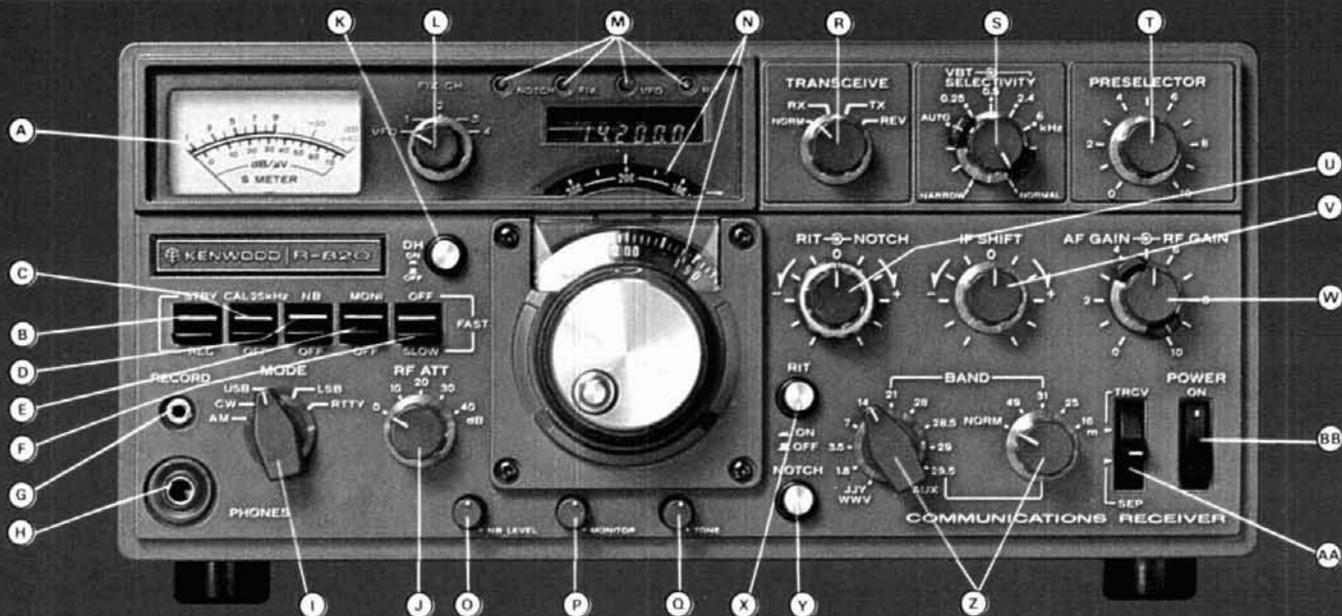
receiver covers 442 to 447 MHz. The receiver includes a five-section helical resonator and a two-pole crystal filter in the IF for improved intermod rejection. Sensitivity is 0.5 μ V for 20-dB quieting. A front-panel switch may be used to activate tone-signaling or other user-provided function. An LED indicates receive crystal functioning. A monitor circuit allows user to listen to his own modulation.

INTRODUCING THE ULTIMATE IN RECEIVER DESIGN ... THE KENWOOD R-820

With more features than ever before available in a ham-band receiver. This triple-conversion (8.33 MHz, 455 kHz, and 50 kHz IFs) receiver, covering all Amateur bands from 160 through 10 meters, as well as several shortwave broadcast bands, features digital as well as analog frequency readouts, notch filter, IF shift, variable bandwidth tuning, sharp IF filters, noise blanker, stepped RF attenuator, 25 kHz calibrator, and many other features, providing more operating conveniences than any other ham-band receiver. The R-820 may be used in conjunction with the Kenwood TS-820 series transceiver, providing full transceive frequency control.

R-820

- A S-METER** Easy-to-read, calibrated to S9 = 40 dB full scale and dB/μV.
- B STANDBY/RECEIVE SWITCH** Disables audio circuits during transmit mode with associated transmitter.
- C CALIBRATOR SWITCH** Built-in crystal calibrator, settable to WWV provides signal every 25 kHz.
- D NOISE-BLANKER SWITCH** A specially designed crystal filter eliminates noise pulses such as ignition-noise interference.
- E MONITOR SWITCH** RF sampling allows user to hear his own voice when using associated transmitter.
- F AGC SWITCH** Automatic gain-control circuit switchable to slow or fast response, or completely off.
- G RECORD JACK** Makes recording off the air simple.
- H HEADPHONE JACK** Provision for plugging in headphones.
- I MODE SWITCH** Selection of AM, CW, upper or lower sideband, or RTTY.
- J RF-ATTENUATOR SWITCH** 10 dB steps of attenuation from 0 to 40 dB, to prevent overloading from nearby stations, and for precise signal comparison.
- K DIGITAL HOLD** Locks counter and display while VFO is tuned to another frequency. Helps return to 'hold' frequency.
- L VFO/CRYSTAL SWITCH** Permits VFO control or crystal control on four selectable frequencies.
- M LED INDICATORS** Light-emitting diodes indicate activation of notch filter, crystal-controlled reception, VFO control, and RIT.
- N DRS DIAL** Satin-smooth VFO tuning dial system provides easy analog frequency readout (useful when digital hold is activated). LSB, USB, and CW frequencies are accurately read from the same pointer.
- O NOISE-BLANKER LEVEL CONTROL** Controls level of blanking, for maximum effect in eliminating noise interference.
- P MONITOR CONTROL** Adjusts level of RF sampling.
- Q TONE CONTROL** Varies audio-output frequency response.
- R TRANSCEIVE SWITCH** Selects frequency tuning from either the receiver or TS-820 series transceiver.
- S VBT/SELECTIVITY CONTROLS** Separate controls on the same shaft provide variable bandwidth tuning as well as selection of four IF filters: 250 Hz*, 500 Hz*, 2.4 kHz, and 6 kHz* (*optional). CW filters function in 455-kHz IF for superior shape factor.
- T PRESELECTOR** Peaks tuned circuits in RF amplifier stage for increased selectivity and sensitivity. RF amplifier coil is dual-tuned.
- U RIT/NOTCH CONTROLS** RIT allows receiver to be tuned off frequency, while not affecting transmit frequency, when in transceive mode with TS-820. Notch control tunes notch within IF passband for eliminating interference. Notch frequency remains the same even when IF shift is utilized.



- V IF SHIFT** Varies (shifts) IF passband away from interfering signal.
- W AF GAIN/RF GAIN** Separate controls adjust volume and RF gain.
- X RIT SWITCH** Allows tuning off frequency with RIT control, and return immediately to VFO frequency by pushing switch.
- Y NOTCH SWITCH** Takes variable notch filter in and out of circuit.
- Z BAND SWITCHES** Selects frequency bands from 15 MHz (WWV), 160 through 10 meters, the 49, 31, 25, and 16-meter shortwave broadcast bands, and an auxiliary band.
- AA TRANSCEIVE/SEPARATE SWITCH** Enables receiver VFO to control the receiver and TS-820 (or TS-820S) frequency (or the TS-820 VFO to control both), or both can function independently.
- BB POWER SWITCH** Turns receiver on and off.

R-820 PERFORMANCE SPECIFICATIONS

Frequency Range:

160 meters (1.8-2.0 MHz)
80 meters (3.5-4.0 MHz)
40 meters (7.0-7.5 MHz)
20 meters (14.0-14.5 MHz)
15 meters (21.0-21.5 MHz)
10 meters (28.0-28.5 MHz)
10 meters (28.5-29.0 MHz)
10 meters (29.0-29.5 MHz)
10 meters (29.5-30.0 MHz)
19 meters (15.0 (WWV)-15.5 MHz)
49 meters (5.9-6.4 MHz)
31 meters (9.4-9.9 MHz)
25 meters (11.5-12.0 MHz)
16 meters (17.7-18.2 MHz)
Auxiliary band

- Modes: AM, CW, USB, LSB, RTTY
- Sensitivity: 160-10 m, 19 m, SSB, 0.25 μV at 10 dB S+N/N
 AM, 1.5 μV at 10 dB S+N/N
 49, 31, 25, 16 m, SSB, 0.5 μV at 10 dB S+N/N
 AM, 3.0 μV at 10 dB S+N/N
- Selectivity: CW (with optional 250-Hz filter) 250 Hz (-6 dB), 500 Hz (-6 dB)
 CW (with optional 500-Hz filter) 500 Hz (-6 dB), 850 Hz (-6 dB)
 SSB (2.4-kHz filter) 2.4 kHz (-6 dB), 4.4 kHz (-6 dB)
 AM (6-kHz filter) 6 kHz (-6 dB), 12 kHz (-6 dB)
- Image Ratio: 160-10 m, 19 m, 80 dB
 49, 31, 25, 16 m, 60 dB
- IF Rejection: 160-10 m, 19 m, 90 dB
 49, 31, 25, 16 m, 50 dB
- Power Requirements: 100/120/220/240 VAC, 50/60 Hz, or 12-15 VDC
- Dimensions: 13-1/8" (333 mm)W x 6" (153 mm)H x 13-3/16" (335 mm)D
- Weight: 26.4 lbs (12 kg)



AT-200

The AT-200 is an antenna tuner, but it's also much more. It's an antenna switch, an SWR bridge and an in-line wattmeter. The AT-200 reduces the clutter and increases the operating efficiency of your station... and at a surprisingly moderate price.

The AT-200 features a seven position rotary switch that selects 1 of 3 antennas and connects it through the antenna tuner circuit or directly to the transceiver. The 7th position allows you to connect a dummy load directly to your transceiver for tune up and testing. Two of the antenna inputs are fitted with SO-239 type coax connectors. A third input allows for easy hook up of a wire antenna with an impedance of 10 to 500 ohms. The AT-200 may be used on all HF amateur bands from 160 to 10

meters. It's handsomely styled to match the TS-820S and TS-520S Series (and TS-820 and TS-520), but can also be used with any HF transceiver or transmitter with less than 200 watts output.

Frequency coverage: Amateur bands 1.8 to 30 MHz • Input impedance: 10 to 500 ohms • Maximum power capability: 200 watts • Insertion loss: 0.5 dB • Power meter: 20 watt/200 watt full scale • SWR meter measures up to 10:1 • Dimensions: 6-1/2" W x 7-3/8" D x 6-9/16" H • Weight: 6.2 lbs.



TV-506

The TV-506 6-meter transverter produces 10 watts on SSB and CW. Simply plug it into your TS-520 or TS-820 Series transceiver. It works with most other transceivers, too.

TV-502S

The TV-502S 2-meter transverter produces 8 watts on SSB and CW. It easily hooks up to the TS-520 and TS-820 Series transceivers.



HS-4

The HS-4 headset is comfortably padded and is completely adjustable, for extended periods of wear. Frequency response is designed for Amateur communications (300 to 3000Hz). Impedance is 8 ohms.

MC-50

The MC-50 dynamic microphone is perfect for any ham shack, and is ideal for all Kenwood equipment as well as many other brands. It includes PTT and LOCK switches, as well as a microphone plug wired for instant connection to any Kenwood rig. It is easily converted to high or low impedance (600 Ω or 50 Ω).

MC-30S & 35S

The MC-30S and MC-35S dynamic mobile microphones provide 150-5000 Hz frequency response (150-4000 Hz when operated as noise-cancelling microphones). The MC-30S impedance is 500 Ω and the MC-35S is 50 Ω .

OTHER KENWOOD PRODUCTS

ACCESSORIES FOR TS-820 Series 160-10-m transceiver
 DG-1 digital frequency display
 VFO-820 deluxe remote VFO
 CW-820 500-Hz CW filter
 DS-1A DC-DC converter
 SP-820 external speaker with audio filters

ACCESSORIES FOR TS-520 Series 160-10-m transceiver

DG-5 digital frequency display
 DK-520 digital adaptor kit for TS-520
 VFO-520 remote VFO
 SP-520 external speaker
 CW-520 500-Hz CW filter

ACCESSORIES FOR 599D Series 160-10-m transmitter and receiver

S-599 external speaker
 CC-29A 2-m converter
 CC-69 6-m converter
 FM-599A FM filter

ACCESSORIES FOR TS-700SP 2-m all-mode transceiver

VFO-700S remote VFO
 SP-70 matching speaker
Other products:
 PS-6 power supply for TR-7500 and TR-8300
 PS-8 power supply for TR-7400A
 VOX-3 VOX for TS-600/TS-700A
 Active filter elements for TR-7400A



TRIO-KENWOOD COMMUNICATIONS INC.
 1111 WEST WALNUT/COMPTON, CA 90220



general-purpose vhf receiver

Design details
of a
receiver
that covers
the popular
vhf ranges,
in one
convenient package

Any uhf enthusiast can appreciate a receiver that monitors all vhf frequencies and modes in one small, convenient package. Alas, this sort of receiver doesn't exist in the amateur marketplace. For years at my station, a Collins 75A2 — supported by a bewildering array of converters — did the job. Soon after my station was remodeled, I developed a strong desire to replace the large, unwieldy (and ugly) receiver rack with smaller and modern equipment. I'm an avid homebrewer always looking for new projects to occupy limited time and pocket money, so plans for a new receiver were soon germinating.

design features

Hf operators and shortwave listeners alike have always enjoyed the convenience of general-coverage receivers, so why not something similar in nature, only intended for the vhf regions and tailored to today's needs for diversified vhf operation? Doug DeMaw¹ was on this track some years back when he described a tunable i-f receiver for use with converters. While its abilities fell short of my receiving requirements, several weeks of daydreaming produced on paper a receiver better able to meet my goals, which would incorporate the following features:

By Peter J. Bertini, K1ZJH, 20 Patsun Road,
Somers, Connecticut 06071

1. Four-MHz coverage, through a 26-30 MHz i-f range, to allow full reception of the 6-, 2- and 1-1/4-meter bands without changing converter crystals. Dial readout was desired to at least 1-kHz resolution with mechanical and electrical stability for smooth CW and ssb reception.

2. Multimode detection for a-m, ssb, and fm with squelch to allow monitoring all the popular vhf modes.

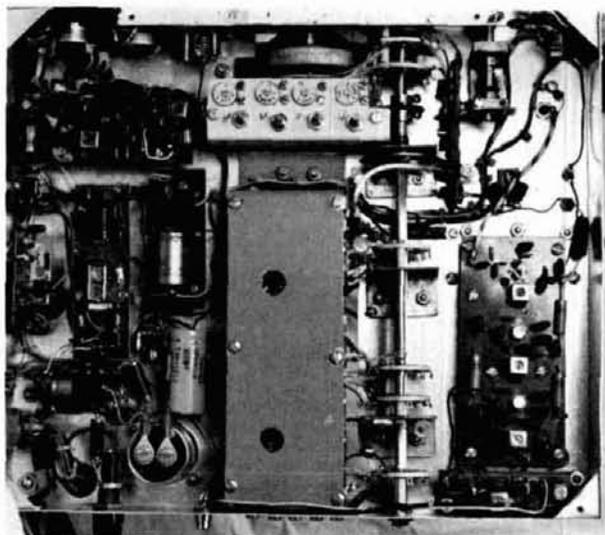
3. Several selectivity positions for mode compatibility and operating convenience.

4. All components self-contained in one neat package including all converters, power supplies, and speaker.

A pretty tall order to fill, and obviously some compromises must be reached. Adequate coverage of 4 MHz was best done in four 1-MHz segments, starting at 26 MHz and ending at 30 MHz. This allows for quick scanning across a band while maintaining a tuning rate comfortable enough for ssb reception. The range of 26-30 MHz was chosen for the i-f because many converters come equipped for this range and the i-f is also high enough for good converter image-rejection.

Performance data for the basic receiver, covering the 26-30 MHz range, is presented in **table 1**. I'd like to point out that no pretense is made of using this receiver as the nucleus for a moon-bounce, scatter, or other demanding station-receiver role. Those so inclined will do better with a special-purpose receiver. Templates or board layouts are not available, and this receiver is not intended as a beginner's project.

Schematics of the vhf receiver are shown in **figs. 1** through **10**. The basic receiver, not considering the vhf converters, is a dual-conversion design using the



Bottom view of the author's receiver. The two boards in the bottom right are the i-f amplifier and the agc detector/amplifier. In the upper left is the bfo/product detector board. Mounted on the left side of the main chassis are the a-m detector and squelch boards. Below and to the right of the product detector is the audio amplifier board. To the left of the audio board is the fm limiter and detector.

standard frequencies of 10.7 MHz for the first i-f and 455 kHz for the second i-f. Motorola MFE 121 dual-gate mosfets were used in the 26-30 MHz rf preselector and in the first- and second-mixer stages. No peaking of the preselector is required across the range on any of the 1-MHz receiver bands. The vfo bandswitch also selects a set of preselector trimmers for each of the four bands; stagger tuning provides broadbanding and uniform gain over each 1-MHz segment.

The basic receiver input allows for direct monitoring of frequencies between 26-30 MHz should the 10- or 11-meter bands be of interest. Special attention to

table 1. Performance data for the basic receiver.

frequency coverage	26-30 MHz in four 1-MHz bands
circuit	superhet, dual conversion; 10.7 MHz first i-f; 455 kHz, second i-f
sensitivity	0.12 μ V detectable in a-m/fm mode; 0.1 μ V detectable in ssb mode
noise figure	not measured — estimated at \approx 2 dB
stability	after 1-hour warmup in stable atmosphere, less than 500 Hz per hour
spurious responses	all greater than 50 dB down
i-f rejection	80 dB down
dial accuracy	1 kHz digital resolution. Dial mechanical backlash less than 200 Hz
IMD performance	two 1-mV signals separated 20 kHz required to produce a third order product equivalent to 1.5 μ V
agc range	agc action begins at 0.3 μ V; i-f distortion at 15 mV
selectivity	2 kHz, 4 kHz, 8 kHz, 16 kHz (13 kHz actual) 455 kHz filters. 10.7-MHz IMD filter is 13 kHz
modes	fm, a-m, ssb and CW detectors
squelch	noise operated, all modes f-m generator and Hewlett-Packard 608D a-m generator used for performance analysis
vhf ranges	2-meter, 6-meter, 1-1/4-meter, 430-434 MHz, 446-450 MHz coverage inboard; three external converter provisions

adequate shielding and power-line bypassing is encouraged. The recent proliferation of 27-MHz CB units increases the likelihood of annoying i-f breakthrough from strong signals in the 27-MHz range. Converters feeding the receiver should be of low or near unity gain to preserve receiver dynamic range. Modern designs without rf amplifiers, especially those employing hot-carrier diodes in double-balanced mixers or mosfet mixer circuits, are ideal. Of course, you can use your own converters; but converters with excessive gain should be followed by an appropriate T-pad attenuator to prevent receiver overload.

An alternative to the i-f attenuator pads to bring the converters to or near unity gain was suggested by *Hamtronics*. The *Hamtronics* C25-series converters produce between 10-20 dB gain, depending upon the band and device alignment.

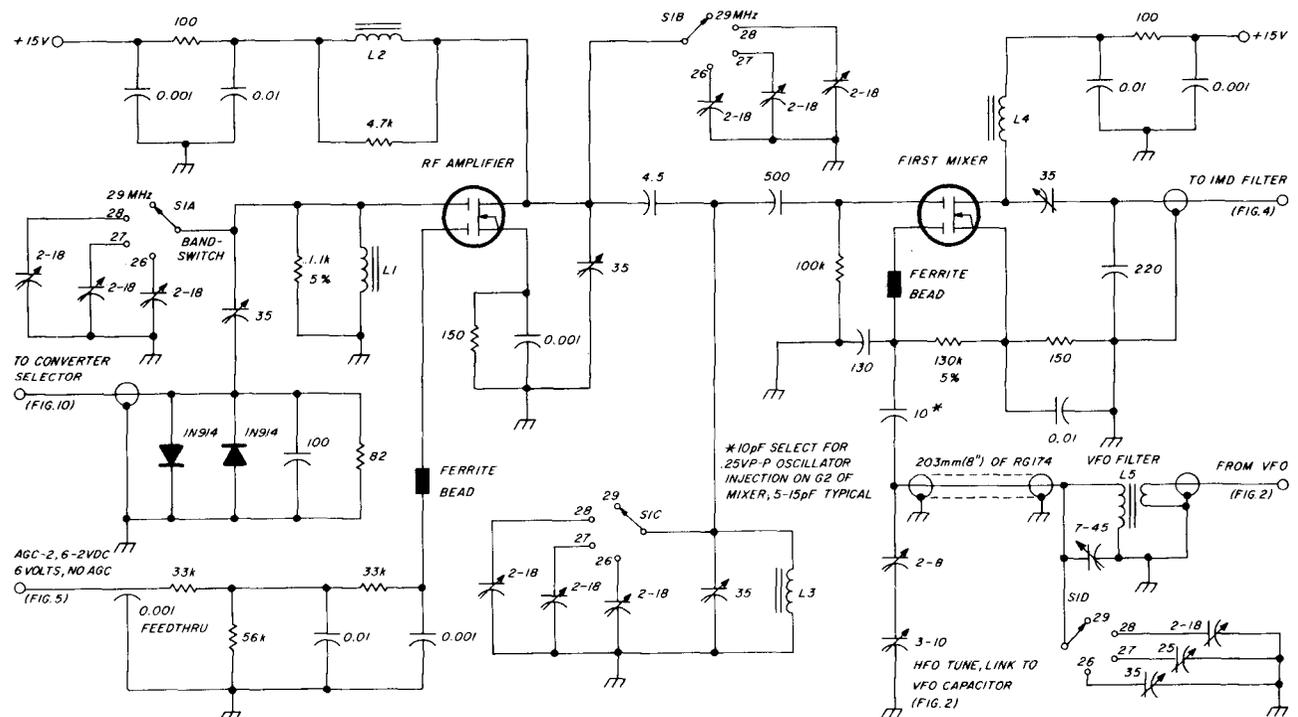
The cascade front-end stage in the C25 converters is broadband; slight stagger tuning of these stages yields the desired 4-MHz bandwidth. However, the i-f output transformer at 28 MHz has a comparatively narrow passband. Resistive loading of the i-f transformer primary broadens the i-f passband while also

decreasing converter gain, which eliminates the need for external attenuators. The approach used on the C25 converters should be adaptable to other makes of converter that exhibit a restricted i-f passband and excessive gain.

The *Hamtronics*-series converters designed by Jerry Vogt, WA2GCF, were used for the vhf converter front ends of this receiver. Kits are available at modest cost. Three *Hamtronics* converters cover the three vhf bands: a P25-50 for 6 meters, a P25-150 for 2 meters, and a P25-220 for 1-1/4 meters.

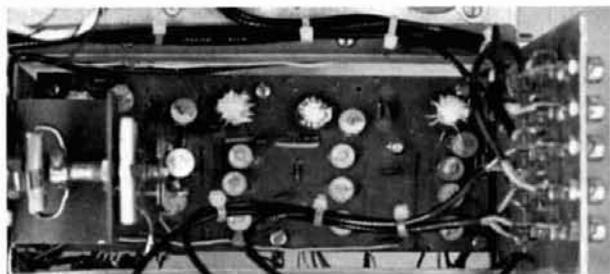
Since the 3/4-meter band is 30 MHz wide, two uhf converters were needed to monitor this band adequately. One is for the 430-434-MHz DX segment; the other is for the 446-450-MHz range, covering the 400-MHz fm repeater output channels for the northeast corridor of the U.S. These are unity-gain converters, and, when used without an external rf amplifier, don't require the T-pad i-f attenuators.

Receiver use is not limited to amateur frequencies. Suitable converters provide many enjoyable hours monitoring commercial and military air traffic, police, radio-telephone, weather bulletins, municipal and federal government, and much other interesting vhf



- | | | | |
|---|-------------------------------|-------------------------------|--|
| L1 - L5 wound on 13 mm (0.5 in.) OD type E toroid cores | L4 | 30 turns 0.5 mm (no. 24) wire | |
| L1 | 12 turns 0.5 mm (no. 24) wire | L5 | 16 turns 0.8 mm (no. 22) wire;
link to vfo is 1 turn 0.8 mm (no. 22) wire |
| L2 | 12 turns 0.5 mm (no. 24) wire | S1 | 6-pole, 4-position bandswitch (see text) |
| L3 | 12 turns 0.5 mm (no. 24) wire | | |

fig. 1. Rf amplifier, first mixer, and vfo filter. The fets are MPF121, SK3050, or equivalent.



Top view of the receiver showing the rf amplifier and first mixer.

activity. Even the hf frequencies can be up-converted, as DeMaw did in his "Receiving Package," reference 1, to produce a truly all-band receiver.

mixers and filter arrangements

Vfo injection from 15.3-19.3 MHz is supplied to gate 2 of the first mixer, (an MFE/MPF121). A Piezo Technology Model 1433 crystal filter, which has 13-kHz bandwidth with a 10.7-MHz center frequency, follows the mixer and acts as an IMD filter, which protects the second mixer from strong out-of-pass-band signals. The 13-kHz bandwidth of this filter sets the maximum receiver selectivity. (It's electrically similar to the KVG XF9A filter.)

A 10.245-MHz crystal-oscillator signal, mixing with the 10.7-MHz i-f signals in the second mixer stage (fig. 4), produces the 455-kHz i-f. Four 455-kHz Collins mechanical filters follow, which select the desired 455-kHz i-f bandwidth. Selectivity positions of 16, 8, 4 and 2 kHz are provided by the four filters.

The use of so many expensive mechanical filters may appear extravagant, but they permit versatility. The 4-, 8-, and 16-kHz filters were salvaged from a demolished R390A i-f strip purchased at a hamfest for \$5.00. The 2.1-kHz filter was purchased at another for only \$18.00. A 2- or 3-kHz filter will serve the majority of ssb and CW vhf requirements, and a simple LC bandpass filter, made up from i-f transformers, will be adequate for fm or a-m reception if inexpensive mechanical filters aren't readily available. Note that in this receiver, the 16-kHz filter pass-band is limited to 13 kHz by the selectivity of the Piezo Technology filter. A 20-kHz, 10.7-MHz filter would have improved this situation, but I used materials on hand. The additional cost of a new filter was not justified.

The vfo (fig. 2) operates in the 15.3-19.3-MHz region in four bandswitched 1-MHz segments. Mechanical rigidity and electrical stability are paramount watchwords for a vfo working this high in frequency. Careful mounting of all vfo components and elimination of chassis flexing and dial backlash are

important for good vfo performance. The chassis was rigidly reinforced. The Eddystone dial assembly serves admirably.

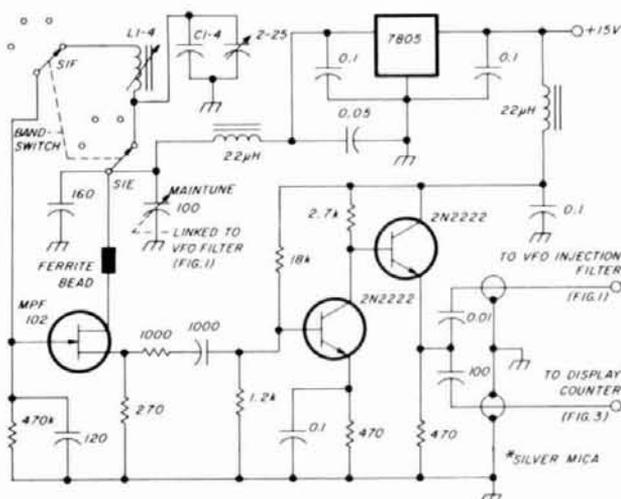
The use of polystyrene caps and good-quality ceramic coil forms and trimmers help contribute to vfo stability. The vfo, built on 3-mm-thick (1/8 in.) glass-epoxy board, was mounted beneath the receiver away from heat-producing components and drafts. The jfet oscillator is powered by a dedicated 5-volt regulator; the low voltage was helpful in reducing drift from rf component heating. The end result is a vfo exhibiting freedom from microphonics and drift, which permits extending monitoring periods without frequent and annoying retuning.

Extensive filtering of the vfo second harmonic was necessary after a problem surfaced during monitoring of 29.6 MHz. Instead of amateur signals, several local CB operators were heard. When monitoring 29.6 MHz, the vfo second harmonic is at 37.8 MHz. When mixed with the 27.1-MHz CB signals spurious responses were produced at 10.7 MHz, the first i-f!

frequency counter

Despite the excellent performance of the Eddystone 898 dial, the 1-MHz spread didn't permit the desired 1-kHz dial resolution, partially because of the physical limitations involved and also because of small inconsistencies in linearity between band segments caused by the bandswitched vfo circuit.

An ideal solution would have been a counter com-



- C1 43 pF NPO (15.3/16.3 MHz vfo; 26/27 MHz receiver)
- C2 27 pF NPO (16.3/17.3 MHz vfo; 27/28 MHz receiver)
- C3 27 pF NPO (17.3/18.3 MHz vfo; 28/29 MHz receiver)
- C4 not used (18.3/19.3 MHz vfo; 29/30 MHz receiver)
- L1 - L4 9.5 mm (3/8 in.) OD core ceramic. 7 turns 0.6 mm (no. 22) enamel wire

fig. 2. Vfo and buffer amplifiers.

putting the vfo, bfo, hfo, and vhf converter oscillator frequencies to give an exact frequency readout. The cost and complexity of such a counter, and the likelihood of generating spurious signals from the counter circuits, quickly ruled it out. The decision was made to use a counter, but to count and display only the vfo frequency (fig. 3). Up to the tens of kHz position, there is a direct correlation between the vfo and operating frequency, so a two-digit display supplies a direct readout of the tens of kHz and the receiver operating frequency (in kHz). Above 10 kHz, the Eddy-stone dial-calibration points and bandswitch position supply the hundreds of kHz and MHz readings. It's easy to include a third display for hundreds of hertz, but remember that, unless the other conversion oscillators are extremely accurate and set on frequency, the cumulative error makes this resolution

meaningless. Of course, even a 1-kHz readout requires careful frequency setting and regulated power supplies.

The basis for the counter was a circuit in the January 1976 issue of *ham radio*.² Its simplicity, small size, and low cost made it appealing for this application. It's built on a small 102 x 102 mm (4 x 4 in.) square of glass epoxy vectorboard and is sandwiched between two aluminum plates that provide shielding and a ground plane for the counter. No birdies from the counter were heard in the finished receiver.

I used an MD-640 incandescent 7-segment display in place of LED displays. The MD-640s are brighter, cheaper, and don't require current-limiting resistors for each segment as in the case of their solid-state counterparts. The display was also more uniform than that produced by most bargain-basement LED

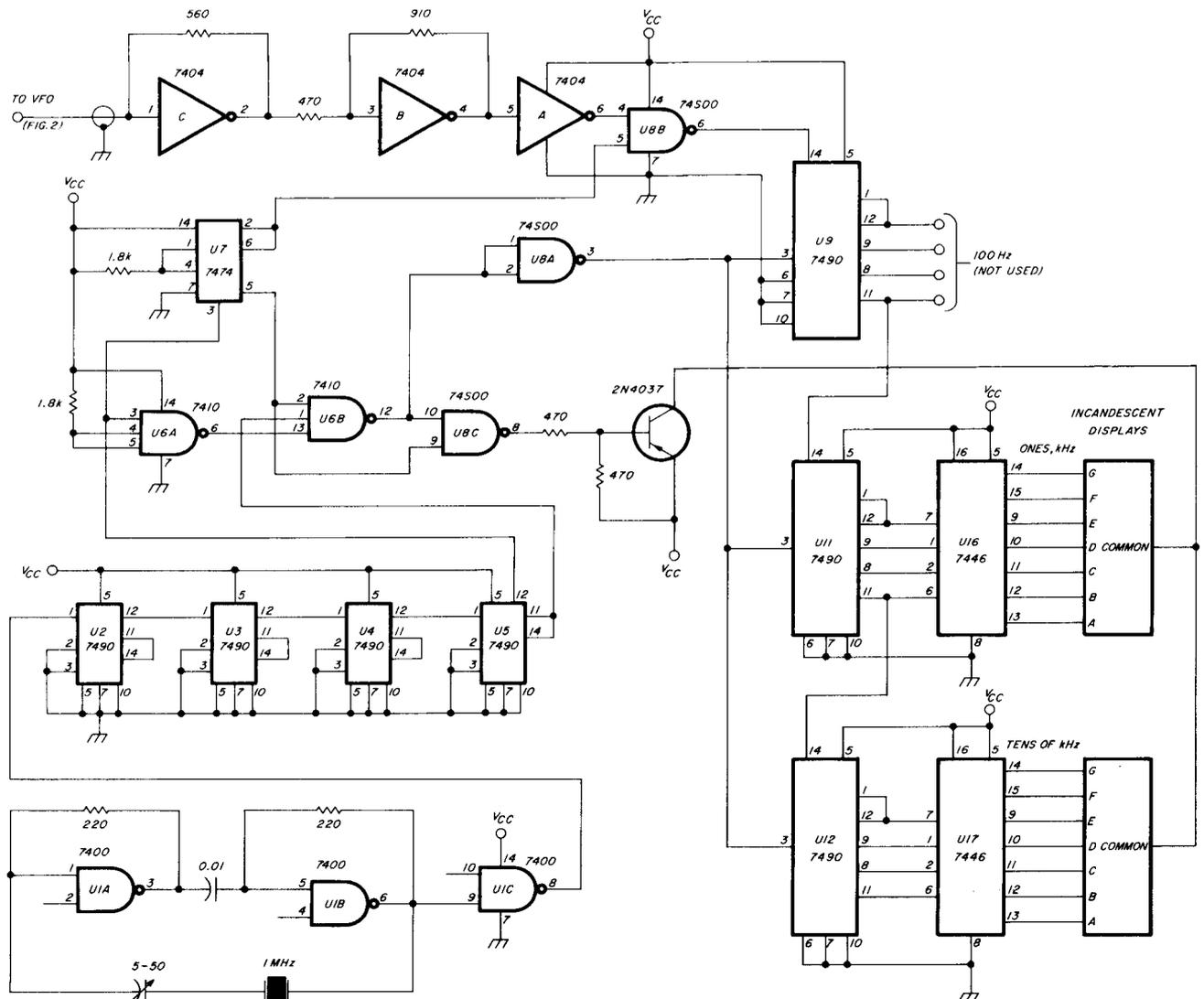
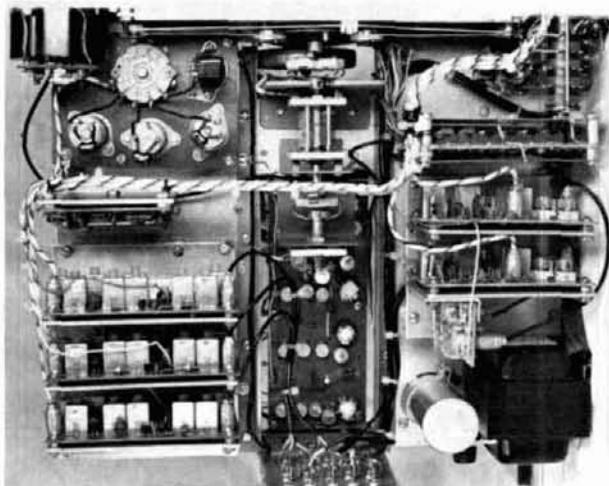


fig. 3. Frequency display. IC numbers refer to the article in the January, 1976, issue of *ham radio* (reference 2).



Overall top view of the receiver. The i-f filters are mounted in the upper left. The ICs for the counter are mounted in the upper right. Note the shielding on both sides of the counter board.

displays. A window for the readouts was carefully milled through the steel Eddystone dial plate. A piece of gray plexiglass behind the window provides contrast for the display readouts.

i-f strip, agc detector, and amplifier

The output of the selected 455-kHz filter goes to our first 455-kHz stage, an MPF102 fet (fig. 5). This stage provides some gain and a good termination for the relatively high filter impedance, but primarily it's an agc attenuator.

The next two stages of i-f amplification employ two high-gain MC1550 ICs. The i-f interstage transformers, from the Radio Shack general replacement line, are resistive loaded rather heavily for stability and to ensure the i-f amplifier is broadband enough not to restrict the broader selectivity positions. (While aligning the i-f amplifiers, it may be necessary to remove the loading resistors to see the peak at resonance.) The last i-f amplifier is a single 2N3904 transistor stage. It is not under agc control and is designed to have sufficient output to drive the agc detector and demodulators.

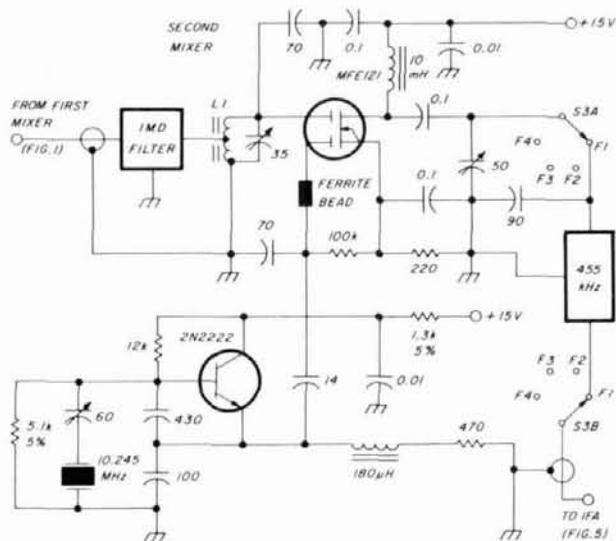
A pair of germanium diodes in a voltage doubler rectifies the i-f signal and presents a proportional dc level to a 2N3053 dc amplifier. The 2N3053 provides an increasing voltage potential during periods of agc action for reducing the MC1550 i-f stage gain and conversely, a decreasing dc potential for gain reduction in the MPF102 i-f amplifier and MFE121 preselector.

A selectable RC time constant controls the agc response; for simplicity the mode switch selects the agc time constant appropriate for the mode selected.

Agc voltage is also used to provide the signal strength meter reference voltage. If the gains of all the converters are equalized, the meter may be calibrated in microvolts instead of just providing a relative signal-strength indication. My converter selector switch also provides agc voltage to the converters, but external agc was not advised for use with the Hamtronics converters.

Ssb or CW reception is accomplished with a hot-carrier-diode product detector circuit inspired by another article.³ The bfo is on the same board. Because of the high cost of 455-kHz crystals and the advantages of a variable bfo, the tunable bfo route was taken. Use of Radio Shack transformers was again made in the product detector and bfo circuits (fig. 6). Note that several volts p-p of bfo energy are required for diode saturation and proper operation of the detector. The bfo signal is amplified to prevent pulling and to develop ample bfo injection voltage. Recovered audio is clean sounding and not fatiguing, indicating low harmonic distortion from this circuit.

The a-m detector (fig. 6) is simple and requires little explanation. A half-wave rectifier, consisting of a slightly forward-biased hot-carrier diode for improved low-level signal detection, rectifies and detects the a-m signal. A low-noise audio preampli-



F1 - F4 Collins 455-kHz mechanical filters (2, 4, 8, and 16 kHz used)

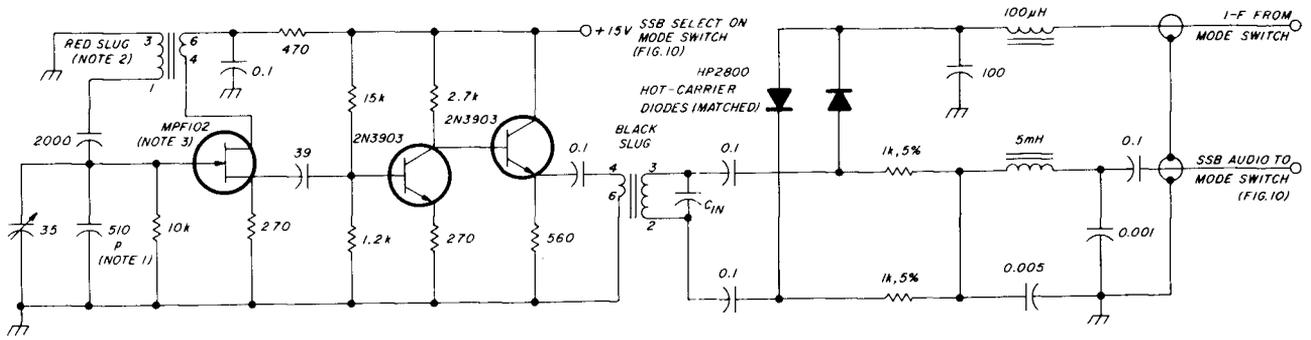
IMD filter Piezo Technology model 1433; 6-dB bandwidth: 13 kHz

L1 wound on 13-mm (0.5 in.) OD type E toroid (red) core. 26 turns 0.8 mm (no. 22) enameled wire; tap 10 turns from cold end

S1 4-position, 2 poles; shield between wafers

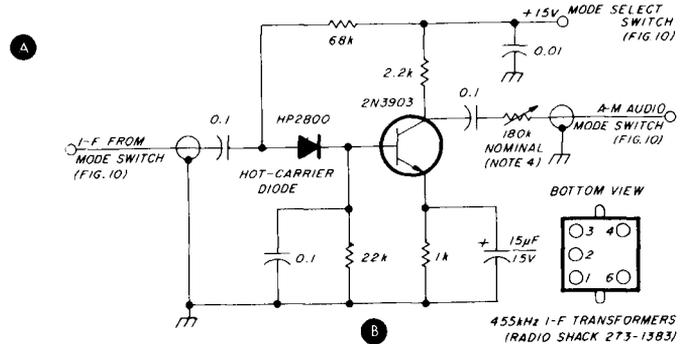
S3 filter switch; 2-pole, 4-position wafer

fig. 4. Second mixer.



- NOTES:
1. p DENOTES POLYSTYRENE CAPACITORS
 2. RED CORE DENOTES BROADCAST-BAND LOCAL OSCILLATOR COIL.
 3. MPF102 SHOULD BE HIGH GRADE. A 2N5486 CAN BE SUBSTITUTED.
 4. REDUCE FOR MORE AUDIO.

fig. 6. Ssb beat-frequency oscillator and product detector, A, and a-m detector, B.



receiver the time constants are mode-switch selected for best performance. A 120 microsecond squelch dropoff is used for a-m and fm signals; an appreciably longer delay is provided for ssb signals. Pin 4 of the 432141 module is an inverted-output of the pin-12 squelch gate, which mutes the LM-380 audio PA. I used pin 4 to light a front panel call lamp through a dc amplifier to indicate band activity.

Recovered fm audio is fed through an active audio filter in the 432141 module (3 dB gain) for conventional 6-dB-per-octave de-emphasis audio processing of the received signal. Note that pins not shown on the schematic for the 432141 module are active and are used for special applications of the RCA radios: quiet channel and fast mute. All unused pins should be unterminated.

As the 432141 requires only 10 volts for proper operation, a 5-volt zener drops the 15 V dc supply bus

to a suitable level. The RCA-module pins are not keyed; refer to fig. 8 for pin alignment. Caution: The chip can be installed 180 degrees around, and will be damaged if done so.

Since the receiver could be used for casual monitoring of various citizen and amateur services over its basic 26-30 MHz i-f tuning range, sensitivity and noise figure were contributing factors in its design criteria. More often than not, converters for frequencies above 400 MHz employ passive mixing devices, often without the aid of an integral rf preamplifier. Since these converters exhibit negative gain, not only does the mixer noise figure play a large role in system performance, but also the noise figure and

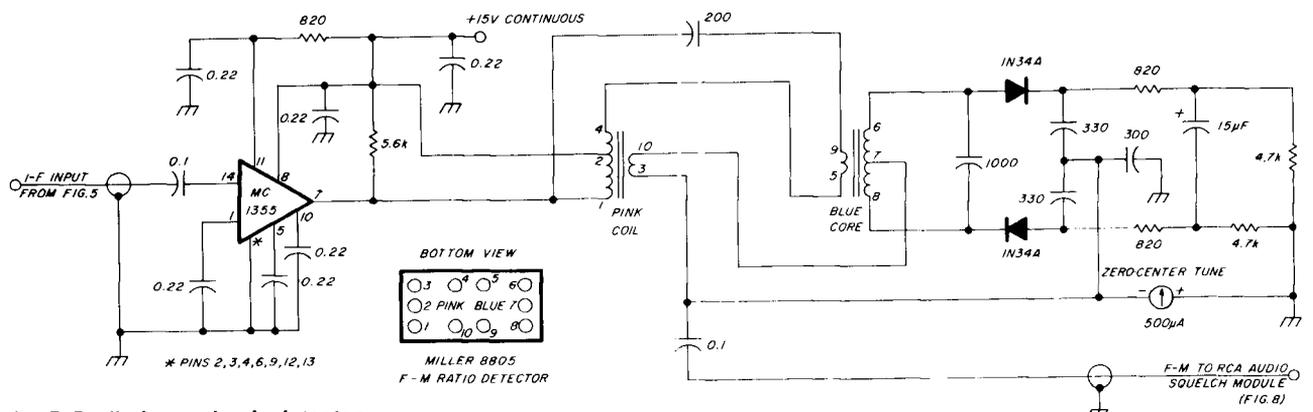


fig. 7. Fm limiter and ratio detector.

sensitivity of the i-f strip are important if optimum results are to be realized.

A single stage rf amplifier is used, using a Motorola MFE121 dual-gate mosfet (fig. 1). Agc control over the RFA is through the dc biasing level on gate 2. A small ferrite bead directly on the gate-2 lead inhibits parasitic uhf oscillations. The 20-30 MHz input and output coils of the RFA are resistive loaded to improve bandwidth, stability, and to reduce front end gain. While the resistive loading provides sufficient bandwidth to allow operation over each 1-MHz range without cumbersome preselector tracking capacitors, additional trimmers are bandswitched on the

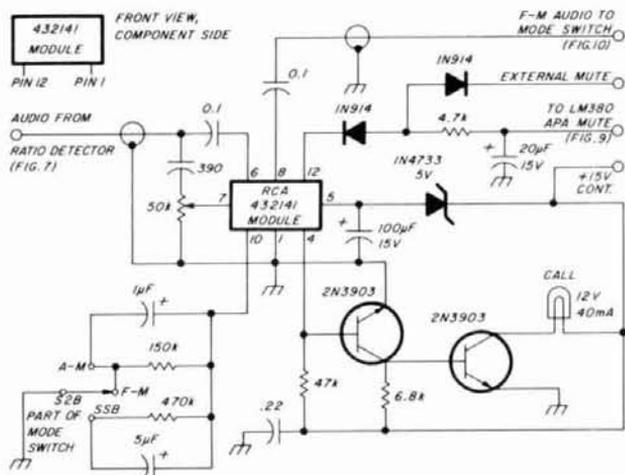


fig. 8. Noise squelch circuit for fm, a-m, and ssb reception.

lower three receiver ranges for proper RFA operation.

A *Minilabs* MLA-1 double-balanced mixer was tried in the first version of the receiver. Exotic power fets for impedance matching, high local-oscillator-injection requirements, and cost soon eliminated this scheme. The old axiom "simplest is often best" was proven in the final circuit used for the first mixer. Another MFE121, using conventional gate-2 local-oscillator injection, is employed. A ferrite bead again is required on gate 2, as in the RFA stage. Impedance transformation between the mixer output and the 10.7-MHz filter is through a capacitive divider across the mixer output tank circuit.

Vfo injection to the MFE121 mixer is filtered through a simple single LC toroidal stage, which reduces vfo harmonics and subsequent spurious receiver responses, as mentioned later. Because the vfo range covers 15.3-19.3 MHz in four 1-MHz steps, bandswitching of trimmers, as in the RFA stages, was required to resonate the filter on three lower ranges. The relatively low vfo injection frequencies

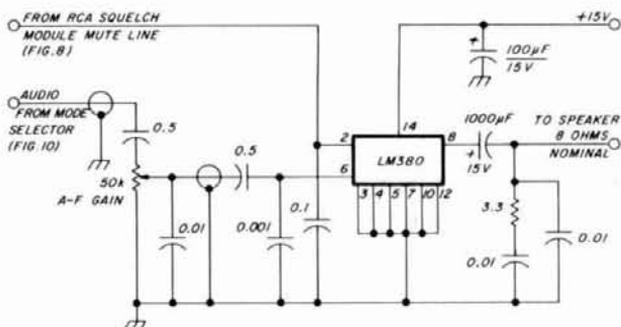
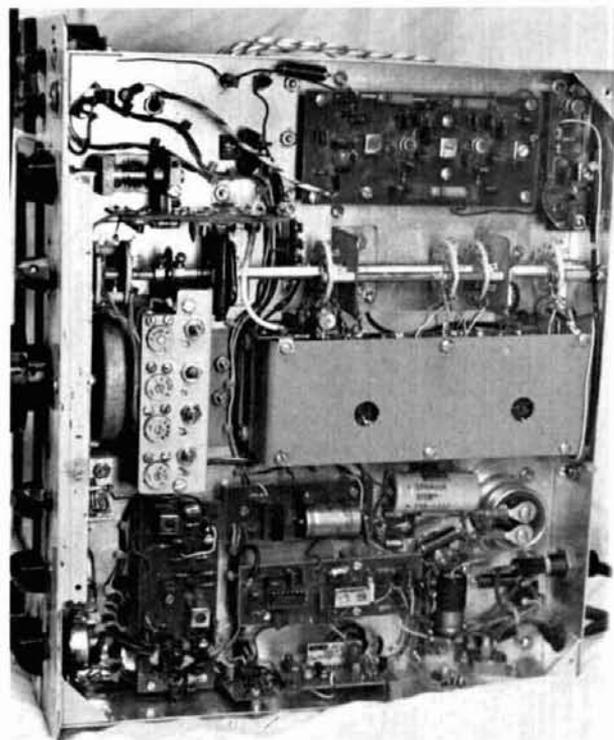


fig. 9. Receiver audio circuit.

and the desired high circuit Q prevented broadbanding of this stage. As a solution, a 3-10 pF variable capacitor mechanically linked to the vfo main tuning capacitor provides filter tracking with the vfo frequency. The purpose of the 2-18 trimmer in series with the vfo tracking capacitor (fig. 1) is to set a 1-MHz tuning range for the filter.

During alignment considerable back-and-forth tuning and peaking are required to adjust the vfo ranges, tracking, and LC-filter range.

The 10 pF capacitor coupling the LO injection to the mixer was empirically chosen. At 29.6 MHz, the



The right-angle gear drive is used to select an i-f filter mounted on the top of the chassis. The four trimmers, behind the dial assembly, are used in the main receiver vfo to select the different frequency segments.

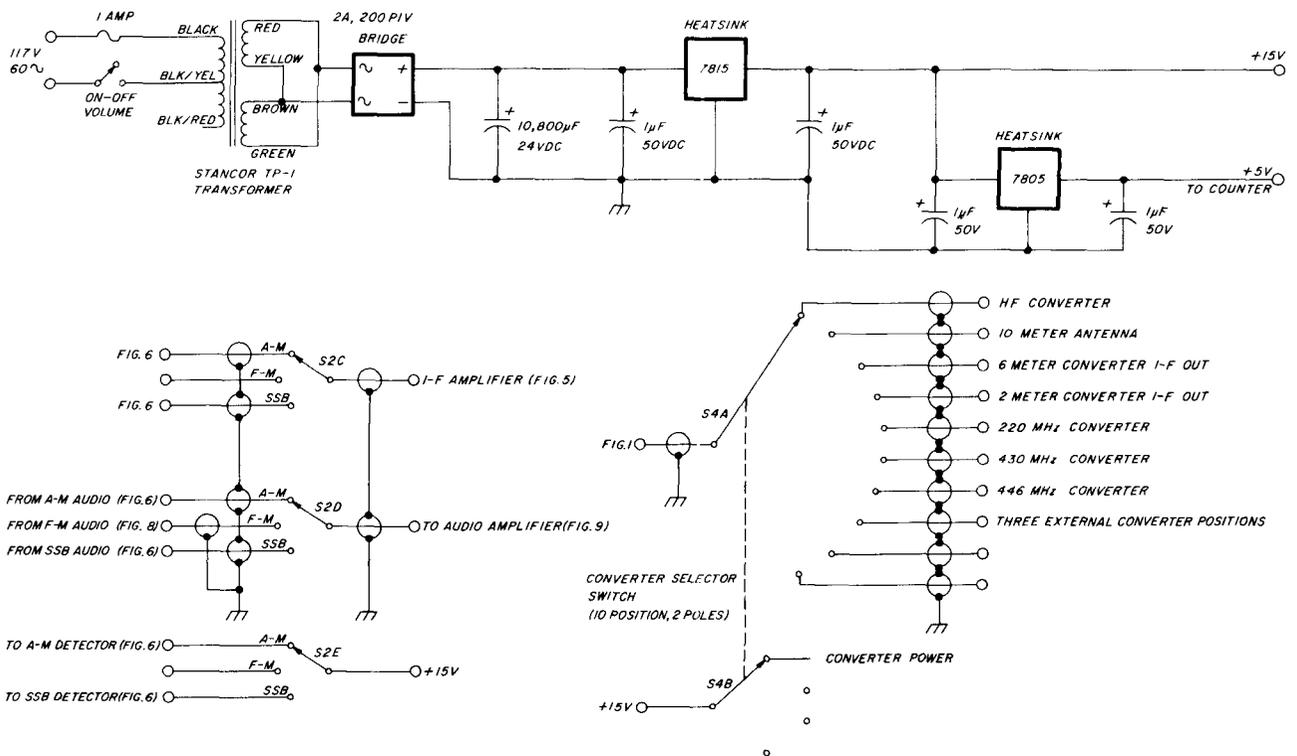


fig. 10. Power supply and switching arrangements.

vfo operates at 18.9 MHz. The second harmonic of the vfo is 37.8 MHz. If sufficient 37.8-MHz harmonic energy reaches the mixer, signals at 27.1 MHz are readily converted to the 10.7 i-f output. If the coupling capacitor is too large the harmonic injection becomes excessive; if it's too low mixer gain suffers. My receiver (worst case) has close to 50 dB of spurious rejection, or a 300-microvolt signal on the spurious frequency will produce a signal equivalent to 1 microvolt on the operating frequency.

construction

The photographs show placement and mounting of the major receiver components. The cabinet and chassis is a LMB type CO-1 enclosure. The five *Hamtronics* converters were mounted vertically on aluminum plates for space conservation and rf shielding. The two uhf converters were mounted on the left top side of the chassis, while the three vhf units flank the right side. The vfo counter is mounted vertically between the front panel and the uhf converters; the counter board is sandwiched between two aluminum plates for shielding. The chassis center was used for the 28-MHz rf amplifier and first mixer. The board is recessed below the chassis for access to the band-switch assembly.

Behind the S-meter and the 455-kHz Collins filters, another vertical shield supports the PC-board assem-

bly for the 10.7-MHz IMD filter, second mixer, and the second conversion oscillator.

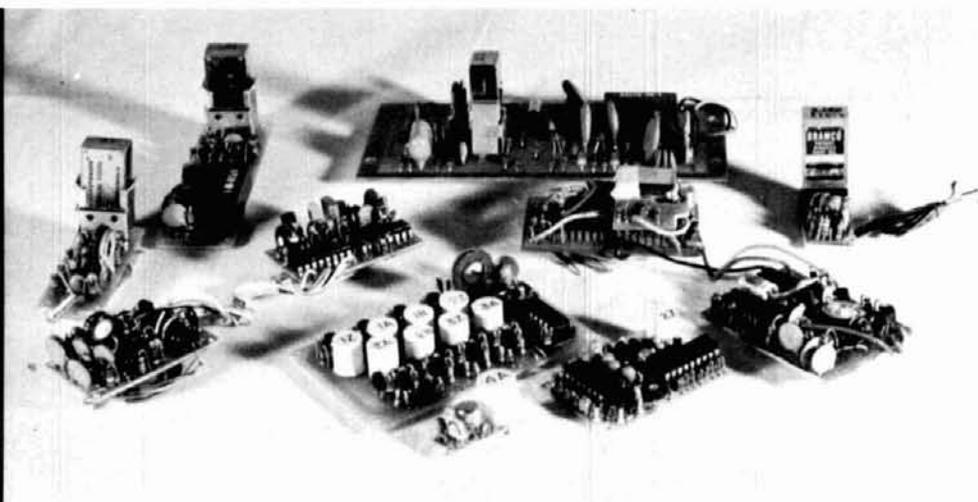
The bottom of the chassis is dedicated to the power supply components, left rear; the 455-kHz i-f stages and agc detector, right side; the vfo components, front center; multimode detectors, the squelch, and bfo, left front.

The chassis was reinforced along the cutout for the 28-MHz front end to minimize chassis flexing. The vfo tuning-capacitor supports are of heavy-gauge metal for mechanical rigidity. The bandswitch assembly transverses the entire width of the chassis, front to rear. It was constructed from several disassembled switches salvaged from flea markets. Low-loss ceramic wafers are recommended. The first two wafers are for vfo bandswitching; the third is for the vfo injection-filter trimmers. Wafers 4, 5, and 6 are for rf amplifier bandswitching. L-shaped aluminum brackets were placed between wafer sections 2 and 3, 3 and 4, and 5 and 6 for mechanical support and rf shielding.

references

1. Doug DeMaw, W1CER, "Receiving Package for 30 to 144 MHz," *The Radio Amateur's Handbook*, American Radio Relay League, 1974 edition.
2. Jim Pollock, WB2DFA, "Six-Digit 50-MHz Frequency Counter," *ham radio*, January, 1976, page 18.
3. Mike Goldstein, VE3GFN, "A Practical Discussion of Product-Detector Operation," *ham radio*, October, 1969, page 12.

ham radio



subaudible tone encoders and decoders

A review of the latest two-meter directory confirms the impression I obtained from vacation trips and from amateurs visiting the greater Cleveland, Ohio area: Most amateur repeaters are still carrier accessed. However, in localities where unoccupied two-meter pairs have become scarce and intermod problems on all bands more prevalent, some form of tone access is becoming more common. This situation is particularly noticeable in the larger east- and west-coast metropolitan areas and along the Great Lakes. Some repeaters have optional guard systems that are turned on and off automatically or by the control operator, as conditions require.

Both tone-burst and *Touch-Tone* access control are used, but the most popular method seems to be continuous subaudible tone, commonly known as *PL*, from the Motorola trade name for the system "Private Line." A selectable guard system¹ that uses *PL* has been in use on the Cleveland 16/76 repeater for some time and has led to considerable interest in various types of encoders and decoders. This article covers experiences that other club members and I have obtained about encoders and decoders we've bought or built, tried and discarded, or adapted to our use.

First, let's look at some of the reasons for using *PL*. The advantages on a control or link frequency to which access is strictly limited are obvious. Anyone who is a repeater control operator in an area where more than one machine can be heard on the same

frequency can appreciate the advantage of having an encoder on the repeater transmitter and a decoder on his monitor receiver.

PL on the repeater input also helps to minimize interference caused by intermod and sources other than amateur transmitters. In crowded areas, individuals or small groups looking for a frequency on which they can experiment or operate a special-purpose repeater, can share the same channel with much less distance between their stations than would be required without *PL*. My point is that many reasons exist for using continuous subaudible tone on input or output other than a wish to operate a closed repeater.

reed-type encoders

The encoders most used on control and link frequencies, and by operators with converted commercial gear, are those in which a resonant vibrating reed establishes the tone frequency. The advantages of these encoders over other types of low-frequency oscillators include 1) reliability and stability under temperature extremes and supply voltage changes, and 2) ability to change frequency by merely plugging in a new reed. In addition, the reed encoder generates a pure sine wave, which does not need filtering.

By Pat Shreve, W8GRG, 2842 Winthrop Road, Shaker Heights, Ohio 44120

Disadvantages include the cost of reeds (particularly if the user wishes to work several repeaters with different *PL* frequencies), and size. It is difficult if not impossible to fit a reed encoder into many of the popular hand-held and mobile transceivers.

Early popularity of reed encoders and decoders led to the adoption of the standard commercial subaudible tone frequencies for amateur use, shown in **table 1**. These have carried over into the design and production of other types of equipment. Two reed encoders, a decoder, and a combination encoder/decoder are shown in the photographs. The circuit of the Communications Specialists* miniature encoder is shown in **fig. 1**. With a Motorola Vibrasponder reed, it produces a clean sine wave to 3.4 volts rms; it will go higher, but the wave peaks will be clipped. These test results and those that follow were obtained with a 12.5-volt regulated supply voltage; Output was measured with a precision ac voltmeter, and the wave form was evaluated by

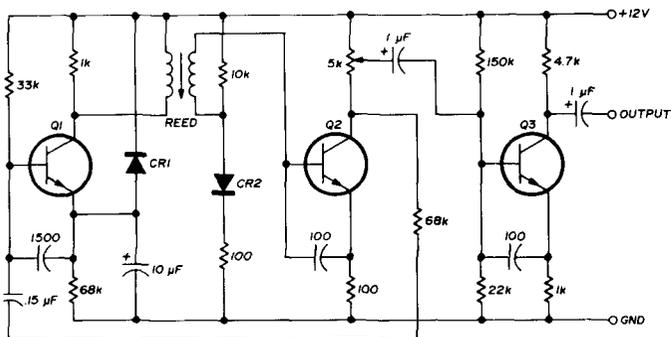


fig. 1. Communications Specialists miniature reed encoder. CR1 and CR2 are silicon diodes. Q1, Q2, and Q3 are general-purpose npn silicon transistors.

comparison with the output of a Heath IG-1B wave generator using a dual-trace oscilloscope. Tone frequency was 110.9 Hz.

Fig. 2 is the circuit of a subminiature encoder built by our club according to a design used in some Motorola equipment. It is smaller than the original or the Com Spec unit. Where space is a problem, the reed and socket can be separated from the PC board. It produces an equally good waveform, but has much lower output: from 0.22 to 0.35 volt rms, depending on the reed. It works well if the transmitter has sufficient audio amplification between the *PL* injection point, which should follow any speech filters, and the modulator.

tunable oscillators

A number of tunable oscillator circuits have been

*Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92667.

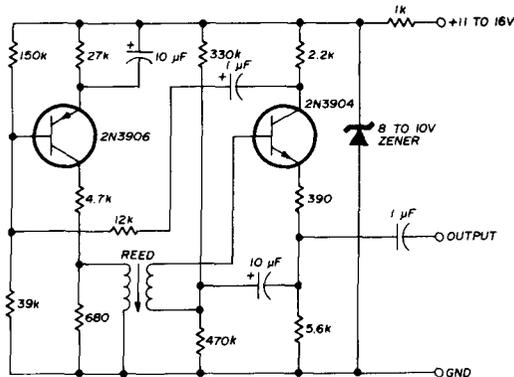


fig. 2. Lake Erie ARA subminiature reed encoder.

tried as *PL* encoders by repeater groups in or near Cleveland. The most popular was the twin-T circuit shown in **fig. 3**, which came to our club from the Great Lakes repeater group in Detroit. It is compact, inexpensive, and can be assembled from readily available parts by anyone with a minimum of experience or equipment. For satisfactory performance the frequency-determining capacitors must be molded Mylar or polycarbonate components, and precision 1% resistors should be used where shown. Even then, the circuit needs to be retuned occasionally and will give trouble in a mobile installation parked in a Lake Erie winter or in a desert sun. The wave shape is satisfactory, but the load and bias resistors may have to be changed for different output frequencies to prevent distortion.

I also experimented with tunable encoders designed around a function generator such as the Intersil 8038. Several pilot units showed promise, but none fully overcame problems of rf sensitivity and need for a more stable supply voltage than was easily obtainable in a mobile installation.

digital encoders

Use of a crystal-controlled oscillator to generate a stable frequency is nothing new; but until multistage dividers on a single IC chip became readily available,

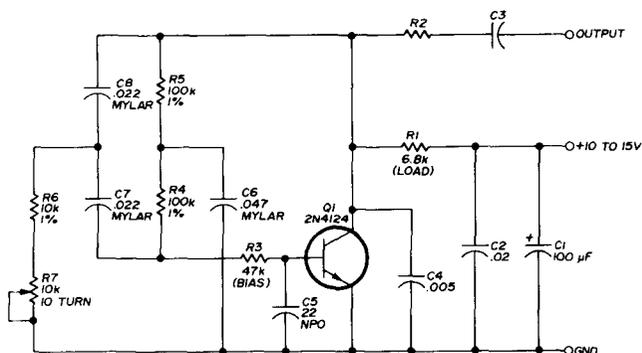
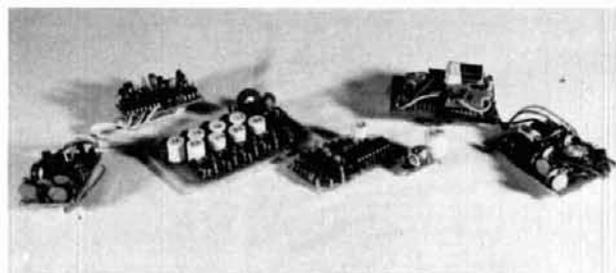
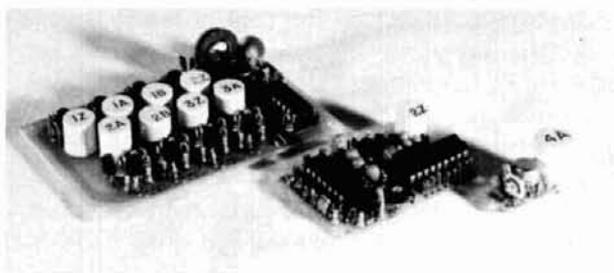


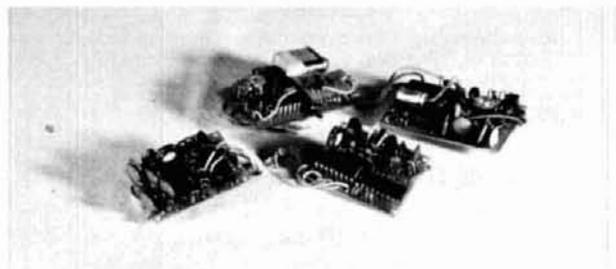
fig. 3. Twin-T oscillator. C1 is not required in battery-powered portables. Select R2 to give desired modulation level.



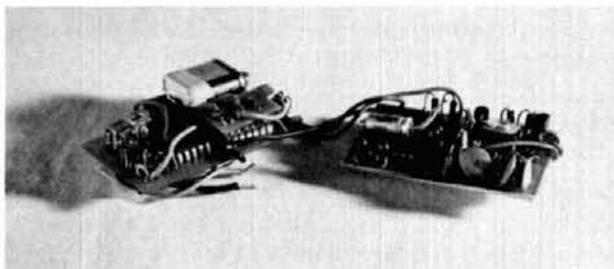
Ceramic resonator and crystal-controlled encoders and encoder-decoders.



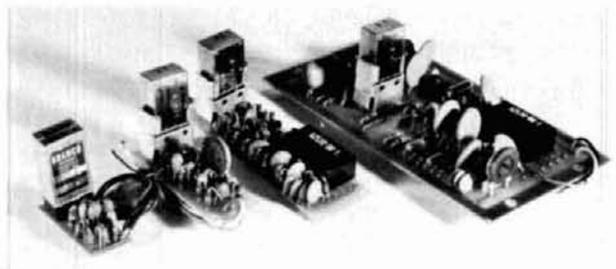
Communications Specialists eight-frequency encoder, encoder-decoder, and sub-miniature encoder.



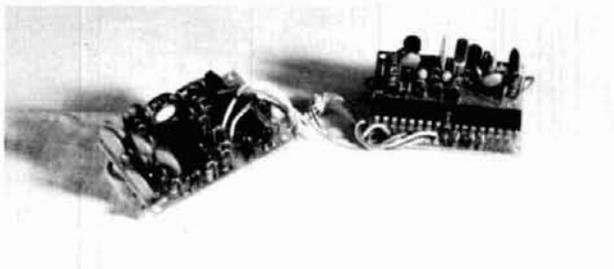
Crystal-controlled encoder-decoders.



Encoder-decoder designed by the author.



Two reed encoders, reed decoder, and encoder-decoder.



Adcom crystal-controlled encoder-decoder.

it was not practical to use high-frequency crystals to generate the low frequencies used in a subaudible tone encoder. Development of CMOS ripple counters, capable of division by factors in the thousands or millions, eliminated the need for bulky divider chains in low-frequency generators and timers and at the same time eliminated the need for a regulated 5-volt power source and sometimes difficult shielding against rf and external noise.

Descriptions of the operation and application of a number of these dividers are found in manufacturers' publications.^{2,3} Those of most interest for *PL* use are the 4020 and 4060, both capable of division by 16,384 (2¹⁴). The 4020 will accept input frequencies to 7 MHz; the 4060 to 4 MHz. The 4060 includes an oscillator circuit that can be crystal controlled.

Two approaches to the use of a multistage divider to reduce a crystal frequency to the *PL* range are possible: Use the full range of the divider and select a crystal that will give the desired output, or program the divider to give any desired output from an available crystal.

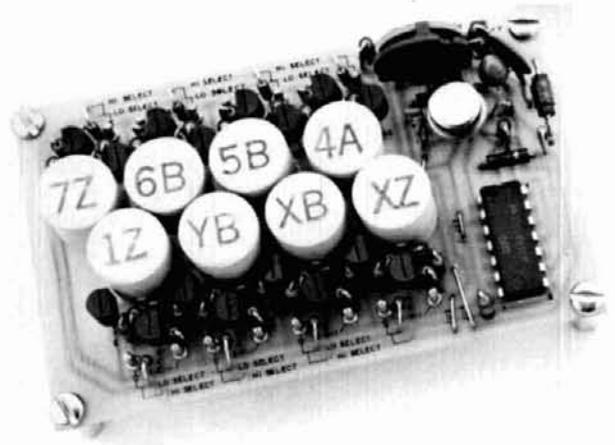
Let's look at the unprogrammed divider, in which

the crystal is selected for a specific output frequency. Only two ICs are needed, a 4020 divider and a 4030 exclusive-OR gate, which serves as crystal oscillator and digital-to-analog (D-A) converter. The crystal frequency is the desired output multiplied by 16,384. The circuit in **fig. 4** is such an encoder, designed and used by members of the Lake Erie Amateur Radio Association (LEARA), which operates the Cleveland 16/76 and 28/88 repeaters. The choice of whether to use a 1.0- or 2.2- μ F output-filter capacitor depends on whether you want a stronger signal (use 1.0 μ F) or a cleaner waveform (use 2.2 μ F). The unit leaves something to be desired in both respects. A better design could be worked out with a 4060 using the internal oscillator and substituting an operational amplifier wired as a lowpass filter for the 4030 D-A converter.

Some may ask why a D-A converter is needed at all. Certainly a square-wave digital output will modulate the transmitted signal; many solid-state CW identifiers use such an output. The trouble is, that on most amateur transmitters, square-wave modulation is far from subaudible. Many of the har-

monics in the square wave fall in the audible range, and the resulting signal can be very unpleasant. Reed encoders have a clean sine-wave output.

Another device equivalent to crystal control is used in several Communications Specialists encoders. It is a ceramic resonator much like an i-f filter operating between 250 and 500 kHz. The small size of the resonator and a special IC make possible the Com Spec microminiature ME-3 tone encoder, which is hardly larger than a postage stamp. The ME-3 circuit is shown in fig. 5. The special IC contains the oscillator, divider, and gates, which form a lowpass square wave to sine-wave converter. The output is a clean sine wave adjustable to any level to 3.2 volts rms. The output frequency can be changed by plugging in a different resonator. Similar circuitry is used in the ME-8 encoder, which provides for selection of one of eight frequencies by electronically switching the resonators, and in the combination encoder-decoder discussed later.



Eight-frequency encoder with ceramic resonators.

is lowest in the range between 3.0 and 10.99 MHz. To illustrate how the divider is programmed, assume a crystal frequency of 3.066 MHz and a desired output of 110.9 Hz. The division factor is

$$\begin{array}{r} 3,066,000 \\ 110.9 \times 4 \quad = 6912 \\ \text{subtract } 2^{13} \quad 4096 \\ \hline 2816 \\ \text{subtract } 2^{12} \quad 2048 \\ \hline 768 \\ \text{subtract } 2^{10} \quad 512 \\ \hline 256 \\ \text{subtract } 2^9 \quad 256 \\ \hline \text{zero} \end{array}$$

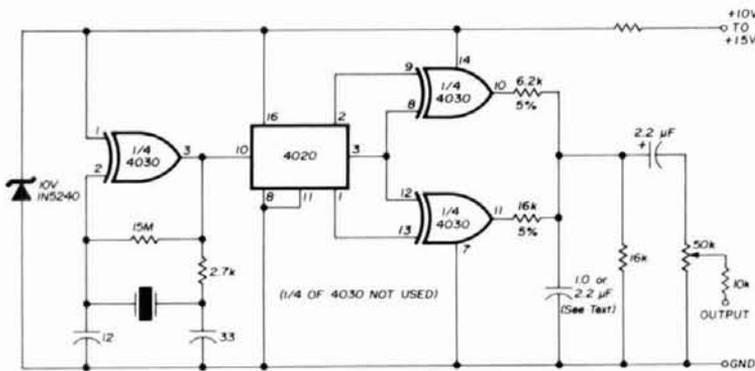


fig. 4. Lake Erie ARA crystal-controlled encoder.

Encoders in which a single crystal is used to generate multiple output frequencies can also be built with a 4020 or 4060 CMOS divider. Both these ICs can be reset to zero at any point in their counting cycle by a high-level input to the reset inverter. Since outputs are available from all divider stages from 4 through 14, diodes can be used to combine outputs to give a reset pulse after any combination of 16 oscillations of the crystal.

In my experiments, the range of crystal frequencies has been limited on the high side by divider capability and on the low side by crystal cost. For reasons explained later, the last divide-by-four step is performed by a separate device, so the divider output should be $4f_{PL}$, where f_{PL} is the desired encoder output.

The lowest of the standard tones in table 1 is 67.0 Hz. The maximum capability of the divider is 2^{14} or 16,384, so the top limit on the crystal frequency is $67.0 \times 4 \times 16,384 = 4.391 \text{ MHz}$.

My lower limit is 3 MHz, based on the price of an International Crystal general-purpose crystal, which

This example shows that when Q9, Q10, Q12, and Q13 outputs are all high at once, the counter will have divided by exactly 6912. If four diodes are connected with anodes to these outputs of the IC, and a common cathode lead is connected to the reset input, the counter will reset to zero after dividing by 6912. A similar calculation will show that diodes con-

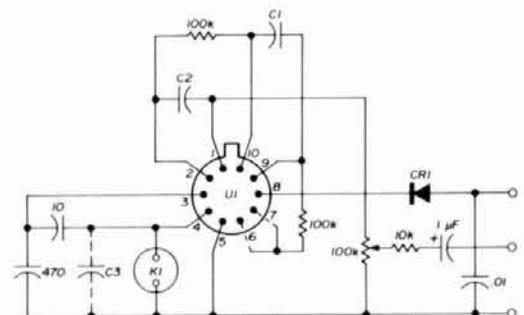
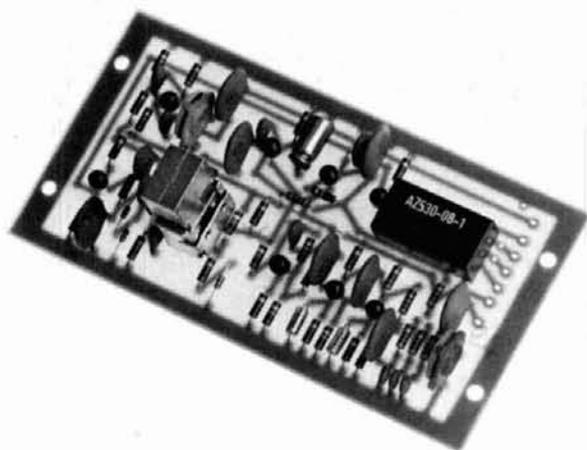


fig. 5. Communications Specialists micro-miniature encoder. Values of C1, C2, and C3 and connection to pin 6 or 7 of U1 depend on frequency. U1 is a custom-made IC; K1 is a ceramic resonator.



Reed encoder-decoder circuit board.

nected to Q10, Q11, Q12, and Q13 will result in a division factor of 6780 and an output of $3,066,000 \div 7680 = 399.21 \text{ Hz}$, which is $4 \times 99.8 \text{ Hz}$ — well within tolerance of a 100 Hz *PL*.

The reason for the external division by four is that a divider output programmed in this way is not the 50% duty cycle square wave desired for easy conversion to a sine wave. If the counter output is used to drive a dual flip-flop, such as a 4013, it will divide by four and give the desired wave; or if the encoder is to be coupled with the decoder described later, the divide-by-four operation can be performed by the decoder shift register.

One encoder of this type is made by Avcom*. No circuit diagram is available and the ICs are unmarked, but the encoder apparently uses a 4020 divider and a 4035 shift register to divide the output of a 3.334-MHz fet crystal oscillator. A 50% duty cycle square wave from the shift register is converted to an approximate sine wave by an RC lowpass filter and amplified by an npn output transistor. Maximum output of the unit I tested is 2.85 volts rms. The output waveform is reasonably good between 1 and 2.4 volts, but peak clipping occurs at higher levels and distortion appears below 0.8 volt.

decoders

Only two types of subaudible tone decoder I've tested have given consistently satisfactory results: the reed and the digital. I've heard of designs that use a linear IC such as the NE567V, frequently used to decode *Touch Tone*, but I've never seen one that will perform satisfactorily at *PL* frequencies.

To work as a *PL* decoder, the circuit should be sufficiently sensitive to respond to any signal that will quiet the receiver, have stability equal to a reed en-

coder, and have a bandwidth sufficiently narrow not to be triggered by a *PL* on an adjacent standard frequency (table 1). The circuit should have a "hang-up" connection that will release the receiver squelch, so that the operator can receive signals that don't have *PL* and can also monitor the frequency before transmitting. Outputs that will permit use with either pull-to-ground or pull-to- $V+$ squelch circuits are desirable.

reed decoders

The receiver on LEARA'S 16/76 repeater has two Motorola reed decoders. One is on the 110.9-Hz access tone and the other discriminates against the 100-Hz *PL* used across the lake in Detroit, which minimizes interference from there when the Cleveland repeater is operating in the fully open carrier access mode. The Motorola circuits are not reproduced here, but part and circuit diagram numbers are given for those interested.^{4,5}

Sensitivity is quite adequate for the excellent receiver with which they are used. Capture bandwidth is less than $\pm 1 \text{ Hz}$ on a signal with a low *PL* level; but once captured, the decoder will follow a shifting tone approximately 2 Hz either side of the nominal frequency. Tone filters are provided to eliminate the subaudible tone from the receiver output. The enable/disable function can be remotely controlled without difficulty.

A reed decoder similar to the Motorola units is obtainable from Communications Specialists either as a separate miniature model or as part of a combination encoder-decoder using the same reed for both functions. The decoder circuit diagram is shown in fig. 6. Sensitivity is 2.5 millivolts at the reed frequency at

table 1. Standard EIA subaudible tone frequencies. Higher frequencies not listed are not commonly used by amateurs.

frequency (Hz)	code	frequency (Hz)	code
67.0	XZ	118.8	2B
71.9	XA	123.0	3Z
74.4	WA	127.3	3A
77.0	XB	131.8	3B
79.7	SP	136.5	4Z
82.5	YZ	141.3	4A
85.4	YA	146.2	4B
88.5	YB	151.4	5Z
91.5	ZZ	156.7	5A
94.8	ZA	162.2	5B
97.4	ZB	167.9	6Z
100.0	1Z	173.8	6A
103.5	1A	179.9	6B
107.2	1B	186.2	7Z
110.9	2Z	192.8	7A
114.8	2A	203.5	M1

*Avcom, Inc., P.O. Box 29153, Columbus, Ohio 43299.

which combine several functions on one chip); lower cost; and a built-in audio filter to remove the *PL* tone from the receiver audio output. Instructions on how to connect it to most amateur equipment are furnished on request. Bandwidth of all digital decoders I've tested is comparable to the reed types. Sensitivity is a little less but adequate for all the receivers on which I have tried them.

decoder-detector and tunable encoder

The encoder shown in **fig. 7** uses a phase comparator and function generator to provide a sine wave output. This is because I designed the circuit as

a tunable *PL* detector, which permits the operator to match and retransmit an unknown *PL* frequency. The block diagram is shown in **fig. 8**. A 4PDT switch is added to the circuit of **fig. 7**, and the VCO portion of the 4046 is used.

In the detect mode, the crystal oscillator is disconnected. The VCO runs at four times the frequency of the function generator; this relationship is maintained by the phase comparator, and the LED connected to pin 1 of the 4046 illuminates when the loop is in lock. The frequency-adjusting potentiometer on the 8038 function generator is accessible to the operator.

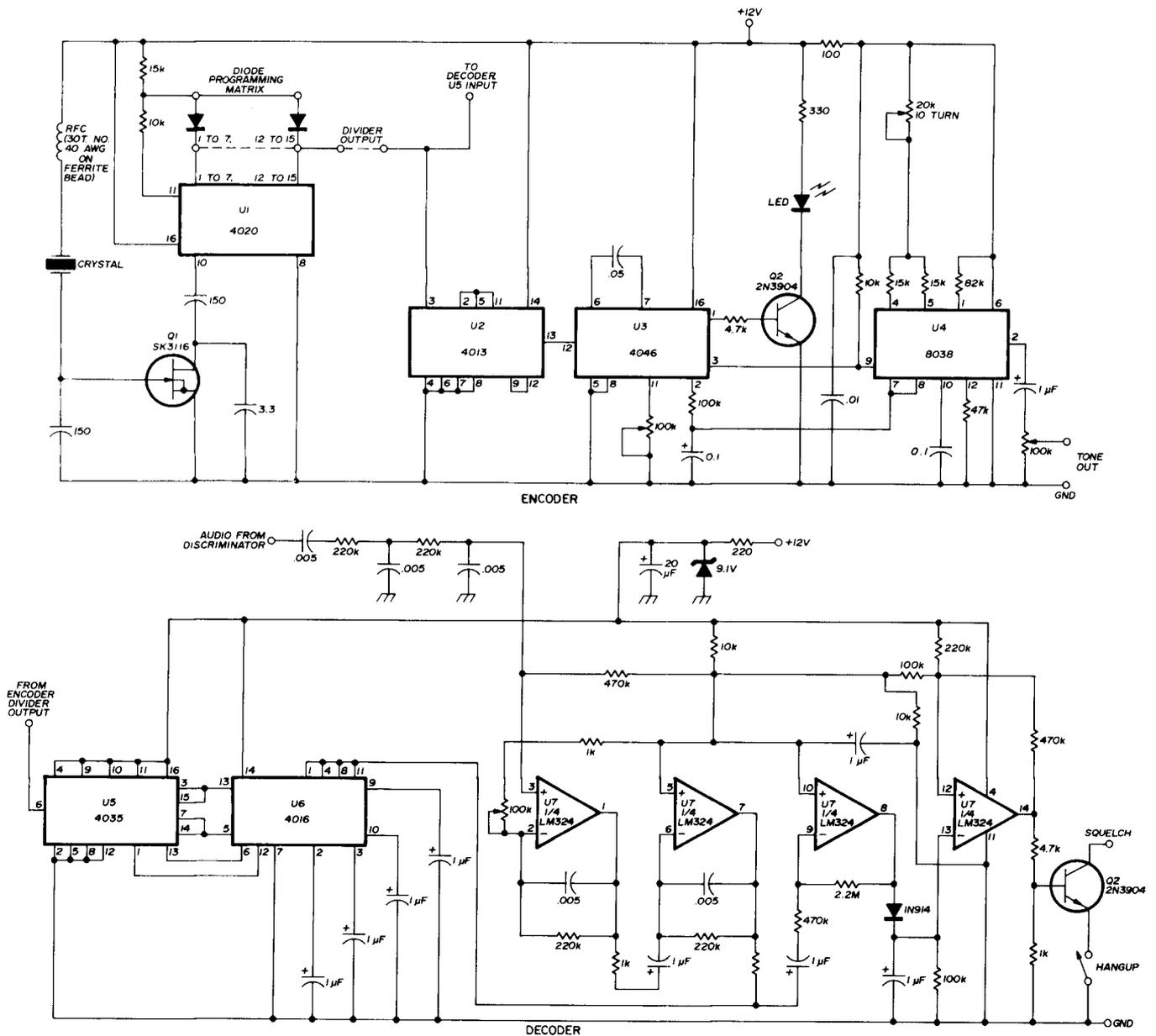
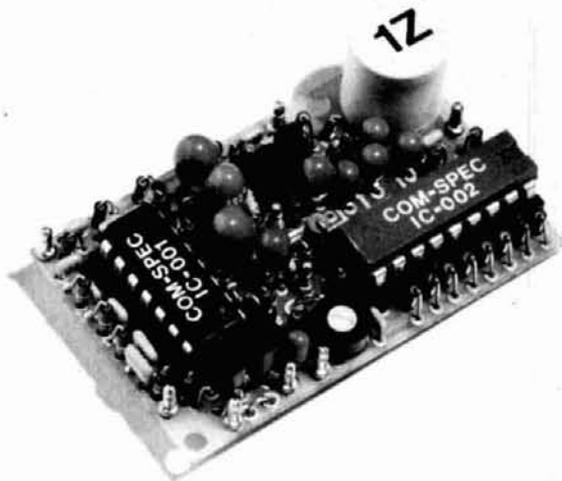


fig. 7. Encoder-decoder with programmable divider. The encoder schematic is at top; LED indicates when phase comparator loop is in lock. The decoder circuit is below; the 100k pot is used to adjust sensitivity.



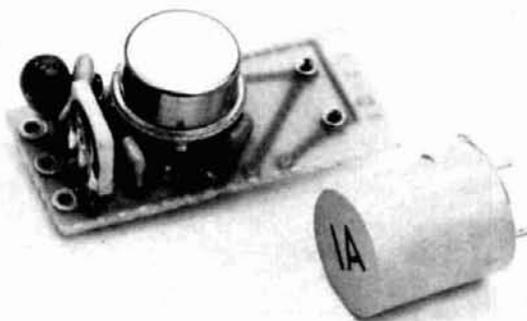
Communications Specialists encoder-decoder.

To match a received *PL* frequency, the switch is thrown to *detect* and the frequency of the free-running function generator adjusted until the decoder output LED shows a frequency match. The encoder will then transmit the same frequency as that received.

It's not ordinarily necessary to adjust the vco, which will hold its lock over a wide range of frequencies. Although the function generator is free-running it will remain within *PL* tolerances for several transmissions. The LED will show the need for re-adjustment whenever the incoming tone is received.

The greatest limitation of this system is that to acquire a repeater with an unknown access tone, you must be able to hear another station on the input frequency unless the tone is being retransmitted. With the switch in the *crystal* position, the unit operates as a normal digital encoder-decoder.

Micro-miniature encoder with ceramic resonator.



Each of the subaudible tone encoders and decoders described has its advantages and disadvantages. Where space restrictions are not a factor and frequencies are not changed often enough to make the cost of reeds prohibitive, a reed-type unit is hard to beat for stability and clean output. The units with ceramic resonators are much more compact, however, and are comparable in performance. They cost less overall if many frequencies are wanted.

My tunable model is for the experimenter or those who enjoy something different. It's not really as valuable to the traveling ham as one might think — even if you succeed in matching the unknown *PL* on that closed repeater that has been tantalizing you, you probably won't get anyone to talk to you when you do get in!

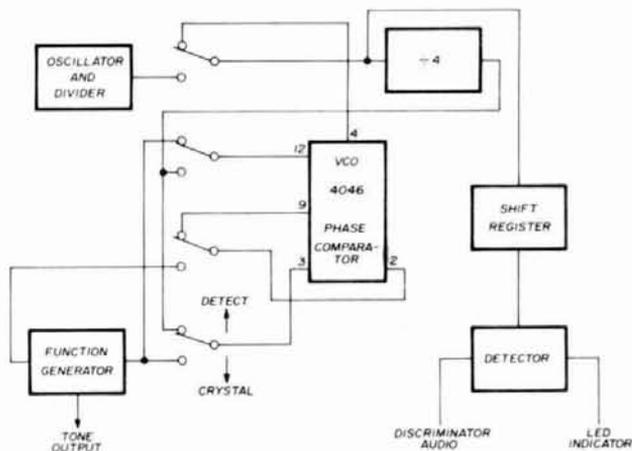


fig. 8. Block diagram of detector and tunable encoder based on encoder-decoder shown in fig. 7.

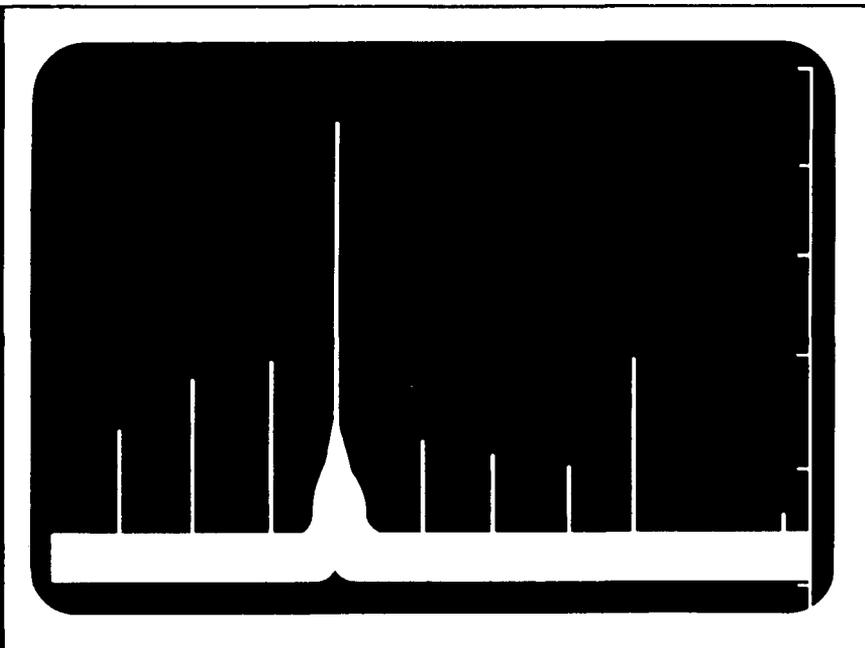
acknowledgement

I'd like to express my thanks to Spence Porter, WA6TPR, of Communications Specialists, Inc., for the opportunity to test and evaluate their products; for his description of the nature and functions of special components; and for permission to reproduce the circuit diagrams in this article.

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2. *McMOS Handbook*, 1st edition, Motorola Semiconductor Products, Inc., Phoenix, Arizona, pages 4-76.
3. *COS/MOS Integrated Circuits Databook SSD-203*, RCA Communications, Inc., pages 109 and 291.
4. *Private Line™ Decoder Model TLN 8401A Used in 136-174 MC MOTRAN Radio Sets*, Motorola Publication EPD-14779C (T-2422A).
5. *Private Line™ Encoder and Decoder Model TLN 4181A (Schematic and Circuit Board Detail)*, Motorola Publication PEPS-3531-0.

ham radio



pseudo-logarithmic display for the microwave spectrum analyzer

This pseudo-logarithmic
circuit for your
home-built microwave
spectrum analyzer
provides good resolution
and 40 dB dynamic range

In a recent article I described a microwave spectrum analyzer which covered dc to 2.5 GHz with up to 2 GHz of dispersion, 2 MHz resolution, and 50 dB of dynamic range.¹ This analyzer was built almost completely from surplus materials and has been well received by the amateur microwave community. However, the instrument has one drawback: the display graduations are linear rather than logarithmic. This limitation was discussed in the original article, and reader suggestions were solicited.

Before my spectrum analyzer article appeared (but after the manuscript was finalized) *ham radio* published a very fine article by Jeff Walker, W3JW, on the design and construction of a high resolution high-frequency spectrum analyzer.² In that article Walker described a simple and effective circuit for providing his analyzer with a pseudo-logarithmic display which allowed him to view 40 dB dynamic range at one vertical deflection setting. It seemed to me that this circuit would, with suitable modification, greatly enhance the performance of my analyzer. I am pleased to report that it did just that.

circuit description

Walker's circuit, shown in **fig. 1**, consists of an audio-frequency detector, lowpass filter, and a unique nonlinear diode limiter arrangement. My analyzer already included an i-f detector diode, the out-

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Sandy Lane, San Jose, California 95124

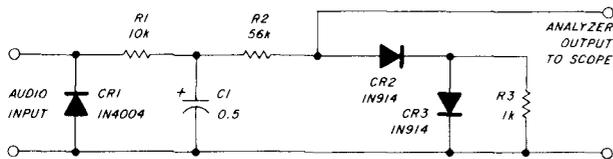


fig. 1. Pseudo-logarithmic signal-processing circuit developed by W3JW for use in a high-frequency spectrum analyzer.²

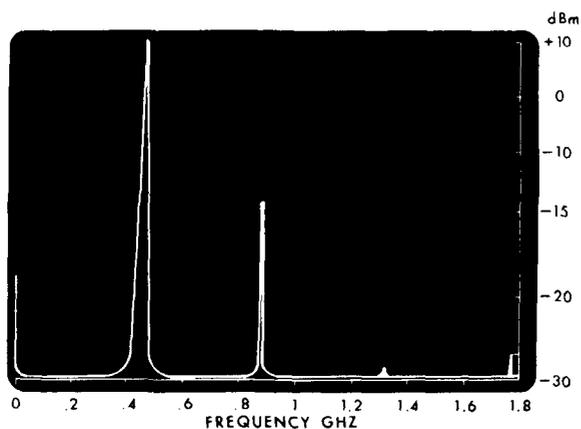
put of which I applied to Walker's filter/limiter circuit. However, I found it necessary to change the value of C1 to achieve the desired video frequency response at high sweep speeds (a value of 1000 pF is acceptable for sweep speeds of up to 60 Hz). For the logarithmic shaper circuit I replaced the 1N914 switch diodes with general-purpose Hewlett-Packard hot-carrier diodes. The final circuit values are shown in fig. 2.

Note that the detector circuit I used in my original analyzer provides a positive-going video output. If one of the more common negative-output detectors were used, it would be necessary to reverse the polarity of the Schottky diodes in the logarithmic shaper circuit.

performance

This shaper circuit enabled me to easily view 40 dB dynamic range (+10 to -40 dBm), with an unusual response which is very nearly logarithmic at 10 dB/cm at very low (-20 to -30 dBm) and very high (-10 to +10 dBm) signal levels. Intermediate ampli-

Spectrum display of a 450-MHz signal source, as viewed on the microwave spectrum analyzer with logarithmic video processing. The desired signal is at +10 dBm; second harmonic is down 23 dB at -13 dBm. Fourth harmonic is clearly visible at 40 dB down (-30 dBm). Also visible is a third harmonic component at approximately -35 dBm. Total display dynamic range easily exceeds 40 dB. Note the non-uniform vertical deflection graduations, discussed in the text.



tudes are "stretched" somewhat, as seen in the scope photograph. However, it is possible to measure signal amplitudes to within one or two dB over the entire 40 dB range, once you get the hang of it. It is possible to view spectral components as far down as -40 dBm, but scale compression at the low end is so great that you can only guess at the actual amplitude.

calibration

The display response indicated in the photograph was achieved on my analyzer with i-f attenuation set at a minimum and video sensitivity at 50 mV/cm. The

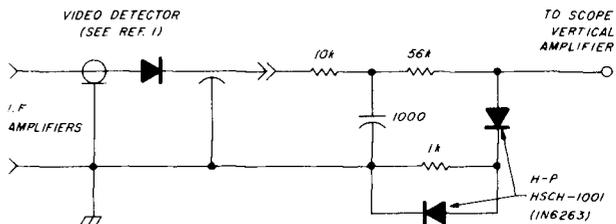


fig. 2. Signal-processor circuit as modified by N6TX for use with his microwave spectrum analyzer.¹

display was calibrated with the aid of a stable 10 mW signal source and a calibrated step-attenuator, by observing changes in the display amplitude as various amounts of attenuation were switched in. Since every analyzer is likely to exhibit its own transfer characteristics, it's a good idea to perform a similar calibration yourself if you duplicate this project.

One further point: When I change from low-band (dc to 2 GHz) to high-band (500 MHz to 2.5 GHz) coverage, the vertical scale calibration changes *considerably*. This is due to the difference in i-f gain with the i-f amplifiers operating at 2 and 1.5 GHz, respectively. Once the analyzer is recalibrated, however, I find it possible to easily resolve signal amplitudes over at least a 40 dB range, with the analyzer operating in either band.

Any feedback from readers who attempt to apply this or other signal-processor circuits would be greatly appreciated. All correspondence which includes a stamped, self-addressed envelope will be answered.

references

1. H. Paul Shuch, WA6UAM, "Low-Cost Microwave Spectrum Analyzer," *ham radio*, August, 1977, page 54.
2. Jeff Walker, W3JW, "High-Resolution Spectrum Analyzer for Single Sideband," *ham radio*, July, 1977, page 24.

ham radio

the weekender



1.2 ampere variable-voltage power supply

Whether you are a neophyte just getting started with electronics, an old-timer who hasn't built anything since the days of the 807, or an amateur in need of a handy bench supply, here is project that you can complete in a weekend, yet does not contain any exotic or hard-to-find parts. To make the project even easier, an etched and drilled printed-circuit board is available. The components are available from standard parts houses such as Allied, James Electronics, Lafayette, and Radio Shack. This should take the hassle out of getting the parts together to start the project. The finished product is a neat package that you can be proud to put your call letters on, and will find extensive use in your shack or on your work bench.

circuit description

The power supply furnishes a regulated dc output that is variable from 1.5 volts to 24 volts at 1.2 amperes. The regulation is excellent and the ripple is so low that you can power just about any type device with it, from a high gain op amp to a little QRP rig. Although the unit is configured as a bench supply, don't overlook its use for new equipment designs, as well as for powering portable or small mobile rigs in the shack. The little supply will even run the kids' HO trains as I found out last Christmas when their power pack went sour on Christmas Eve. They used the meter on the power supply for a speedometer to see how fast the trains would go before jumping the track.

The circuit, depicted in **fig. 1**, consists of three basic sections: a standard dc supply, a modern three-terminal regulator, and a metering circuit. The ac input (117 Vac 60 Hz) enters through a three-wire cord

for safety, placing the case of the supply at ground potential. A fuse is placed in the hot side of the ac line in case of a catastrophic failure, such as a shorted power transformer. The power supply is turned off and on by S1, which is coupled to the output voltage level control R3. With this arrangement you will not be so apt to connect a five-volt device to the power supply and flip on the power switch with the level control set at twelve volts. This feature can save a part or two from an unexpected smoke test.

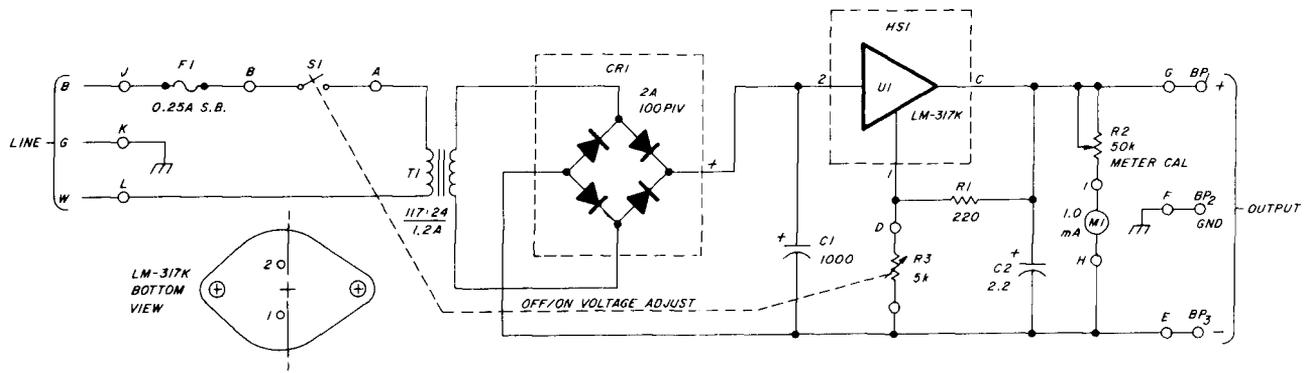
The power transformer steps down the 117 Vac to 24 Vac and isolates the circuitry from the ac line. The transformer output is applied to a full-wave bridge rectifier circuit, CR1, which rectifies the 60 Hz ac and furnishes 120 Hz pulsating dc. The dc is then filtered by the input filter capacitor C1. The basic power supply furnishes about 35 volts dc when lightly loaded.

The output from the basic power supply is applied to the input of the voltage regulator, U1. The output of the voltage regulator is controlled by a voltage divider network formed by resistors R1 and R3. As the

The author's completed power supply.



By Ken Powell, WB6AFT, 6949 Lenwood Way, San Jose, California 95120



- | | | | |
|-----------|---|-----------|-------------------------------------|
| R1 | 220-ohm 1/2-watt resistor (RS 271-000) | FH1 | Fuse Holder (RS 270-739) |
| R2 | 50k-ohm trimmer resistor (RS 271-219) | F1 | 1/4-A slow blow fuse (RS 270-1288) |
| R3 | 5k-ohm w/switch (RS 271-1714 & 271-1740) | HS1 | Heat Sink (Allied Radio 957-2650) |
| C1 | 1000 µF/35Vdc capacitor (RS 272-1032) | U1 | LM-317K (James Electronics LM-317K) |
| C2 | 2.2 µF/35Vdc capacitor (RS 272-1407) | Case | (RS 270-254) |
| T1 | 24 volt/1.2A transformer (RS 273-1480) | Spacers | (RS 270-1393) |
| CR1 | 2A/100 PIV bridge rectifier (RS 276-1152) | Heat Sink | |
| M1 | 0-1 mA meter (RS 22-052) | Compound | (RS 276-1372) |
| BP1, 2, 3 | Binding Posts (RS 274-662) | | |

fig. 1. Schematic of the variable-voltage power supply. The lettered terminals are used to indicate where the leads enter and leave the printed-circuit board. All resistors are 1/2 watt tolerance; capacitors are rated at 35 volts dc. RS part numbers are available from Radio Shack.

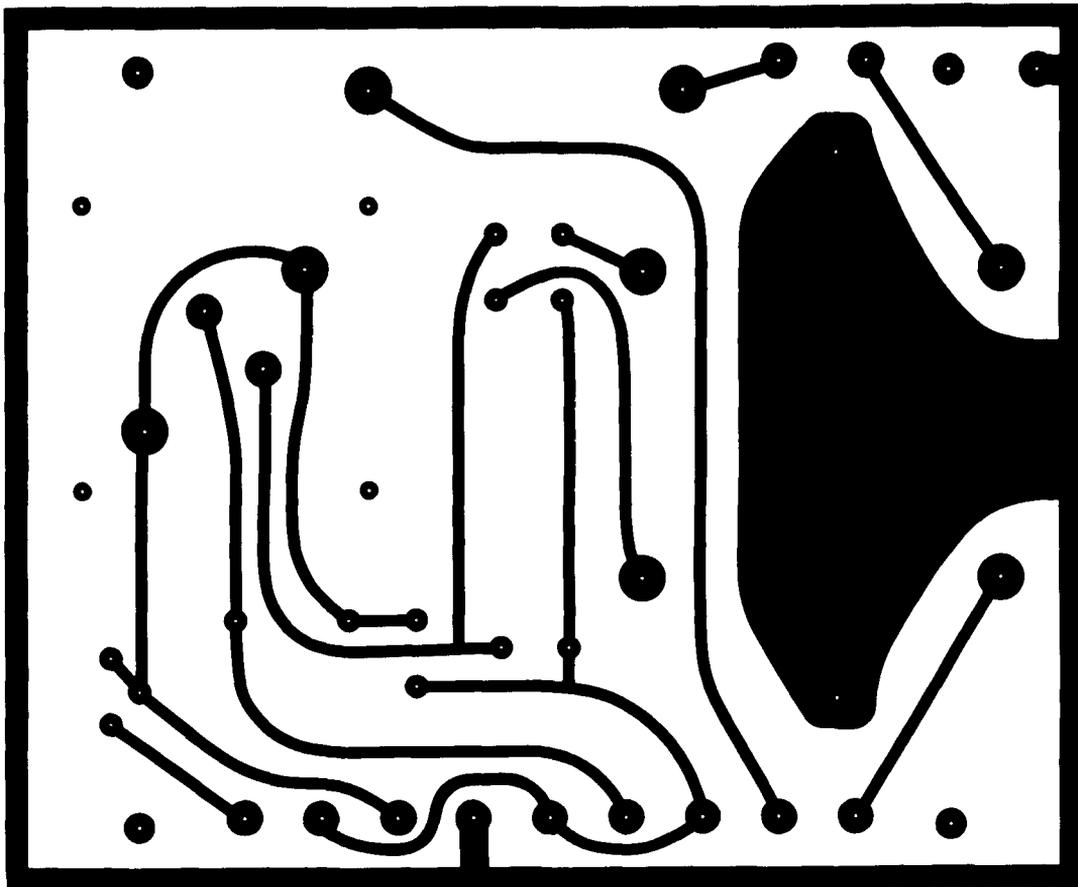


fig. 2. A full-size foil layout for the printed-circuit board. An etched and drilled board is available for \$4.00, postpaid, from J. Oswald, 1436 Gerhardt Avenue, San Jose, California 95125.

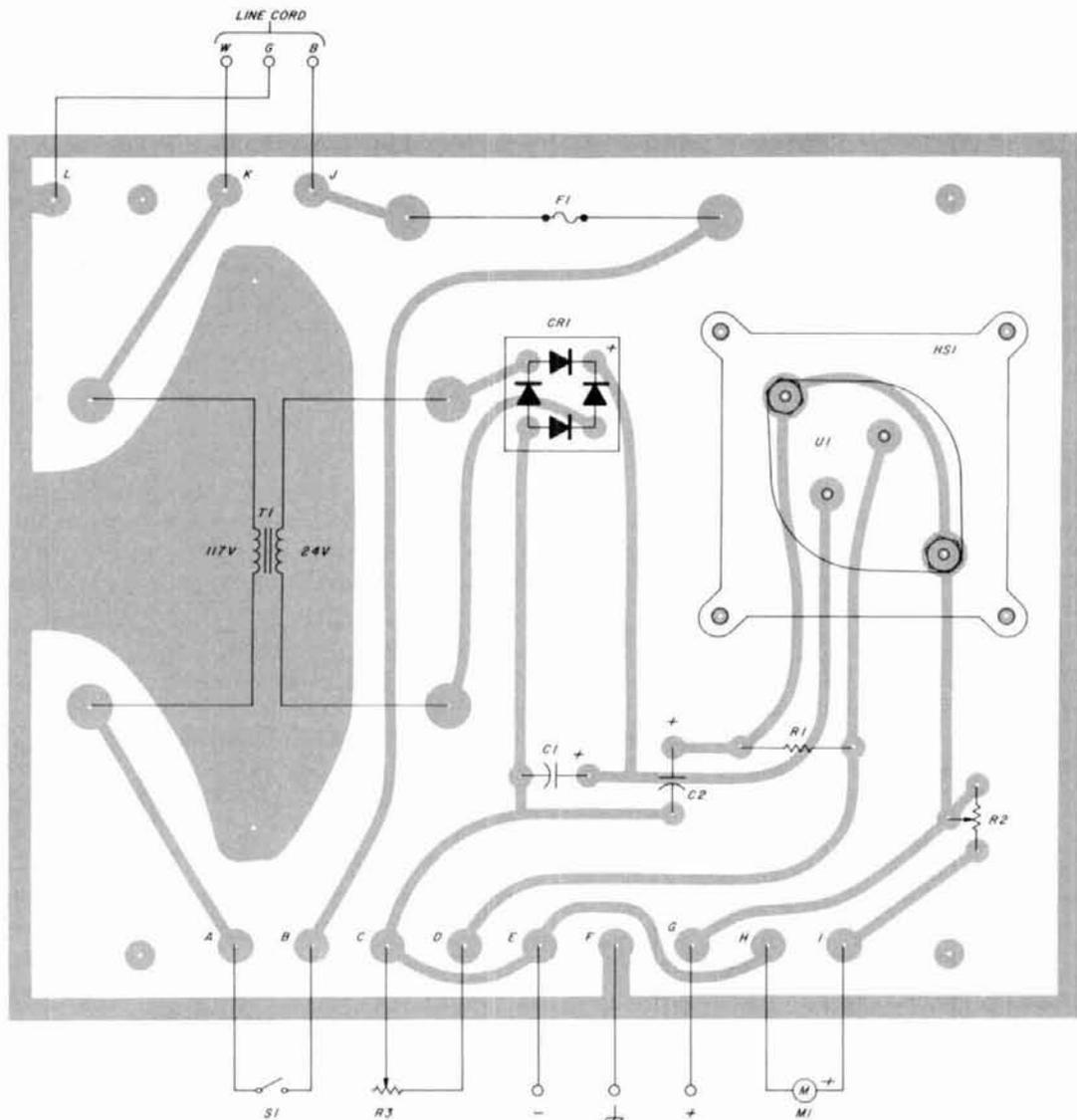


fig. 3. Parts placement diagram for the circuit board.

value of R3 is varied, the output voltage from the regulator varies accordingly; C2 is added to improve the performance of the regulator. A metering circuit is included to indicate the output voltage. A 0-1 mA meter was chosen since this seems to be the most common value available, with surplus units being advertised as low as \$1.50.

A small variable resistor, R2, in series with the meter provides an accurate means of calibration. The power supply outputs, both plus and minus, are isolated from ground so the unit may be used as a positive or negative power supply. A ground terminal is also brought out to the front panel should its use be required under certain conditions.

A full-sized printed-circuit board layout is shown in fig. 2. This layout assumes the components are the

same size as the ones specified in the parts list. If you etch your own board, I would advise using glass-epoxy board, rather than the lighter phenolic type board, since it must support the weight of the power transformer. The heavier board will provide a sturdy and stable package.

When starting construction I temporarily mounted the four corner screws and standoff spacers to the board to protect the foil side of the board while it was handled during construction. Next, mount the power transformer as this will make a sturdy base to hold the board while the smaller parts are mounted and soldered. Coat the bottom side of U1 with heat-sinking compound to form a good thermal junction between it and the heat sink HS1, and mount them to the board. The remainder of the components can

now be mounted and soldered. Fig. 3 illustrates the component layout and care should be taken to observe the polarity of C1, C2, and CR1. This completes assembly of the basic board.

If you are going to install the printed-circuit board in a chassis box as shown in fig. 4, it is best to install the interconnect wiring and the ac line cord prior to mounting the board. Slip a grommet over the line cord and solder the cord to the board. Next, solder the wires to the interconnect terminals at the front edge of the board and run them off to the left edge of the board and then double them back to the right edge of the board. Now, install the printed-circuit board in the chassis box and solder the wires from the front edge of the board to their respective components on the front panel, breaking them out at right angles to the board, parallel to their respective components. The loop left in the wiring between the terminals and the front panel components will allow the board to be removed and turned over for service, should it ever be required. The ac line cord and grommet are now placed in the cutout at the left edge of the rear panel. Again this is done to facilitate service to the board without unsoldering any wires.

If you use the meter shown in the parts list, and wish to convert the scale to read volts rather than the original milliamperes, remove the plastic cover from the meter and the two small screws retaining the meter face. Then, you can erase the numbers with a typewriter eraser. With rub-on or decal numbers, replace the original markings as follows: change 0.2 to 5, 0.4 to 10, 0.6 to 15, 0.8 to 20, and 1 to 25, leaving the zero digit alone. With a little care, you can do a very nice job on the meter and the neatly graduated

Inside view of the power supply showing the printed-circuit board and most of the internal components. The bridge rectifier is mounted behind the heatsink for U1.

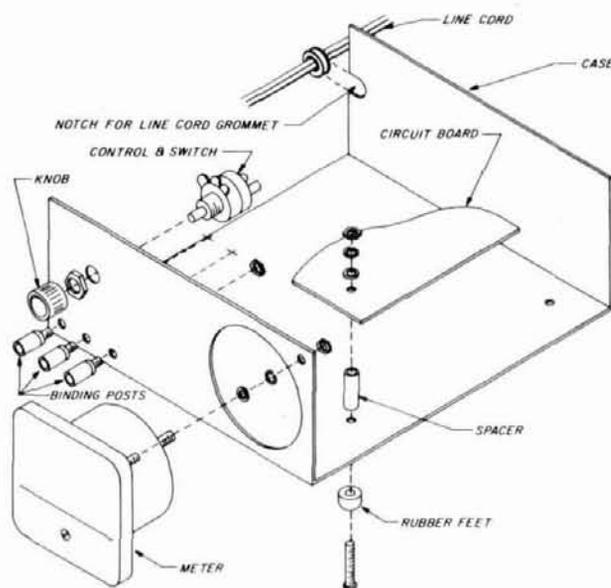
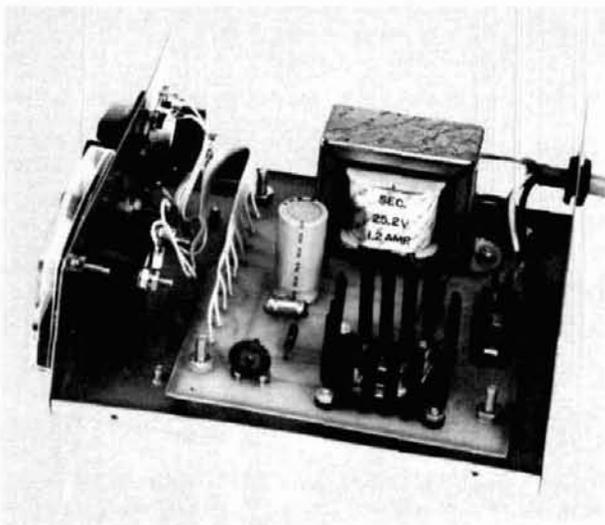


fig. 4. Mechanical details of the enclosure.

scale will be 0.5 volts per division. If you want to skip the meter work, install a 0-25 volt dc meter such as the Lafayette 99P51039V, but in this case be sure to set the calibration trimmer R2 to its minimum resistance position.

test and calibration

The first step, providing you have used a 0-1 milli-ampere meter, is to set the calibration resistor to its maximum resistance position. Now set the meter to zero with the meter adjusting screw on the front plastic meter cover. Connect a VOM or VTVM, set to 25 volts dc or higher, to the front panel output jacks. Plug the power supply into 117 Vac, advance the output level control to turn on the power supply, and adjust the control until the VOM or VTVM reads 25 volts. Now, adjust the calibration trimmer, R2, for a full scale reading of 25 volts on the panel meter M1. Next, check the readings at 20, 15, 10, and 5 volts. The panel meter should track your VOM or VTVM readings quite closely, with the greatest accuracy being achieved at the upper end of the scales.

To check the load regulation, set the power-supply output at 6 volts and apply a load, such as three no. 47 pilot lamps in parallel, to the output jacks. No change in the meter readings should take place as the load is applied and removed. If you have a scope, you can look at the power supply output under load, but under moderate load it is virtually ripple free. In the absence of a scope you can listen to the output with a pair of high-impedance headphones coupled through a 0.1 μ F capacitor, with silence being the rule. The ripple on both of the supplies I've con-

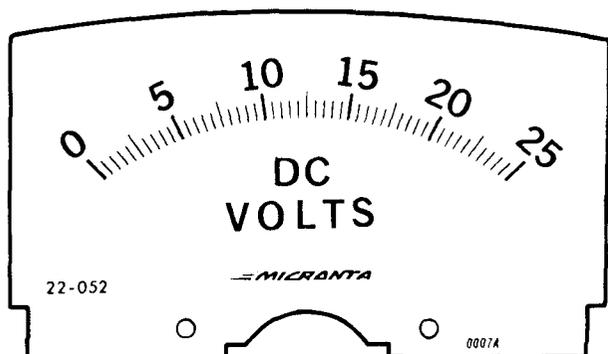


fig. 5. Full-scale meter face after modification.

structed was so low that I could not measure it with my old scope. If the supply meets the above parameters it is time to put it to work; don't worry about hurting it because it can take just about all the abuse you can dish out.

I have built two of these units and use them on the work bench, as I always seem to need both plus and minus voltages at the same time. Both supplies have been excellent performers. The esthetics of the finished product is proportional to the effort you put into it, but I found that you can actually build one of these supplies in a single weekend, have it look as good as a commercial product, and still have time for a late night QSO or two.

There are many variations that could be made, such as placing two boards in a single enclosure and making a dual output supply, or adding a switch and a meter shunt to allow the reading of output current. The fact that all the components are easily obtained, a ready made board is available, and there are no critical adjustments make this bench supply an enjoyable project; I hope you get as much satisfaction out of building and using it as I have.

ham radio

keyboard cleaning on the HP-35 calculator

Owners of HP-35 and equivalent pocket calculators may be experiencing some problem with keyboard operation. Problems such as a double entry or intermittent function is usually due to dirt under the keyboard contacts and is easily corrected.

The Hewlett-Packard series uses thin spring strips for key switches separated by a single, thin plastic sheet from the buttons. The sheet provides a barrier to prevent dirt and moisture from entering the contact area and will wear through after a year or two of operation. The sheet is the major source of trouble, not the contacts.

Plastic sandwich bags of polyethylene are a good source of replacement material for the sheet and may be used in one or two-layer thicknesses.* A common problem is how to open the case.

Models 35, 45, and 55 all use six screws for the main case. Two are easily accessible in the battery compartment, two are under the bottom feet, and the remaining two are hidden by the instruction label. The label is made of aluminum foil stock and its adhesive backing allows easy removal; if you have had one this long, you don't need the instructions. Keep the keyboard side down when removing the screws. When open, the small circuit board screws are easily visible but be careful of the double-wire contacts joining it to the main board.

The main board is screwed to the case top and removal will expose the barrier sheet and key buttons. Use the old sheet for a pattern, tracing the outline and access holes with a felt marker (Sanford *Sharpie* or equivalent). Be sure to keep the old sheet for future repair.

A clean, fine-bristle artist's brush is good for cleaning the area between switch spring strips and contact surface. It is better to work "dry" than to use commercial cleaners since these usually leave a residue. Isopropyl alcohol is good and ordinary rubbing alcohol is suitable even though it contains some water. Inspect the contact area with a magnifier for any stray hairs; a good artist's brush will have bristles firmly attached but some may break off.

This is also a good time to clean the buttons and case front. A lattice-like frame of plastic holds the buttons from the back. Use extra caution in removing this. Once removed, the buttons will simply fall out. Ordinary hand soap and water is an excellent cleaner and will not harm the plastic or markings. Use a bowl to contain the buttons and soapy water — all buttons are individual and it is too easy to lose one or two down a basin drain. The slide switch has a separate contact with special lubricant and the contact must be removed and set aside.

On reassembly, check button locations with the owner's manual. Do *not* force the screws into the plastic case or use too much torque; the original threads are quite adequate.

Leonard H. Anderson

*Mylar and Teflon sheet has been tried by some but does not work as well as the polyethylene sandwich bag plastic. *Glad* bags and *Baggies* material has been very successful locally.



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radio sounding system

An unusual application of amateur radio for atmospheric studies

How many of you vhf enthusiasts have experienced the thrill of "an inversion DX contact" and later wondered just what caused it? Such a phenomenon is caused by weather. Here's a sounding system that you can use to find out what's happening in your area. The heart of the system is called a *radiosonde*.

A radiosonde (or sonde) is a remote weather sensor that uses radio signals to furnish, by telemetry, data to a ground-based receiver and recorder. A radiosonde is usually carried aloft by a helium-filled balloon. From high above it sends back atmospheric information. Weather services use such devices daily all over the world to develop weather forecasts. While the radiosonde described here is a scaled-down version of its bigger cousins, it will allow interested vhf experimenters to study the atmosphere up to several thousand feet (or kilometers). Maybe you can forecast the next big tropo opening!

system description

Fig. 1 illustrates the amateur weather telemetry system. It consists of the airborne radiosonde and a ground-based station that includes a uhf converter, i-f stage, oscilloscope, frequency-to-voltage (F/V) converter, and a chart recorder.

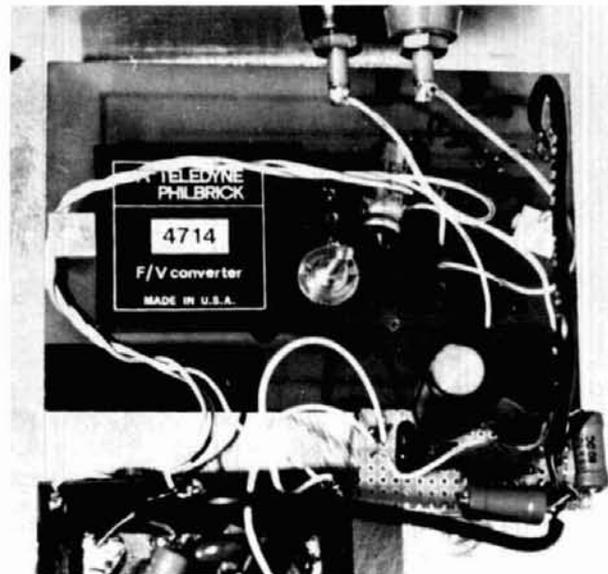
Radiosonde. A schematic of this little unit appears in fig. 2. It consists of a sensor, modulator, and a uhf transmitter that operates in the 420-425 MHz portion of the amateur 70-cm band.

Anything set aloft on a balloon doesn't stand much chance of being seen again, so I've kept the circuits simple and the costs down. This is especially important if you're planning to use these circuits in

any quantity. (A parachute design is included to help increase the odds of retrieval.)

The sonde shown in fig. 2 was originally modeled after one built by the Argonne National Laboratory for use in the 403-406 MHz band. It operates around 422 MHz and has few circuit modifications. The transmitter doesn't drift more than ± 1 MHz, so operation near the band edge is quite safe.

The sonde measures temperature changes and transmits the data to the ground station. A small thermistor, RT, changes value with temperature. Thermistor RT and capacitor CX form an RC circuit that provides an audio signal, which varies as a func-



View showing the inside of the audio-processing stage and the frequency/voltage converter.

tion of temperature. This tone modulates Q2, the transmitter, which provides a uhf fm signal. Although power output is only milliwatts, when the sonde is several thousand feet (or several km) up, its signal can be heard for hundreds of square miles.

The sondes are constructed on small epoxy PC boards (fig. 3). Half the board holds the compo-

By Larry L. Jack, KL7GLK/3, 1 East Lake Drive, Bay Ridge, Annapolis, Maryland 21403

nents, while the other half is a convenient place to tape the 9-volt battery that powers the sonde.

Receiver. Now what's needed is something to receive these interesting weather signals. I use an inexpensive fm broadcast tuner (Lafayette ST-22) for a variable i-f amplifier. It has afc and the wideband

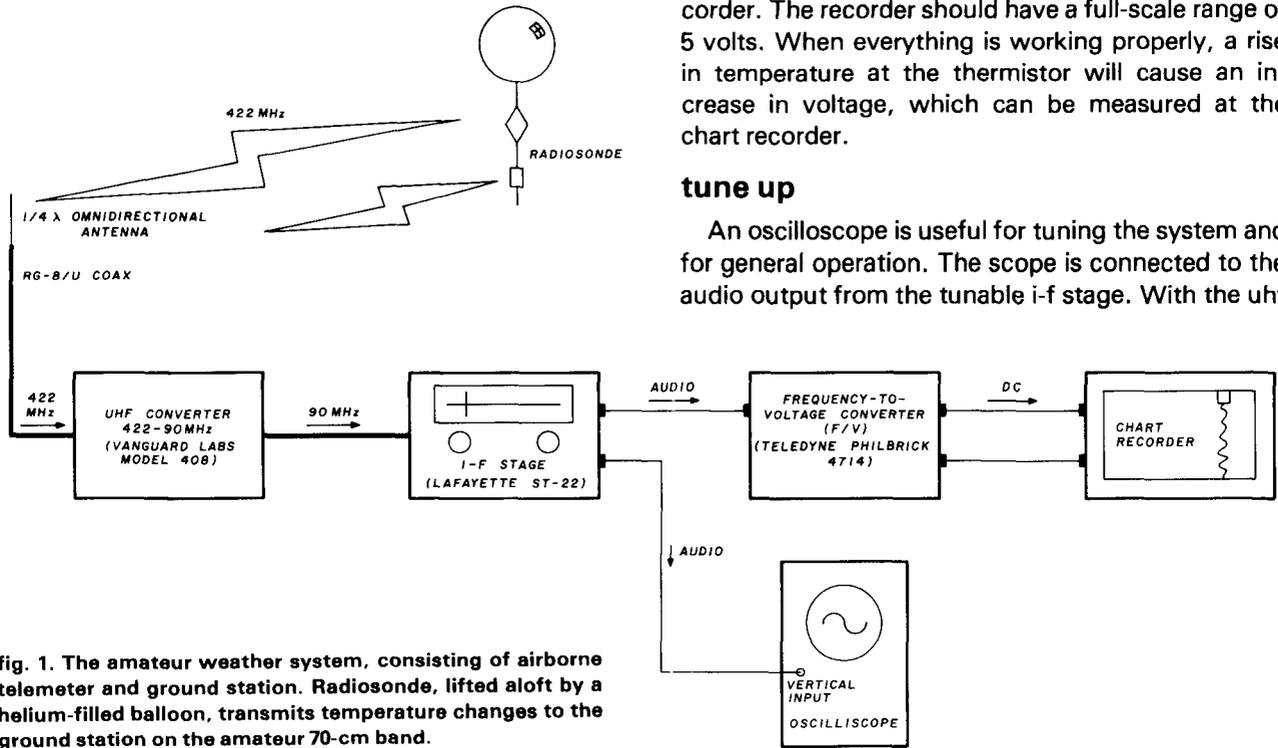


fig. 1. The amateur weather system, consisting of airborne telemeter and ground station. Radiosonde, lifted aloft by a helium-filled balloon, transmits temperature changes to the ground station on the amateur 70-cm band.

capability to receive the sonde's broad, drifting signal. It costs less than the parts needed to build a comparable i-f; but if you have the urge, feel free to experiment.

A Vanguard Labs model 408 uhf converter feeds 422 MHz, which is down converted to 90 MHz, to the i-f amplifier. You now have a receiver over which you can hear the varying temperature-dependent tone from the radiosonde.

FV converter

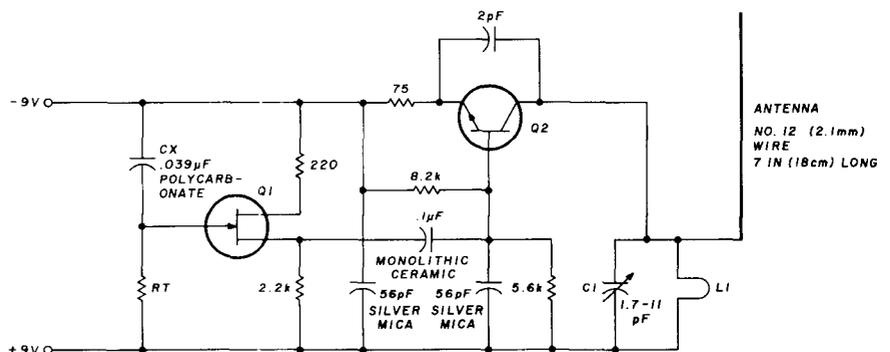
The next stage accepts the audio signals from the i-f amplifier and converts them to a dc voltage. It's a model 4714 frequency-to-voltage converter made by Teledyne Philbrick and is driven by two audio stages that provide limiting and amplification. (See fig. 4.) A dc voltage from the FV converter drives a chart recorder. The recorder should have a full-scale range of 5 volts. When everything is working properly, a rise in temperature at the thermistor will cause an increase in voltage, which can be measured at the chart recorder.

tune up

An oscilloscope is useful for tuning the system and for general operation. The scope is connected to the audio output from the tunable i-f stage. With the uhf

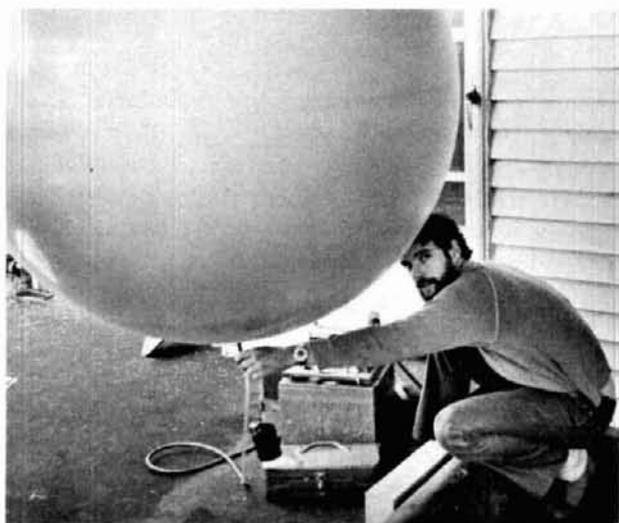
- C1 Johanson 9301 trimmer
- L1 on PC board
- Q1 2N4852
- Q2 2N3563
- RT Fenwall thermistor GA45J1
- Resistors are 5% tolerance 1/8 W

fig. 2. Radiosonde schematic. Circuit was modeled after one built by the Argonne National Laboratory. Parts count and cost are kept low, because retrieval chances are small.



converter and the i-f stage on, a characteristic noise signal will appear on the scope. With the FV converter stage on, a noise trace will appear on the chart recorder. R1 (50k) in fig. 4 is adjusted to give a 2.5-volt trace for 200 Hz into the FV converter. This gives a noise trace at approximately 4.5 volts on the chart recorder. R2 allows fine adjustments at the full-scale end of the chart recorder.

Place the sonde on an elevated nonmetallic stand,



Author KL7GLK filling balloon with helium before launch.

the thermistor will cause the trace to increase in amplitude then decrease as the thermistor cools.

If you wish to measure other weather data, other resistance-variable sensors can be used; for example, a hygistor can be substituted for the thermistor to measure humidity.

calibration

When the sonde and receiver are working properly, the sonde is ready for calibration. Begin by using a

such as a small cardboard box, and connect a 9-volt battery. Tape the battery in place on the sonde. Adjust the transmitter output tuning capacitor (C1, fig. 2) using an insulated tool. Watch the scope and chart recorder. At a point on C1 a sawtooth wave will appear on the scope. The recorder trace will smooth to a straight line between 60-70 per cent of full scale.

Where each sonde operates in this range will be a function of air temperature and the tolerance of RT, CX, and other components. A warm breath of air on

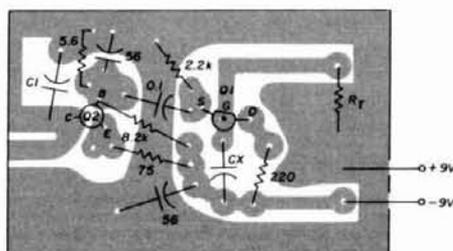


fig. 3. PC-board layout, A, and component placement, B, for the radiosonde.

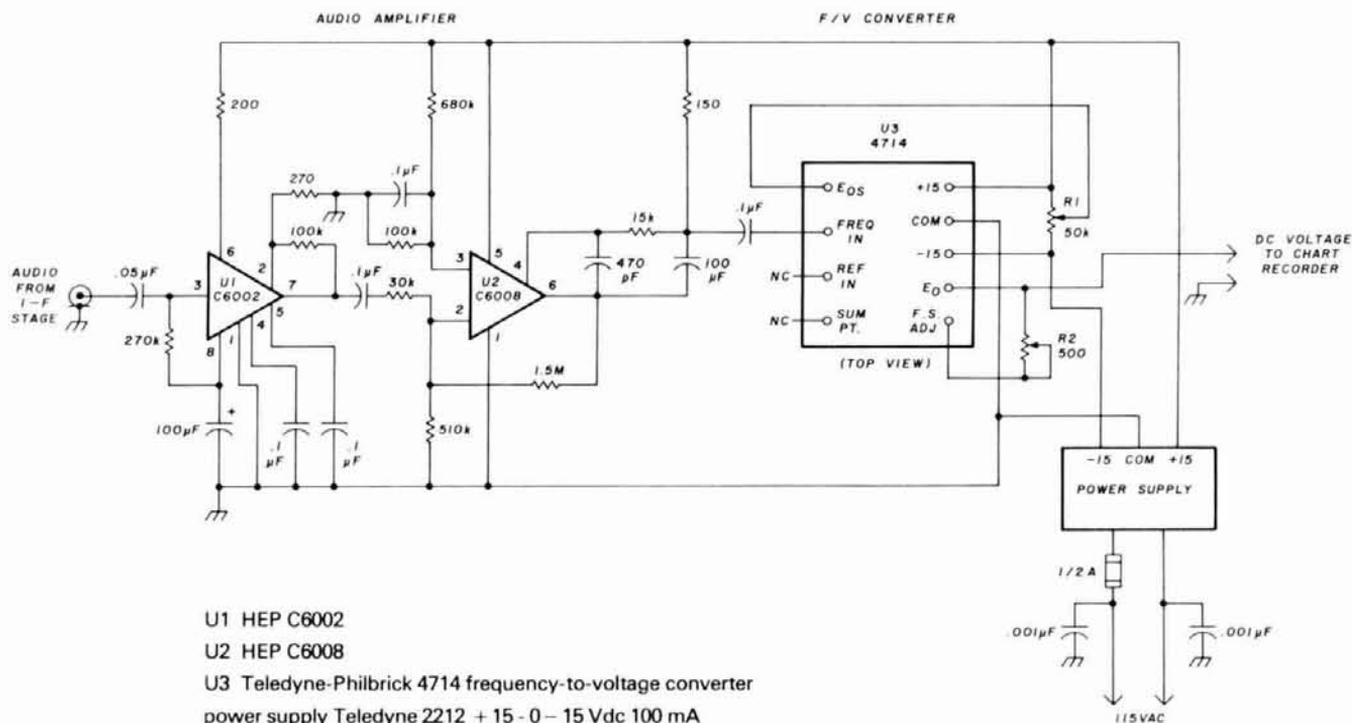


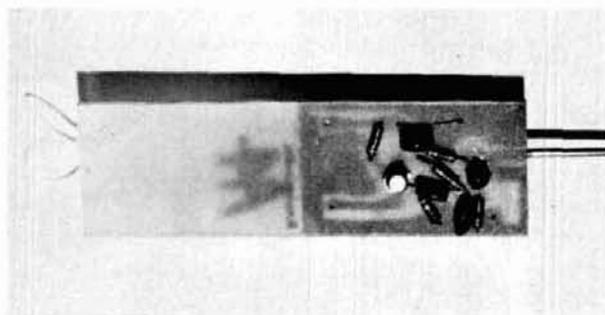
fig. 4. Schematic of the audio amplifier and frequency-to-voltage (FV) converter. The FV converter drives the chart recorder.

thermometer to measure the room temperature that the sonde is monitoring. Mark this value on the chart recorder alongside the trace. Place the sonde in a chamber that can be cooled (by ice, for example) to 68F (20C) below room temperature. When the sonde stabilizes at this new temperature, mark this value on the chart next to its corresponding trace.

The system has a linear temperature response from 23 to 86F (-5 to 30C). Using the two calibration points to form a temperature-to-voltage slope, any temperature point along this slope can be interpolated. A calibration factor can be calculated by dividing the change in temperature by the change in voltage. For example, if the two calibration points were 36 and 75F (2 and 25C), for a voltage change of 1.0 volt there would be a 36F (2.3C) change per 100mV; *i.e.*, $25 - 2 = 23$, and $23/1.0 = 2.3$.

Most chart-recorder paper is divided into 100 lines. On a 5-volt range, each line is 50 mV, so a change of ± 1 of these divisions is a change of $\pm 3.6F$ ($\pm 2.3C$). This number, (2.3C or 3.6F), is assigned to the sonde as its *calibration factor*.

Before flying the sonde, measure the outdoor air temperature and mark it on the chart. Once the sonde is in flight, the calculated temperature can be subtracted from this point, giving the temperature



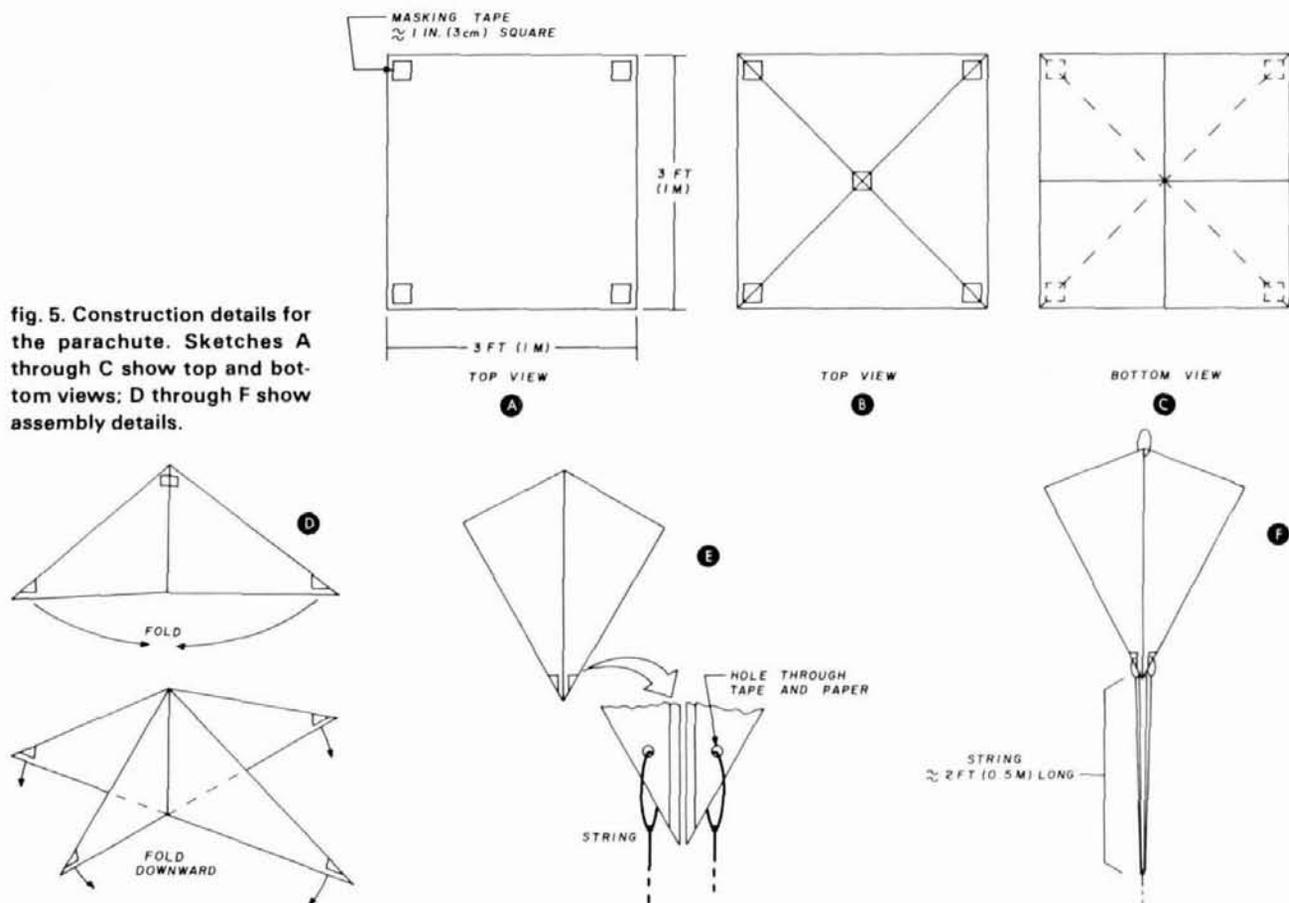
The 422-MHz radiosonde without the 9-volt battery attached.

the sonde is measuring. Because of differences in tolerances of each sonde's components, each sonde will have a slightly different calibration factor. It's therefore a good idea to calibrate each sonde individually.

preflight

The question of determining altitude for the corresponding sonde data now arises. It's necessary to know the rate of the balloon's ascent, so that the altitude can be calculated as a function of time, as measured by the chart recorder; *i.e.*, chart speed =

fig. 5. Construction details for the parachute. Sketches A through C show top and bottom views; D through F show assembly details.



2 inches (51mm) per minute. A rough rule of thumb is: A 1-ounce (30-gram) helium-filled balloon, filled to just lift a 4.9-ounce (139-gram) weight, will ascend alone at 613 feet (187 meters) per minute.

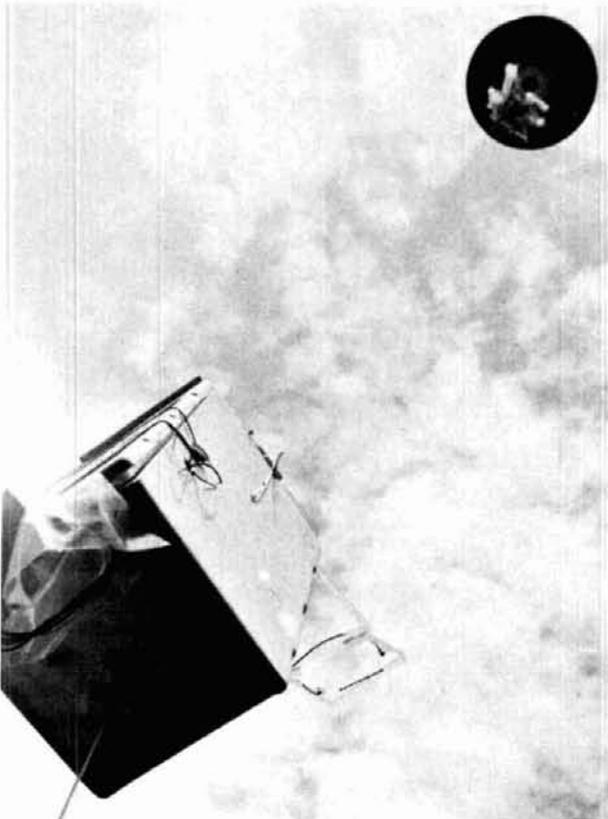
Attaching a sonde and a parachute will upset these figures, so if this method of determining the rise rate is used, some experimentation will be necessary. The best method of determining altitude is to double track the balloon with theodolites and calculate the altitude by triangulation.

The balloons used to lift the radiosondes and their parachutes (there is a good reason for the parachutes, by the way) are meteorological balloons.*

A 1-ounce (30-gram) balloon is overfilled to lift the sonde but it works. These overfilled balloons will burst sooner than others (that's why there's a parachute), but usually long after they have drifted out of radio range. As mentioned before, the balloons are filled with helium. Small cylinders of helium are available from firms selling compressed gases.

the parachute

When the balloon bursts, your little sonde could hurtle down through someone's property if the parachute and sonde didn't descend nice and slowly. If



Radiosonde balloon and parachute just before launch. This is the type of equipment used by the United States Weather Service (Beuker's Laboratory model 1207).

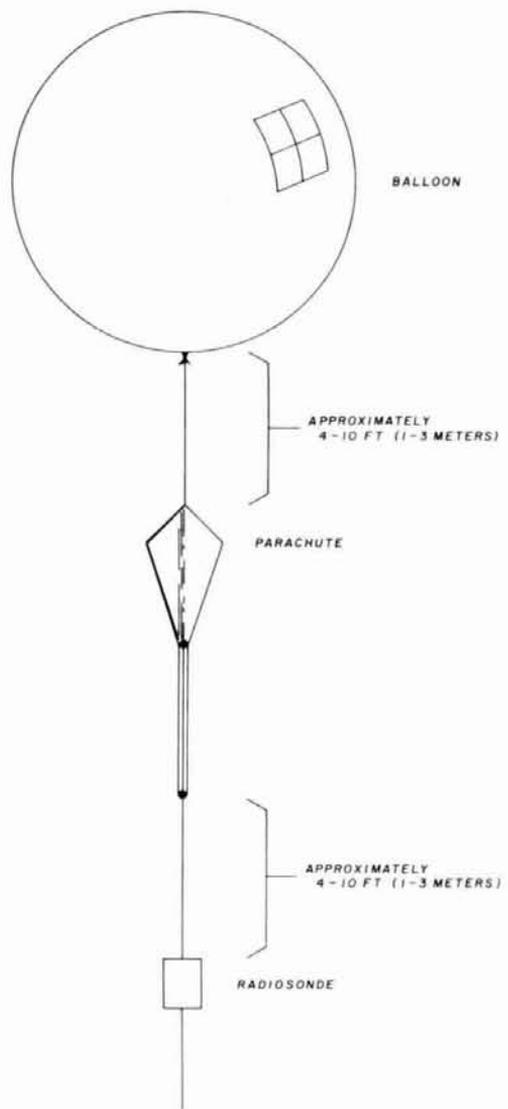


fig. 6. Telemeter assembly showing the balloon, parachute, and radiosonde. When the balloon bursts, the parachute opens, allowing the sonde to drift gently to earth. Your name and address attached to the sonde should increase the chances of its retrieval.

you attach a return address label, maybe someone will mail it back. We get a good number returned this way. You can make your own parachutes with some paper folding and string (see figs. 5 and 6).

flight

The system is now ready for use. I'll describe the procedure my compatriots and I use to fly the sonde. First of all it's necessary to file a *Notice to Airmen* by calling the Federal Aviation Administration (FAA).

*Available from Weather Measure Corporation, P.O. Box 41257, Sacramento, California 95841.

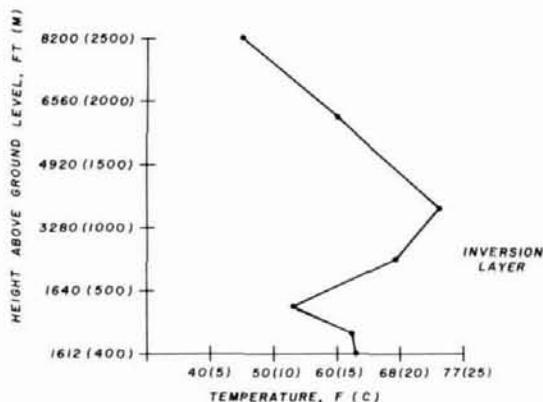


fig. 7. Inversion sounding trace showing temperature changes as a function of altitude between 400 and 2500 meters (1312 and 8200 feet).

Look for a flight service station in your telephone book. They request you tell them the number and times of the flights and the location of the launch. There's nothing to get upset about, but be sure you file your notice.

All right, the balloon is filled, and the train of string, parachute, and sonde is then attached.

Turn on the receiving and recording equipment, attach the radiosonde battery, and tune the receiver to the sonde frequency. Make a note on the chart of the sonde's calibration factor. Then make a side-by-side comparison of the sonde temperature with that of a thermometer; mark the thermometer reading beside the chart-recorder trace. If theodolites are being used to track the balloon, alert the operators. Release the balloon and mark this moment on the chart recorder. As the sonde ascends tune the receiver to follow any transmitter frequency drift.

The chart recorder will begin to show the changes in the air temperature as it takes a cross section, or profile of the atmosphere. Depending on conditions, the temperature changes will range from mild to dramatic.

For instance, the inversion phenomenon mentioned in this article's introduction would look similar to the profile in fig. 7.

onward and upward

From here on, I refer you to the vast number of meteorology books. As more experience is gained using radiosondes, your questions on weather phenomena will quickly outdistance the scope of this article. The system described won't put you in competition with the National Weather Service, but it will provide an interesting medium for a personal study of the atmosphere and an unusual use of amateur radio. Have fun with it.

ham radio

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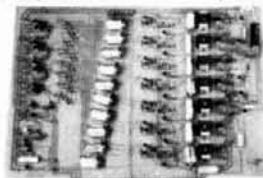
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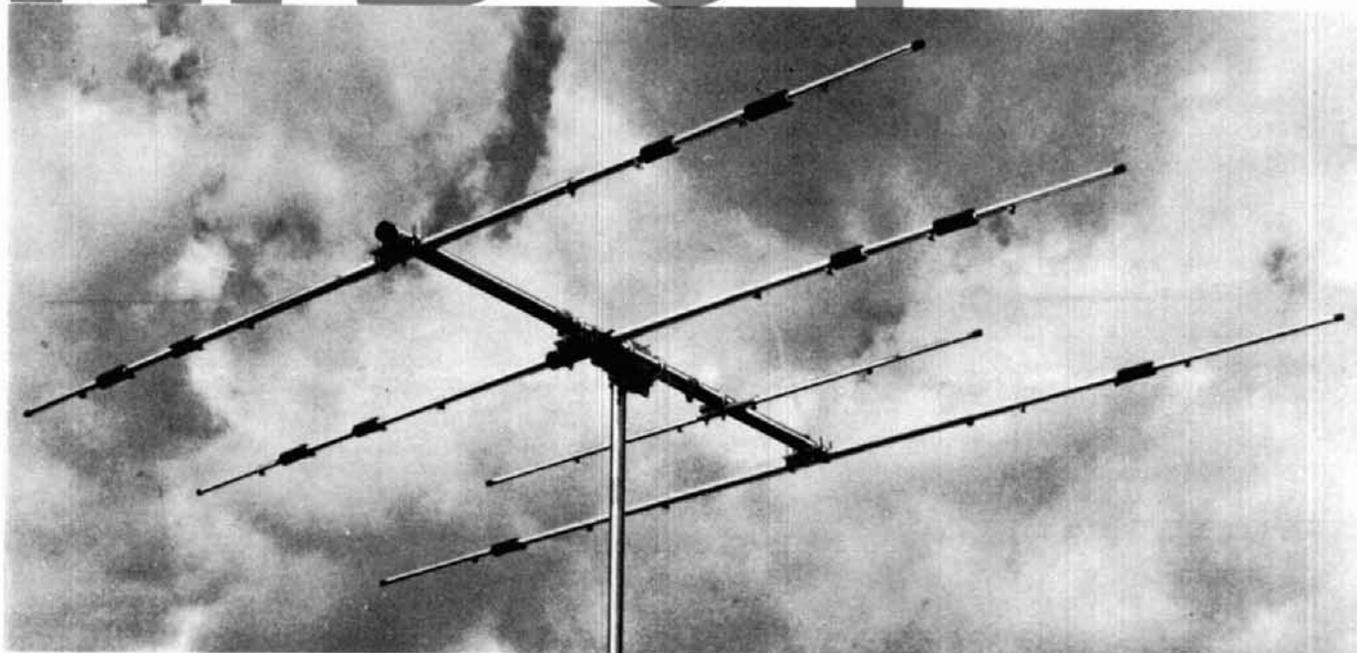


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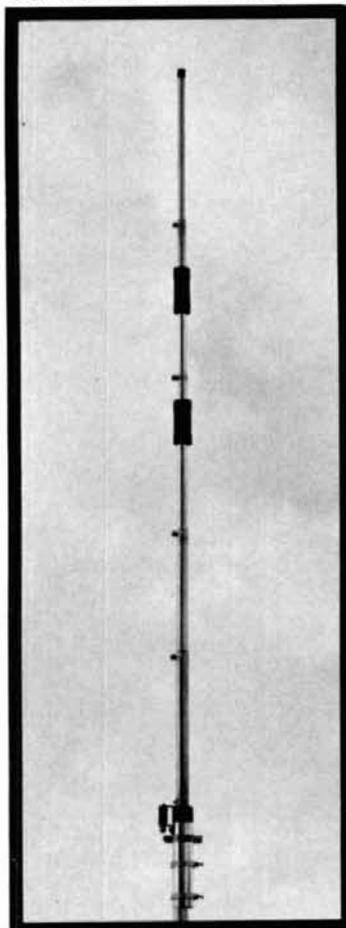
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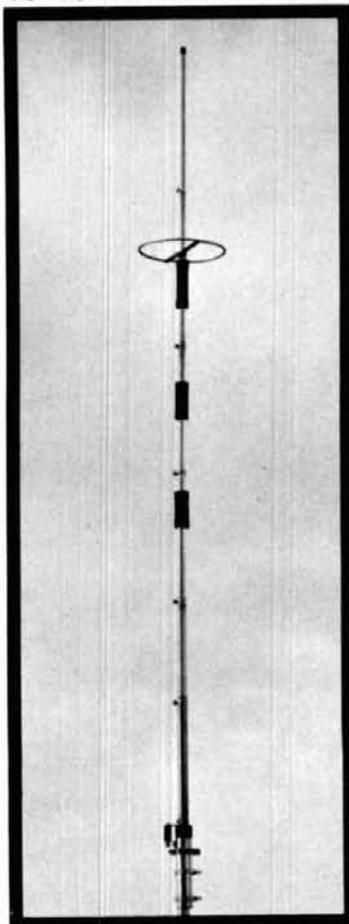
HF VERTICALS BY CUSHCRAFT

10-15-20 METERS



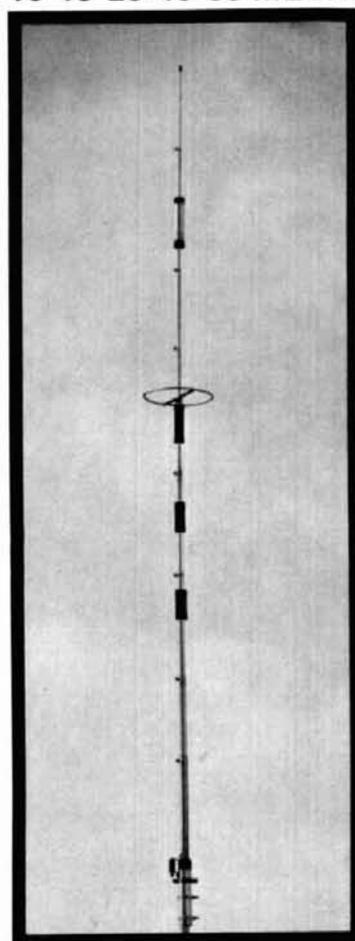
ATV-3 Cushcraft's ATV-3 multiband vertical provides low VSWR operation for both SSB and CW on 10, 15, and 20 meters. Matched to 50 ohms, built-in connector mates with standard PL-259. Stainless-steel hardware is used for all electrical connections. The ATV-3 is a compact 166 inches (4.2 meters) tall. Rated at 2000 watts PEP.

10-15-20-40 METERS

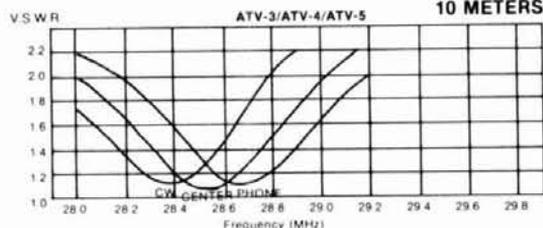
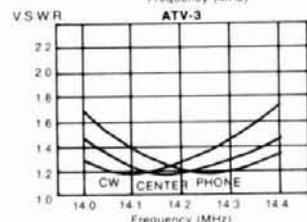
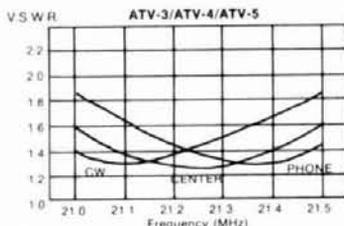


ATV-4 The Cushcraft ATV-4 four-band vertical antenna has been optimized for wide operating bandwidth on 10, 15, 20, and 40 meters. SWR is less than 2:1 over the CW and SSB segments of 10, 15, and 20. The 2:1 SWR bandwidth on 40 meters is approximately 240 kHz, may be quickly and easily adjusted to favor any part of the band. Coaxial fitting takes 50-ohm transmission line with PL-259 connector. Overall height, 233 inches (5.9 meters). Rated at 2000 watts PEP.

10-15-20-40-80 METERS



ATV-5 The ATV-5 trapped vertical antenna system has been engineered for five-band operation on 80 through 10 meters. The high Q traps are carefully optimized for wide operating bandwidth: 2:1 SWR bandwidth with 50-ohm feedline is 1 MHz on 10 meters, more than 500 kHz on 15 and 20 meters, 160 kHz on 40 meters, and 75 kHz on 80 meters. Instructions are provided for adjusting resonance to your preferred part of the band, CW or SSB. Built-in coaxial connector takes PL-259. Nominal height, 293 inches (7.4 meters). Rated at 2000 watts PEP on all bands.



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More Details? CHECK — OFF Page 142



outboard LED frequency display

for the Heath HW2036

Operating the HW2036
no longer
has to be done
in the dark —
this outboard display
shows the frequency
set in the switches

Picture yourself driving along the freeway one evening and your Heath HW2036 comes alive with an interesting QSO on the 01/61 repeater. Just before the punch line, the operator decides to move to 147.18 simplex. If you want to hear the rest of the joke, you'll have to either turn on the dome light in your car so you can see where to set the thumbwheel frequency-selection switches on your rig or you can set the switches to a common starting point and count the clicks as you advance them to the correct frequency. By that time, though, the joke is past history. Having owned a HW2036 for a year now, I found that the only real problem was quickly changing frequency at night. There are no lights provided on the front of the panel to illuminate the thumbwheel switches. After many hours of fumbling in the dark with the frequency switches, it occurred to me that the switches must provide binary-coded information to program the synthesizer variable dividers. Why not use that same information to drive some LEDs, thus displaying the selected frequency?

By **Bill Stephens, WB8TJL**, 217 West Reed Street, Bowling Green, Ohio 43402

circuit details

Switches S3, S4, and S5 provide BCD information to the variable divider ICs (U401, U402, U403 respectively). If you are setting your frequency to 146.94, for example, you dial up 6 9 4, since the 1 and 4 remain constant across the band.

I found it was possible to tap off of the four leads at the rear of the three thumbwheel switches and use that information as inputs to the three 7447 ICs (see

didn't find the prospect of wiring 42 resistors too appealing. Therefore, I decided to limit the current going to the common-anode connection of each LED; that required only six resistors.

The only drawback to this technique is that as different digits are displayed, using more or less segments, the overall intensity of that particular digit will be slightly brighter or dimmer than other digits. For instance, if the digit 1 is displayed, only two segments of the LED are lit (segments *b* and *c*). On

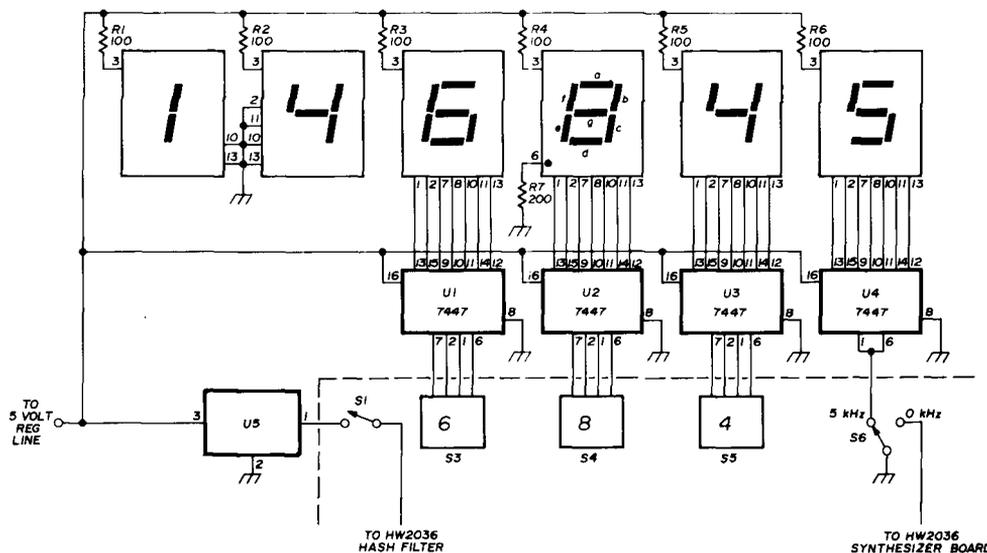


fig. 1. Schematic diagram of the addition to the HW2036 to add an outboard frequency display. Instead of using 7 resistors for each LED, a common current-limiting resistor is inserted in the anode line. The LEDs must be of the common-anode type. The 5-volt regulator is a 7805.

fig. 1). The 7447s convert the BCD data from the switches into a seven-segment format necessary to drive the LEDs. This in no way alters the operation of the synthesizer, nor does it degrade the performance of the rig. All of the outboard components can be easily housed in a small box and set on top of or beside the HW2036. A flat cable containing 15 leads is used to connect the HW2036 and the outboard unit.

display LEDs

Since LEDs 1 and 2 always displayed the digits 1 and 4, it was decided to conserve costs and space and not use a 7447 decoder/driver. Instead, they are permanently wired to display 1 and 4. LEDs 3 thru 6 are driven by U1 through 4, respectively. Each of the LEDs has a 100-ohm 1/2-watt resistor in series with pin 3 and the 5-volt line. This resistor limits the current to each segment to no more than 25 mA. It would be better to connect one resistor in series with each of the seven leads of each of the six LEDs, but I

the other hand, if 8 is displayed, all seven segments (*a*, *b*, *c*, *d*, *e*, *f*, *g*) are used and they must share the same amount of current as the two segments in the first example. With a 100-ohm resistor and a 5-volt line this works out to 25 mA per segment when two segments are on and 7 mA per segment when all seven are lit. In practice this poses no real problem because the change in intensity is barely noticeable.

obtaining power

Referring to the schematic of the HW2036, 12 volts is obtained from lug 2 of the on-off switch S1. This line and a chassis ground are fed thru a cable to the outboard display unit's 5-volt regulator. An LM309K could be used, although the smaller package TO-220 7805 was used in my unit. The 7805 is rated at one ampere of current and is capable of powering the four 7447s and the six display LEDs, although it does run a bit warm at times. Mounting the regulator on a small heatsink or bolting it to the display unit case should overcome this problem.

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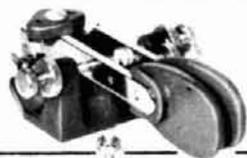
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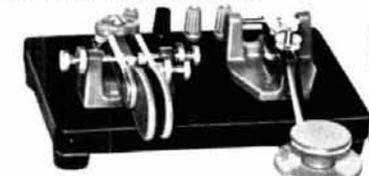
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decoder/drivers

Each switch has a 1, 2, 4, and 8 lead coming from it; these are connected to pins 7, 1, 2, and 6, respectively, of the 7447s. When a thumbwheel switch is set to program the digit 6, it grounds the two unused leads of the 1 and 8 lines. With pins 6 and 7 grounded, the outputs ground the appropriate leads of the associated LED, causing it to form the digit 6. Each of the 7447s is connected to its associated LED by seven leads, one for each of the seven segments in the LED.

If the operator decides to program in a split channel, for instance 146.945, S6 is used to control U4. Within the rig, S6 grounds one lead of U405 when the frequency ends in a zero; it lifts that same lead above ground when a split channel is selected. To display the digit 5, when the switch is set in the 5 kHz position, bring a lead off of the unused side of S6 and tie it to pins 1 and 6 of U4. Thus, when the switch is in the 0 position there is no input to U4 and no digit appears. When the switch is thrown to the 5 kHz position, pins 1 and 6 are grounded, displaying a 5.

construction details

I built my unit on perf board with 2.5 mm (0.1 inch) hole spacings. One board, measuring about 9 cm² (3-1/2 inch square) contains the four 7447s and the voltage regulator. A second board, measuring about 9 cm x 3.8 cm (3-1/2 inches x 1-1/2 inches), contains the six LEDs and is connected to the first board by 30 jumpers. While perf board is entirely acceptable for this project, it is suggested that printed-circuit techniques be used due to the relatively large number of board-to-board jumpers and the small space in which to work. The many connecting jumpers tend to make the interconnecting harness very stiff and difficult to work with.

My unit is housed in a homebuilt box measuring 5 x 9 x 10.2 cm (2 x 3-1/2 x 4 inches) with a sloping front panel for easier viewing of the LEDs. The size and shape of your box will be determined by where and how you are going to mount it. Four-conductor flat cable was used throughout the construction. Four lengths of cable were used to bring the 15 leads out of the back of the HW2036 case (they fit nicely between the top of the PA board assembly and the case). Individual strands of the same wire were stripped and used for the 5-volt line and ground bus on the two perf boards.

If the unit is to be mounted in an outboard case, it is recommended that a plug be installed to disconnect the outboard unit from the HW2036 for ease of servicing.

I wish to thank WB8NQW and K8TT for their help and thoughts in completing this project.

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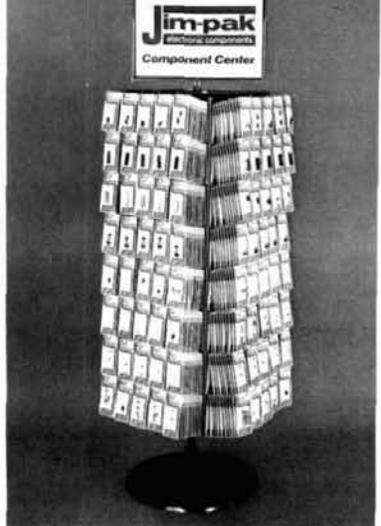
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phase-locked loops

basic building blocks for frequency manipulation

A basic discussion of phase-locked loops and how they are used in communications systems

In **electronic systems**, information can be expressed in four ways: voltage, current, frequency, and phase angle. In modern electronics, operational amplifiers have become the basic building blocks for circuits which manipulate voltages and currents. But what about frequency and phase angle? Enter the phase-locked loop or PLL.

The first widespread use of the phase-lock system was in TV receivers to synchronize the horizontal and vertical sweep oscillators to the transmitted sync pulses. Lately, narrowband phase-locked receivers have proved to be of considerable benefit in tracking weak satellite signals. This is due to the superior noise immunity of PLL systems. Although it's not well known, the synchronous reception of radio signals using the PLL technique was first described in the early 1930s; it was known as a "homodyne" receiver.

In the early days, applications using PLLs had to be implemented using discrete components. Even after the advent of transistors, the PLL circuit was considerably complex. Thus, the use of PLL methods in most electronic systems was both expensive and impractical.

In the late 1960s Signetics Corporation developed monolithic circuit versions of the PLL system. This

development of single chip PLLs changed things considerably. A single package device, used with a few external components, offers all the benefits of PLL operation while making their use practical, uncomplicated, and economical.

PLL theory

Just what is a PLL and how does it do all this frequency manipulating? The op amp is a voltage/current feedback system; that is, a portion of the output is fed back to the input. In an op amp circuit, this feedback component is a current. The PLL is a feedback system but in this case the component fed back is a frequency. **Fig. 1** shows a block diagram of a feedback system.

A phase-locked loop is basically an electronic servo loop. The function of a PLL is to detect and track small differences in phase and frequency existing between the input and a reference signal. A block diagram of a basic PLL system is shown in **fig. 2**. In this circuit the voltage-controlled oscillator is driven in the direction that will minimize the error signal. Note the similarities between **figs. 1 and 2**.

Like most other complex circuits, phase-locked loops have special terms associated with them; understanding their operation is easier when you become familiar with the language. The following is a brief glossary of terms encountered with PLLs:

Capture range. The range of frequencies over which the loop can detect a signal on the input and respond to it. This is sometimes called the lock-in range (lock-in range refers to how close the signal must be to the center frequency before acquisition can occur; thus it is one-half the capture range).

Current controlled oscillator. An oscillator in which the frequency is determined by an applied current.

Damping factor. In a PLL this refers to the ability of a loop to respond quickly to an input frequency step without excessive overshoot.

Free-run frequency (f_o). Also called the center frequency; it is the frequency of the vco with no input signal.

By **Bob Marshall, WB6FOC**, Analog Applications, Signetics Corporation, Post Office Box 9052, Sunnyvale, California 94086

Lock range. The range of frequencies over which the loop will remain in lock; also called tracking range.

Loop gain (K_v). Product of the dc gains of all the loop elements; in units of sec^{-1} .

Loop noise bandwidth. A loop relating to damping and natural frequency which describes the effective bandwidth of the input signal.

Lowpass filter. A filter which permits only dc and low frequencies to travel around the loop; it determines the capture range of the loop.

Natural frequency. The characteristic frequency of the loop (not to be confused with free-running frequency).

Phase detector gain factor (K_d). The conversion factor between the phase detector output voltage and the phase differences of the input and vco signals; expressed in volts-radians.

Phase detector. A circuit which compares the relative phase between two inputs and produces an error voltage dependent on the difference. This error voltage corrects the vco frequency during tracking. Sometimes called a phase comparator or mixer.

Quadrature phase detector. A phase detector operated in quadrature (90° out of phase) with the loop detector.

VCO conversion gain (K_o). Conversion factor between vco frequency and control voltage in radians/sec/volt.

Voltage Controlled Oscillator or VCO. An oscillator whose frequency is determined by an applied control voltage.

loop operation

As was mentioned earlier, the PLL is a feedback system; therefore, it can be characterized mathematically

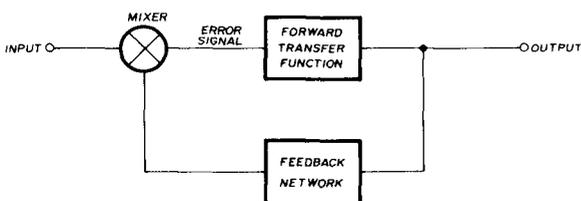


fig. 1. Block diagram of feedback system shows the basic arrangement of both operational amplifiers and phase-locked loops. Current feedback is used in an op amp whereas a frequency component is fed back in a phase-locked loop.

ically by the same equations that apply to other, more conventional feedback systems. However, the parameters in the PLL equations deal with phase rather than a current or voltage.

A mathematical analysis of a PLL can get pretty hairy but a qualitative analysis will explain the basic principle of PLL operation. During the following discussion, it will be helpful to refer to fig. 3.

With no input signal applied, the error voltage V_d is zero. The vco will operate at a set frequency f_o or the free-run frequency. When an input signal is applied to the system, the phase detector compares the phase and frequency of the input with the vco frequency. This generates an error voltage $V_e(t)$ that is related to the phase and frequency difference between the two signals; this error voltage is then filtered, amplified, and applied to the control terminal of the vco. In this manner, the control voltage $V_d(t)$ forces the vco frequency to vary in a direction that reduces the frequency difference between f_o and the input signal.

If the input frequency f_i is sufficiently close to f_o , the feedback nature of the PLL causes the vco to synchronize or lock with the incoming signal. Once in lock, the vco frequency is identical to the input signal except for a finite phase difference. This net phase difference θ_o is necessary to generate the corrective error voltage V_d to shift the vco frequency from its free-running value to the input signal and thus, keep the PLL in lock. This self correcting ability of the system allows the PLL to track frequency changes of the input signal once it is locked.

Another way of describing the operation of the PLL is to observe that the phase detector is, in actuality, a multiplier circuit that mixes the input signal with the vco signal. The mixer produces the sum and difference frequencies ($f_i \pm f_o$). When the loop is in lock, the vco duplicates the input frequency so that the difference frequency component ($f_i - f_o$) is zero; hence the output of the phase comparator contains a dc component. The lowpass filter removes the sum frequency component ($f_i + f_o$) but passes the dc component which is amplified and fed back to the vco. Notice that with the loop in lock, the difference frequency component is dc and independent of the band edge of the lowpass filter.

lock and capture

What happens before the loop is locked? Let's assume for a moment that there is a frequency on the input to a PLL. The phase comparator mixes this incoming frequency with the free running vco frequency. If the difference frequency ($f_L - f_o$) is greater than the band edge of the lowpass filter, the input to the vco is still zero so the vco remains at its free-run fre-

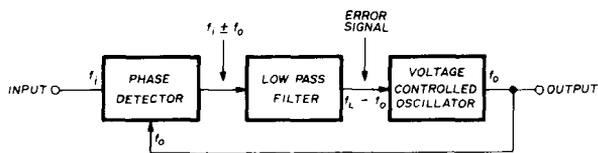


fig. 2. Basic phase-locked loop consists of a phase detector, lowpass filter, and voltage-controlled oscillator (vco). The vco is driven in a direction which minimizes the error signal.

quency. As the input frequency approaches that of the vco, $f_i - f_o$ decreases and approaches the band edge of the lowpass filter. Now some of the difference component is passed to the vco control. This in turn decreases the frequency difference component which allows more information through the filter. This positive feedback mechanism causes the vco to snap into lock with the input signal. Thus, the capture range is again defined as the "frequency range centered about the vco free-run frequency over which the loop can acquire lock."

Once the loop is locked, $f_i - f_o$ is essentially dc and thus unaffected by the lowpass filter. The lock range is limited by the range of the error voltage that can be generated and the corresponding deviation in vco frequency which is produced.

It is important to distinguish the difference between *capture* range and *lock* range. Lock range is defined as the frequency range, usually centered about the initial vco frequency, over which the loop can track the input signal once lock has been achieved. The total time required for the vco to obtain a locked condition is called the pull-in time. Pull-in time depends on the initial frequency and phase difference between the two signals and the overall loop gain as well as the lowpass filter.

effects of the lowpass filter

In the operation of the loop, the lowpass filter serves a dual function. First, by attenuating the high-frequency error component ($f_L + f_o$) at the output of the phase comparator, it improves the interference-rejection characteristics; second, it provides a short term memory for the PLL and ensures a rapid recapture of the signal if the system is thrown out of lock because of a noise transient. Reducing the lowpass filter bandwidth has the following effects on system performance:

1. The capture process becomes slower and increases the pull-in time.
2. Capture range decreases.
3. Interference-rejection properties improve since the error voltage caused by an interfering signal is attenuated further by the lowpass filter.

4. The transient response of the loop (the response of the PLL to sudden changes of the input frequency within the capture range) becomes undamped.

This last effect also produces a practical limitation on the lowpass loop filter's bandwidth and roll-off characteristics from a stability standpoint. A detailed analysis of a PLL under lock condition using Laplace transforms will prove that if either the loop gain or the filter time constant is too large, the loop itself will break into sustained oscillations.

The lock range of the PLL, f_L , can be shown to be numerically equal to the dc loop gain K_v

$$4\pi f_L = 2\omega_L = 2K_v$$

Since the capture range, f_c , denotes a transient condition, it is more difficult to derive, but with a simple lag filter the capture range can be approximated as

$$4\pi f_c = 2\omega_c \approx \sqrt{\frac{2\pi f_L}{\tau_I}} = \sqrt{\frac{K_v}{\tau_I}}$$

where τ_I is the time constant of the loop and f_L is the lock frequency. Thus, the capture range increases as the time constant of the filter decreases, while the lock range is a function of the dc loop gain.

Fig. 4 shows the typical frequency-to-voltage transfer characteristics of the PLL. The input is assumed to be a sine wave whose frequency is swept slowly over a broad range of frequencies. The vertical scale is the corresponding error voltage of the loop.

Fig. 4A shows the loop error voltage with an increasing frequency. The loop does not respond until the frequency reaches f_1 , which corresponds to the lower edge of the capture range. At that time the loop suddenly locks on the input and causes a negative jump of the loop error voltage. As the frequency continues to increase, the loop error voltage increases. Notice that V_d is zero when the incoming signal f_i equals f_o . The loop continues to track the incoming signal until the frequency equals f_2 ; this corresponds to the upper edge of the lock range. The PLL then loses lock and V_d returns to zero.

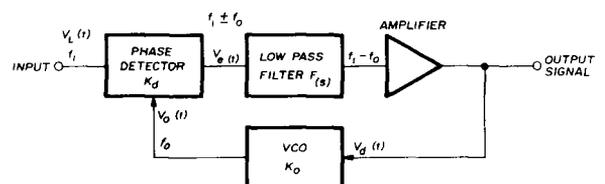


fig. 3. Diagram of a phase-locked loop showing the frequency and voltage relationships in the system. See text for description of operation.

The slope of V_d is equal to the reciprocal of the vco gain ($1/K_o$) measured in volts/radians/sec. If the input frequency is swept slowly back (illustrated by fig. 4B), nothing happens until the incoming frequency reaches f_3 . The loop continues to track until f_4 where it breaks lock and the error returns to zero. The total lock and capture ranges of the system are:

$$2f_L = f_2 - f_4 \text{ (lock)}$$

$$2f_c = f_3 - f_1 \text{ (capture)}$$

As indicated by the transfer characteristics of fig. 4, the PLL has an inherent selectivity about the center frequency set by the free running vco frequency f_o . It can also be seen that the loop will respond to input signal frequencies that are separated from f_o or f_L depending on whether the loop starts with or without an initial lock condition. The linearity of the frequency-to-voltage conversion characteristics for the PLL is determined solely by the vco conversion gain. For most PLL applications, the vco is required to have a highly linear voltage-to-frequency transfer characteristic.

functional applications

Now that you are familiar with what a PLL is, you probably wonder what it all means and what it can do for you. As a functional building block the PLL is suitable for a wide variety of frequency related applications. These applications generally fall into one or more of the following categories: fm demodulation, frequency synthesis, frequency synchronization, signal conditioning, a-m demodulation, and frequency modulation.

fm demodulation

If the PLL is locked to an fm signal, the vco will track the instantaneous frequency of the input signal. The filtered error voltage V_d , which forces the vco to maintain lock with the input signal, then becomes the demodulated fm output. The linearity of the vco's voltage-to-frequency transfer characteristic determines the linearity of the demodulated signal. Some typical fm demodulation applications are discussed below.

Broadcast fm detection. In this application, the PLL can be used as a complete fm i-f strip, limiter, and fm detector. It can be used to detect wide- or narrow-band fm signals with greater linearity than can be obtained by other means for frequencies within the range of the vco (presently up to about 50 MHz). One increasingly popular use of the PLL is in scanning receivers where a number of broadcast channels may be sequentially monitored by simply

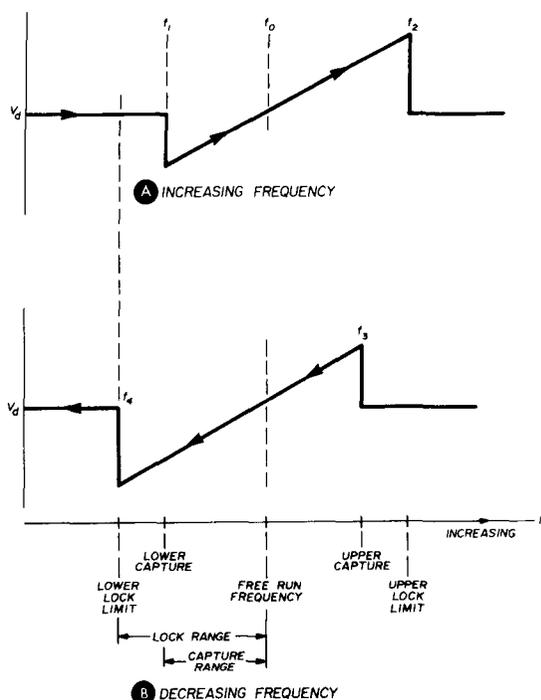


fig. 4. Frequency-to-voltage transfer characteristics of a phase-locked loop for increasing input frequency, A, and decreasing input frequency, B.

varying the vco free-run frequency. (Scanning receivers are also using a digital PLL technique which is, in principle, similar to linear PLLs).

Fm telemetry. This involves demodulation of a frequency-modulated subcarrier. One example is the use of the PLL to recover the SCA (storecast music) signal from the combined signal of commercial fm broadcast stations. The SCA signal is a 67-kHz fm subcarrier which puts it above the frequency spectrum of the normal stereo or monaural program material.

Frequency-shift keying (FSK). This is essentially digital fm. Frequency-shift keying is a means for transmitting digital information by a carrier which is shifted between two discrete frequencies (as in RTTY, for example). In this case, the two discrete frequencies correspond to a digital 1 (mark) and a digital 0 (space), respectively. When the FSK signal is connected to a PLL, the demodulated output (error voltage) shifts between two discrete voltage levels which correspond to the demodulated binary output.

frequency synthesizer

Frequency multiplication can be achieved with the PLL in two ways: locking to a harmonic of the input signal, or insertion of a counter (digital frequency divider) in the loop. Harmonic locking is the simpler

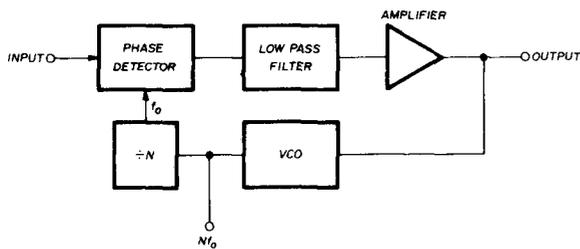


fig. 5. Phase-locked loop frequency synthesizer using a digital frequency divider.

and is achieved by setting the vco free-run frequency to a multiple of the input frequency and allowing the PLL to lock. A limitation of this scheme, however, is that the lock range decreases as successively higher and weaker harmonics are used for locking. This limits the practical harmonic locking range. For large multiples, the use of a digital frequency divider provides better results; the basic arrangement is shown in fig. 5. The loop is broken between the vco and the phase detector and a counter is inserted. In this case, the fundamental of the divided vco frequency is locked to the input frequency so the vco is actually running at a multiple of the input frequency. The amount of multiplication is determined by the counter.

In frequency multiplication applications, it is important to remember that the phase comparator is actually a mixer and its output contains the sum and difference frequency components. The difference frequency component is dc and is the error voltage which drives the vco to keep the loop in lock. The sum frequency components (basically twice the frequency of the input), if not well filtered, will induce incidental fm on the vco output. This happens because the vco is running at many times the input frequency. The sum frequency component appearing at the control voltage input of the vco causes a periodic variation of its frequency about the desired multiple. For frequency multiplication, it is generally necessary to filter quite heavily to remove this sum component. The tradeoff is reduced capture range and a more underdamped loop transient. The size of the loop filter limits the minimum input frequency.

frequency synchronization

Using a PLL system, the frequency of a less precise vco can be phase locked with a low level but highly stable reference signal. The vco output reproduces the reference signal at the same per unit accuracy but at a much higher power level. In some applications, the synchronizing signal can be a low duty cycle burst at a specific frequency. The PLL can be used to regenerate a coherent CW reference fre-

quency by locking onto the short synchronizing pulse. An example of this is phase-locked chroma-reference generators in color television receivers.

In digital systems the PLL can be used for a variety of synchronization functions. For example, two clocks can be phase locked to each other so that one can function as the backup for the other. Other popular applications include locking to National Bureau of Standards' station WWVB to generate an inexpensive laboratory frequency standard.

signal conditioning

By selecting the proper vco free-run frequency, the PLL can be made to lock to any number of signals present at the input. The vco output tracks that frequency while attenuating the undesired frequencies of sidebands present at the input. If the loop bandwidth is sufficiently narrow, the signal-to-noise ratio at the vco is better than that at the input.

a-m demodulation

A-m demodulation can be achieved with a PLL by using the scheme shown in fig. 6. In this mode of operation the PLL functions as a synchronous a-m detector. The PLL locks onto the carrier of the a-m signal so the vco output has the same frequency as the carrier, but with no amplitude modulation. The vco will track the input but with a 90° phase shift; if the input is now sent through a 90° phase-shift network and fed into a multiplier, the output of the second multiplier will be directly proportional to the amplitude of the input signal. The PLL still exhibits the capture phenomenon; thus the loop maintains a high degree of selectivity centered about the free-run vco frequency. Since this method is essentially a coherent detection technique which involves the averaging of two compounded signals, it offers a higher degree of noise immunity than a conventional peak-detector type a-m demodulator.

fm modulation

Since the PLL has a voltage-controlled oscillator, it's possible to inject a signal into the loop and cause the vco to change frequency. This signal can be injected at the lowpass filter or across the timing component; the per cent of modulation is controlled by

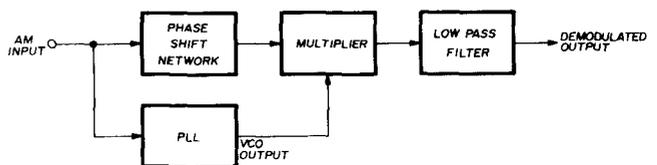


fig. 6. Coherent a-m detector circuit using a phase-locked loop.

the amplitude of the injected signal — linearity is a function of the voltage-to-frequency transfer characteristics of the vco.

PLL design considerations

Many integrated circuits use phase-locked loops along with specialized circuitry to form specialized circuit functions. The versatility of these special systems has been sacrificed for convenience and circuit optimization. Circuits such as the μ A758 Stereo Multiplex Decoder, TDA2541 Video I-F System, and others have built-in PLLs which have been optimized

expected input frequency range. Since the loop's capture ability is a function of the difference between the incoming and free-running frequencies, the band edges of the capture range are always centered about the free-run frequency. Typically the lock range is also centered about the free-run frequency. By choosing a center frequency offset from the incoming signal, the detection or tracking range of the loop is limited to one side. This permits rejection of an adjacent higher (or lower) frequency signal and still permits wideband operation (narrowband operation reduces tracking speed).

table 1. User's guide to phase-locked loop ICs.

	upper freq (MHz)	max lock range (%F _o)	fm distor- tion	output swing ± 5% deviation (volts p-p)	center frequency stability (ppm/°C)	frequency drift/w supply voltage (%/V)	input resist- ance	a-m output avail	typical supply current (mA)	supply voltage range (volts)
NE560	30	40%	0.3%	1	± 600	0.3	2k†	no	9	+ 16 to + 26
NE561	30	40%	0.3%	1	± 600	0.3	2k†	yes	10	+ 16 to + 26
NE562	30	40%	0.5%	1	± 600	0.3	2k†	no	12	+ 16 to + 30
NE564	50	30%			± 200					+ 5 to + 10
NE565	0.5	120%	0.2%	0.15	± 200	0.16	5k†	no	8	± 6 to ± 12
SE565	0.5	120%	0.2%	0.15	± 100	0.08	5k†	no	8	± 6 to ± 12
NE567	0.5	14%	5%*	0.20	35 ± 60	0.7	20k†	yes*	7	+ 4.75 to + 9
SE567	0.5	14%	5%*	0.20	35 ± 60	0.5	20k†	yes*	6	+ 4.75 to + 9
NE566	0.5	N/A	0.2%	30%/V§	± 200	0.16			7	+ 12 to + 26
SE566	0.5	N/A	0.2%	30%/V§	± 100	0.08			7	+ 12 to + 16

* A-m and fm outputs are available, but are not optimized for linear demodulation.

† Input biased internally.

§ Figure shown is vco gain in per cent deviation per volt.

to perform best in those applications. But there are also PLLs which can be used as basic building blocks; these offer the most flexibility in circuit design. Signetics offers three basic classes of single-chip PLL circuits: the general purpose PLL, a PLL with an added multiplier, and the PLL tone decoder. National Semiconductor has a general-purpose PLL and a tone decoder. A more complete list of the PLLs available from various manufacturers is detailed in table 1.

To obtain the optimum performance from a PLL circuit it is important that the user become familiar with the tradeoffs that can be made. To be more specific, the following discussion will be directed at the 560, 561, 562, 564, 565, 566, and 567 phase-locked loops. The tradeoffs and loop conditions, however, will hold true for all basic PLLs. Generally speaking, the user is free to select the frequency, tracking or lock range, capture range, and input amplitude.

center frequency selection

Setting the center frequency is accomplished by selecting one or two external components. This free-running frequency is usually set in the center of the

As was mentioned earlier, the loop uses a multiplier in which the input signal is multiplied by a unity square wave at the vco frequency. The odd harmonics present in the square wave permit the loop to lock to input signals at these odd harmonics. Thus the center frequency may be set to 3 times or 5 times the input signal. The tracking range will, however, be considerably reduced as higher harmonics are used.

It should also be noted that the PLL can lock onto the harmonic of the desired signal. If this is not acceptable, steps must be taken (such as prefiltering) to prevent undesired lockup of the loop to harmonically related signals.

In evaluating the loop for a specific application, compute the magnitude of the expected signal component nearest f_o . This magnitude can be used to estimate the lock and capture range.

The PLLs are stabilized against center frequency drift due to power supply variations and the 565 and 567 are temperature compensated over the full military temperature range (-55 to 125°C). All of the loops are affected by external components which must have equal (or better) stability over the desired operating temperature range.

Two things limit the lock or tracking range of a PLL. First, any vco can only swing so far; if the input signal frequency goes beyond this limit, lock will be lost. Second, the voltage developed by the phase detector is proportional to the product of both the phase and the amplitude of the in-band component to which the loop is locked. If the signal amplitude decreases, the phase difference between the signal and the vco must increase to maintain the same out-

the lowpass filter should have a high cutoff frequency. However, a lowpass filter with a high cutoff frequency will attenuate the sum frequencies to a lesser extent so the output contains a significant and often bothersome signal at twice the input frequency. (Remember that the multiplier forms both the sum and difference frequencies; during lock the difference frequency is zero, but the sum frequency at twice the locked frequency is still present.) If necessary, the

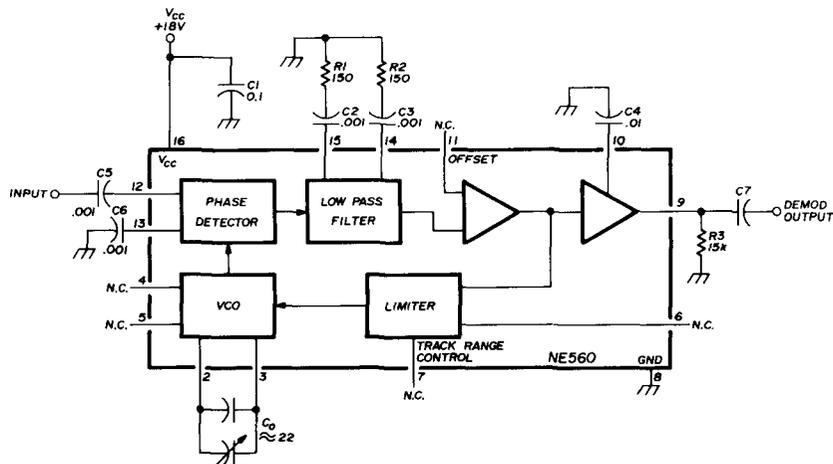


fig. 7. 10.7-MHz fm demodulator using an NE560 PLL IC. Design details are given in the text.

put voltage and, hence, the same frequency deviation.

It often happens with low input amplitudes that even the full $\pm 90^\circ$ phase range of the phase detector cannot generate enough voltage to permit tracking over wide deviations. When this occurs, the effective lock range is reduced. Therefore, if the input signal is weak, you must give up some tracking capability and accept greater phase errors. Conversely, a strong input signal will allow you to use the entire vco swing capability and keep the vco phase (referred to the input signal) very close to 90° throughout the range.

Note that tracking range does not depend on the lowpass filter. If a lowpass filter is in the loop, however, it will have the effect of limiting the maximum rate at which tracking can occur. Obviously, the voltage across the lowpass filter capacitor cannot change instantly, so lock may be lost when sufficiently large step changes occur. Between the constant frequency input and the step-change frequency input is some limiting frequency slew rate at which lock is just barely maintained. When tracking at this rate, the phase difference is at its limit of 0° to 180° . It can be seen that if the lowpass filter's cutoff frequency is low, the loop will be unable to track as fast as when the lowpass filter's cutoff frequency is higher. Thus, when maximum tracking rate is needed,

unwanted sum frequency component can be filtered out with an external lowpass filter.

capture range control

There are two main reasons for making the lowpass filter time constant large. First, a large time constant provides an increased memory effect in the loop so that it remains at or near the operating frequency during momentary fading or loss of signal. Second, the large time constant integrates the phase detector output so that increased immunity to noise and out-band signals is obtained.

In addition to the lower tracking rates of large loop filters, other penalties must be paid for the benefits gained; the capture range is reduced and the capture transient becomes longer. Reduction of capture range occurs because the loop must use the magnitude of the difference frequency component at the phase detector to drive the vco toward the input frequency. If the cutoff frequency of the lowpass filter is low, the difference component amplitude is reduced and the loop cannot swing as far. Thus, capture range is reduced.

choice of input level

Whenever amplitude limiting of the in-band signal occurs, whether in the loop input stages or prior to

the input, the tracking (lock) and capture ranges become independent of signal amplitude.

Better noise and out-band signal immunity is achieved when the input levels are below the limiting threshold since the input stage is in its linear region and the creation of cross-modulation components is reduced. Higher input levels will allow somewhat faster operation due to greater phase detector gain and will result in a lock range which becomes constant with amplitude as the phase detector gain becomes constant. Also, high input levels will result in a linear phase-versus-frequency characteristic.

lock-up time and tracking speed control

In tracking applications, lock-up time normally has little consequence, but occasions do arise when it is desirable to keep lock-up time short to minimize data loss when noise or extraneous signals drive the loop out of lock. Lock-up time is of great importance in tone decoder type applications. Tracking speed is important if the loop is used to demodulate an fm signal. Although the following discussion dwells largely on lock-up time, the same comments apply to tracking speed.

No simple expression is available which adequately describes the acquisition or lock-up time. This will be better appreciated after you have reviewed the following factors which influence lock-up time:

1. Input phase
2. Lowpass filter characteristic
3. Loop damping
4. Deviation of input frequency from center frequency
5. In-band input amplitude
6. Out-of-band signals and noise
7. Center frequency

Fortunately, it is usually sufficient to know how to improve the lock-up time and what must be traded off to obtain faster lock-up. Suppose you have set up a loop or tone decoder and find that occasionally the lock-up transient is too long. Remember all the factors that influence lock, what can be done to improve the situation?

1. Initial phase relationship between the incoming signal and the vco; this is the greatest single factor influencing the lock time. If the initial phase is wrong, it first drives the vco frequency away from the input frequency so the vco frequency must walk back on the beat notes. The only way to overcome this variation is to send phase information all the time

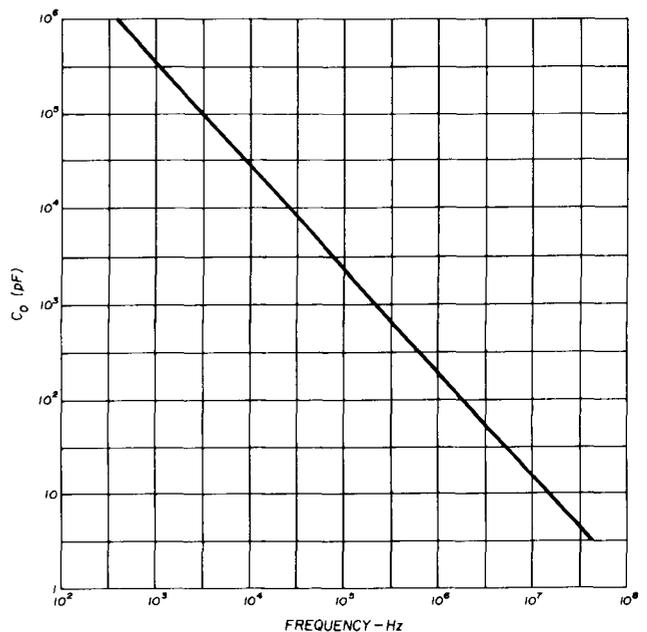


fig. 8. Free-running oscillator frequency as a function of vco timing capacitance.

so that a favorable phase relationship is guaranteed at $t = 0$. Usually, however, the incoming phase cannot be controlled. Using two TTLs with the voltage-controlled oscillators synchronized 90° apart reduces worst-case lock-up time by one-half because the input can never be more than 45° out of phase with one of the loops.

2. Lowpass filter. The larger the time constant of the lowpass filter, the longer the lock-up time. You can reduce the lock-up time by decreasing the filter time constant, but in doing so you sacrifice some of the noise immunity and out-of-band signal rejection which caused you to use a larger filter in the first place. You must also accept a sum frequency (twice the vco frequency) component at the lowpass filter and greater phase jitter resulting from out-of-band signals and noise. In the case of the tone decoder (where control of the capture range is required since it specifies the device bandwidth) a lower value low-pass capacitor automatically increases bandwidth. You gain speed only at the expense of added bandwidth.

3. Loop damping for a simple time constant low-pass filter is:

$$\zeta = \frac{1}{2} \sqrt{\frac{1}{\tau K_v}}$$

Damping can be increased not only by reducing τ , as discussed above, but also by reducing the loop gain K_v . By using the loop gain reduction to control band-

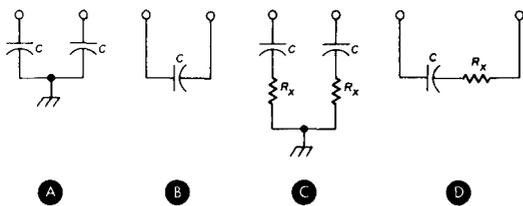


fig. 9. Typical RC lowpass filters. Approximate formulas describing their frequency response are presented in the text.

width or capture and lock range, you achieve better damping for narrow bandwidth operation. The penalty for this damping is that more phase detector output is required for a given deviation so that phase errors are greater and noise immunity is reduced. Also, more input drive may be required for a given deviation.

4. Input frequency deviation from free-running frequency. Naturally, the further an applied input signal is from the free-running frequency of the loop, the longer it will take the loop to reach that frequency because of the charging time of the lowpass filter capacitor. Usually, however, the effect of this frequency deviation is small compared to the variation resulting from the initial phase uncertainty. Where loop damping is very low, however, it may be predominant.

with the reduced phase detector output (see 4 above).

6. Out-band signals and noise. Low levels of extraneous signals and noise have very little effect on the lock-up time, neither improving or degrading it. However, large levels may overdrive the loop input stage so that limiting occurs, at which point the in-band signal starts to be suppressed. The lower effective input level can cause the lock-up time to increase, as discussed in 5 above.

7. Center frequency. Since lock-up time can be described in terms of the number of cycles to lock, fastest lock-up is achieved at higher frequencies. Thus, whenever a system can be operated at a higher frequency, lock will typically be faster. Also, in systems where different frequencies are being detected, on average the higher frequencies will be detected before the lower frequencies. Because of the wide variation due to initial phase, however, the reverse may be true for any single trial.

In all of the above design considerations, it is important to remember that the PLL is a dc loop. Any dc level change injected into the loop will affect its operation. The loop is also sensitive to temperature changes because most voltage-controlled oscillators have a temperature coefficient of around 600 ppm/C°. The resistors and capacitors used in the fre-

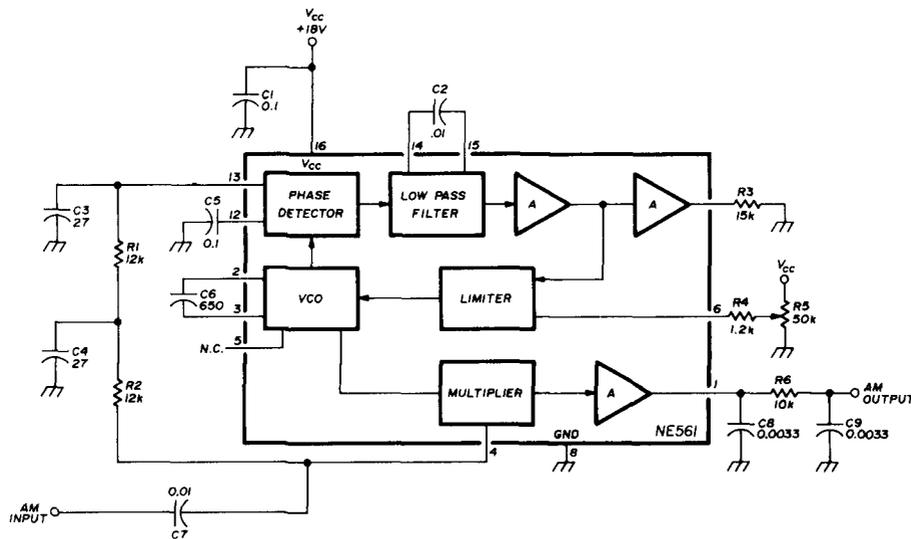


fig. 10. A-m detector for 455-kHz i-f using the NE561 PLL.

5. In-band input amplitude. Since input amplitude is one factor in the phase detector gain K_d , and since K_d is a factor in the loop gain K_{LV} , damping is also a function of input amplitude. When the input amplitude is low, the lock-up time may be limited by the rate at which the lowpass capacitor can charge

frequency determining network also have temperature coefficients which must be considered when designing circuits using the PLL.

design examples

Let's take a look at a practical design example

using the NE560 PLL as a 10.7-MHz fm demodulator (see fig. 7).

Supply voltage. Generally, the operating voltage is determined by the available power supply or the device data sheet. The manufacturers specify the V_{CC} at which the device parameters were measured. This simplifies the V_{CC} selection since the device is tested at an optimum voltage; for the NE560 this voltage is 18 volts. Capacitor C1 is a decoupling capacitor for the supply and should be located as close as possible to the V_{CC} pin of the IC.

VCO free-run frequency. Since this is a 10.7-MHz detector, the approximate timing capacitance can be found on the data sheet graph. This graph (fig. 8) shows that the timing capacitor should be about 22 pF. An approximate value for the timing capacitor can be found from:

$$C_o \cong \frac{300}{f_o}$$

where C_o is in pF and f_o is in MHz. Using this formula, the capacitor is calculated to be about 28 pF. The design example uses a 22-pF capacitor in parallel with an 8-pF trim capacitor for fine tuning.

Lowpass filter. The output of the phase detector is the sum and the difference of the input fm signal and the vco frequency. The loop filter must remove the sum component. Because the modulation on the incoming signal contains the information desired, it is necessary to retain the difference frequency. In addition, the attenuation of the high-frequency component increases the interference rejection characteristics. To maintain loop stability at all signal levels, the loop cannot cause more than 12 dB per octave rolloff.

Fig. 9 shows several lowpass filter configurations. The circuits of figs. 9A and 9B can provide 6 dB per octave rolloff at the desired bandwidth frequency; resistor R_x in 9C and 9D will break the response up at higher frequencies. R_x is typically between 50 and 200 ohms. Because of the complexities of the transfer functions, which many designers use to characterize these filters, it is usually easier to use approximation formulas for the lowpass filter. For fig. 9A and 9B the formula is

$$C = \frac{26.6}{f} \mu F$$

where f is the desired cutoff frequency. The lowpass filters of fig. 9C and 9D can be approximated by

$$C = \frac{13.3}{f} \mu F$$

At frequencies greater than 5 MHz, the filters of fig. 9C and 9D will provide greater loop stability.

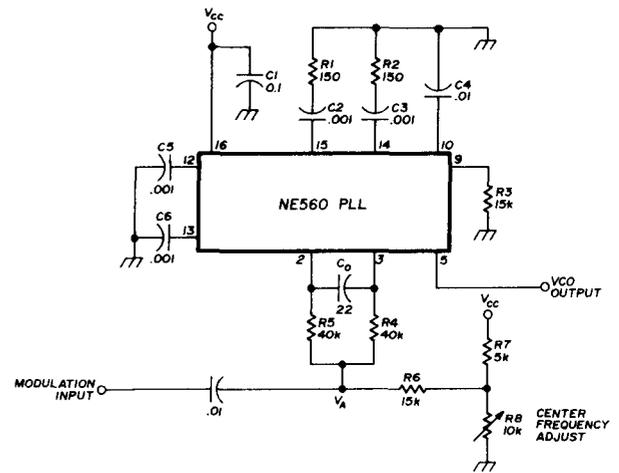


fig. 11. 10.7-MHz fm modulator using the NE560 PLL. Resistor R8 provides $\pm 10\%$ variation of the center frequency. For best waveform linearity resistors R4 and R5 should be equal; minimum value for R4 and R5 is 20k.

Assuming that the desired frequency bandwidth of the demodulated signal is 15 kHz, the filter capacitor values (C2 and C3 in fig. 7) are

$$C = \frac{13.3 \times 10^{-6}}{15 \times 10^3} = 886 \text{ pF}$$

or approximately 0.001 μF

Resistors R1 and R2 were selected to be between 50 and 200 ohms; I used 150-ohm resistors.

De-emphasis filter. Standard fm broadcast stations use a de-emphasis time constant of approximately 75 microseconds. According to the NE560 data sheet, the internal resistance into pin 10 is 8k. The 3 dB roll-off frequency is given by

$$f_{3dB} = \frac{1}{2\pi R_D C_D} \text{ where } R_D \text{ is } 8000$$

For a time constant of 75 microseconds

$$C_D = \frac{1}{2\pi f R_D} = \frac{75 \times 10^{-6}}{R_D} = 0.0094 \mu F$$

In practice, we can use a 0.01 μF capacitor since the internal resistance has a ± 20 per cent tolerance.

Output. The output, pin 9, has a 15k dc load and ac coupled output. The detected output amplitude for 75 kHz deviation, according to the data sheet, is 30 mV minimum. The demodulated output is emitter coupled and requires a dc path to ground.

Input level. The input level required for a constant tracking range is 2 mV. Also, the a-m rejection char-

table 2. Dedicated ICs which use phase-locked loops.

function	part number	manufacturer
FSK modulator/ demodulator	XR210	Exar
	HCTR0340	Hughes
CB synthesizer	MC15104	Motorola
	MM55104	National
	NE575	Signetics
	SP8920	Plessey
TV synthesizer	NC6410	Nitron
Tone decoder	NE567	Signetics
	LM567	National
	XR567	Exar
Stereo decoder	μ A758	Signetics
	LM1310	National
	LM1800	National
VCO function generator	NE566	Signetics
	LM566	National
	11C58C	Fairchild
	MC4024	Motorola

acteristic for high input signal levels is reduced for signals greater than 30 mV, so the input signal should be between 2 and 20 mV.

synchronous a-m detector

Because the NE561 has a multiplier incorporated on the chip, it is possible to accomplish synchronous a-m detection; fig. 10 shows the circuit. The vco tracks the input to the phase detector with a 90° phase shift. By shifting the input signal 90° the vco and the input become in phase. These signals are mixed in the multiplier — the output is the a-m signal (due to the amplitude difference of the input, the constant amplitude of the vco, and the 90° phase difference).

To accomplish this phase shift of the input, a simple RC lag network is used. The values of R1, C3, R2, and C4 are calculated using the relationship

$$R1C3 = R2C4 = \frac{1}{2\pi f_o} \text{ (assume } R1 = 12k\text{)}$$

$$\text{Thus } C3 = C4 = \frac{1}{2\pi \times 55 \times 10^3 \times 12 \times 10^3} = 29 \text{ pF}$$

(Use a standard value, 27 pF.)

The timing capacitor is calculated as

$$C_6 \approx \frac{300}{f_o} \text{ pF (} f_o \text{ in MHz)}$$

$C_6 \approx 660 \text{ pF}$ (use a standard value, 680 pF)

By controlling the current into pin 6 the vco can be fine tuned so the absolute value of C6 is not critical. By injecting or loading current into pin 6, the vco frequency can be adjusted ± 20 per cent (with 4 mA

control current). Since no information is taken from the lowpass filter, the filter capacitor, C2, is not critical; the value is selected to maintain loop stability. I used a 0.01 μ F for C2 which worked quite well.

Capacitors C8, C9, and resistor R_c form a post a-m filter to eliminate carrier feedthrough; R3 is optional because that output is not used.

frequency modulation

Since the PLL IC contains a vco, frequency modulation is accomplished by summing a modulating signal with the error signal or by changing the dc level to control the vco at the modulating rate. Referring to fig. 10, if the current into pin 6 becomes a variable current, the vco will change frequency at the modulated rate. The vco output is taken at pin 5.

Another method of changing the frequency is to insert offset currents across the timing capacitor. To build a 10.7-MHz fm modulator, the circuit of fig. 7 can be modified as shown in fig. 11. Note that the input is ac grounded and the vco output is taken from pin 5. Frequency deviation can be approximated by the formula given below.

$$f = f_o \left[1 - \frac{V_A - 6.4}{1300 R} \right]$$

By incorporating a switching arrangement the circuit can be used as both an fm demodulator and modulator.

summary

In addition to the basic PLL, many manufacturers have ICs for dedicated applications which incorporate PLLs into the design. For example, the CA3089 is an fm i-f strip which contains all the functions of an i-f amplifier and audio detector. Because these parts are dedicated to a specific application, however, they lose flexibility. Table 2 is a partial list of parts which use PLLs in dedicated applications. This list is by no means complete, but is an indication of the various types of circuits which are presently available.

In summary, the PLL is a versatile building block for use in frequency manipulating circuits. By careful design, an awareness of the PLL's limitations, and knowledge of the design tradeoffs, you will find them as easy to use as operational amplifiers.

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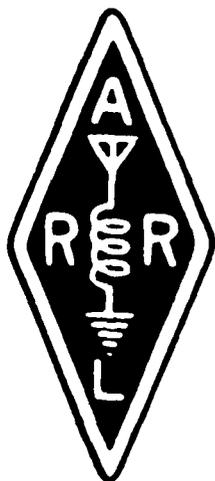
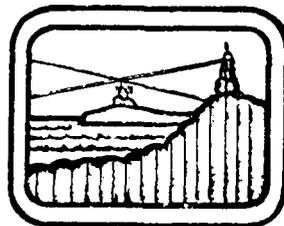
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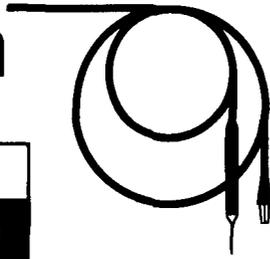
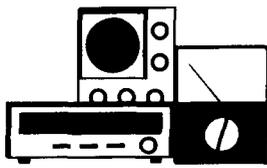
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repair bench



Bob Stein, W6NBI

semi-precision voltage calibrator for digital voltmeters

If you are relying on your digital voltmeter for accurate voltage measurements, you may have a problem which you have never considered — how well is your voltmeter calibrated? The specifications for the instrument tell you its guaranteed accuracy, and should also state the period of time that this accuracy will be maintained after calibration. Typically, the period is between three months and one year, depending on the manufacturer's tests or specmanship. After that, you must accept the DVM's accuracy on faith, or have it recalibrated. If you are fortunate enough to have access to a calibration-standards laboratory, recalibration is no problem. Otherwise, you must return the voltmeter to the manufacturer's service center at a cost of at least \$20 plus shipping, plus the inconvenience of being without it for several weeks.

An alternative is to build the voltage calibrator described in this article. Although it will not permit the same degree of calibration accuracy which can be achieved with precision voltage standards, it will indicate whether or not your DVM requires recalibration and will allow you to perform the calibration if you are willing to accept about 0.3 per cent accuracy on dc and 1 per cent on ac.

Because of its simplicity, the calibrator has several limitations, all of which are based on the assumption that most digital voltmeters owned by hams are inexpensive, 3-1/2-digit types. First of all, both the ac and dc calibration adjustments on the DVM must be made by a single control for each, on the lowest ac and dc voltage ranges (accuracy of the higher volt-

age ranges is dependent upon the accuracy of the input voltage dividers in instruments of this type). Second, the DVM calibration voltages specified in the instruction book must be within 25 per cent of either 100 millivolts or 1 volt; the voltage calibrator is designed to allow you to choose one of these two calibration voltages. And third, the dc input resistance of the DVM must be at least 10 megohms, although lower input resistances can be accommodated at the expense of accuracy.

In addition to the voltage calibrator, an audio-frequency generator having a low output impedance is required. This generator must supply a voltage equal to the specified ac calibration voltage, at the frequency specified by the manufacturer of the DVM. Other than that, the voltage calibrator is self-contained and self-calibrating.

circuit description

The voltage calibrator circuit is shown in **fig. 1**. Two LM308 op amps are used in a precision full-wave rectifier circuit to obtain an ac calibrating voltage by substitution. A reference voltage source, to be described later, provides the dc calibrating voltage and a reference for the rectifier circuit. Either a 100.0-millivolt or a 1.000-volt reference voltage source is used, depending on the requirements of the DVM undergoing calibration. It can be seen that when switch S1 is in the DC CAL position, the reference voltage source is connected directly to the DVM, permitting calibration of the dc voltage range.

Since the precision full-wave rectifier is dc coupled, it will respond to dc as well as ac; its output can therefore be calibrated from the reference voltage source. The ac signal from the audio generator is adjusted to provide an equivalent dc output from the rectifier, which is measured using the DVM (already calibrated on its dc range). Since the ac input is now known in terms of the dc output from the rectifier, the DVM is connected directly to the audio generator and calibrated for ac voltage.

Let's now examine the precision rectifier, described by Dobkin in **reference 1**, in more detail. LM308 op amps were selected because of their low input-bias currents, ensuring that the bias current supplied from the reference voltage source would not exceed 14 nanoamperes. When an ac voltage is applied to the inverting inputs of both op amps, U1 functions as a half-wave rectifier and produces the output shown in **fig. 2B**. The positive half-cycle of the input is inverted in the op amp and applied to the inverting input of U2 through CR1 and R6. The inverted negative half-cycle of the input is clamped at approximately 0.7 volt by CR2, but is isolated from U2 by CR1. Therefore the output of U1 which

reaches U2 is a true half-wave rectified version of the input, having the same peak amplitude, e_{pk} .

To simplify the explanation, it is best to consider the two inputs to U2 separately. The output of U1 is applied to U2 through R6. Therefore the gain of U2 for this input is established by the ratio $(R7 + R8)/R6$. If this gain is set at 2.222 by means of GAIN control R8, the output resulting from this input would be as shown in fig. 2C. Simultaneously, the original ac input is applied to U2 through R4, with the gain of U2 for this input being determined by the ratio $(R7 + R8)/R4$. Since R4 is twice the value of R6, within the tolerances of 1-per cent resistors, the gain becomes 1.111 for the same setting of R8; the resultant output is shown in fig. 2D. Waveforms C and D are summed (more rigorously, the input currents are summed), producing the actual output shown in fig. 2E. Note that this is a full-wave rectified version of the ac input, but amplified by a factor of 1.111. The dc level at the output of the rectifier circuit is equal to the average value of the input signal times the gain of the circuit. Since the average value of a sine wave or a full-wave rectified sine wave is 0.6366 times the peak value, the dc output becomes $0.6366 \times 1.111e_{pk}$, or $0.707e_{pk}$, which is the rms value of the ac input.

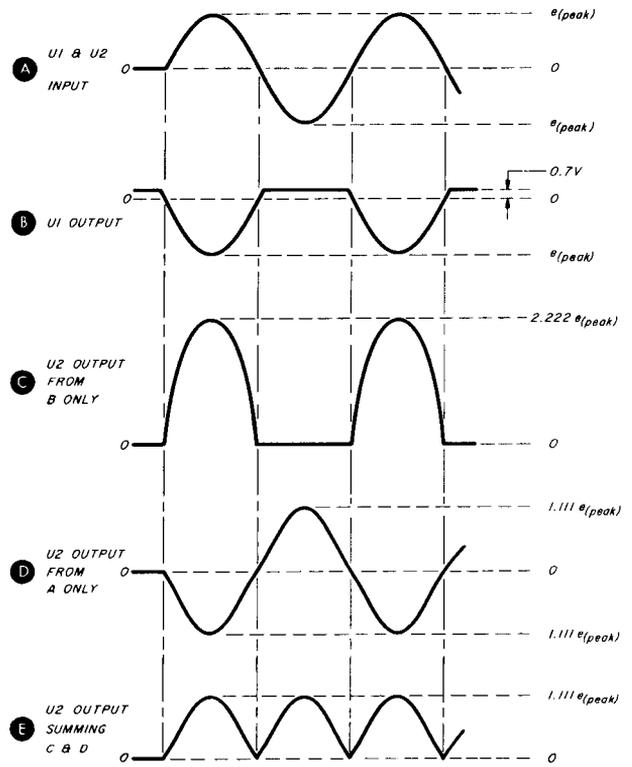


fig. 2. Waveforms of the precision full-wave rectifier circuit.

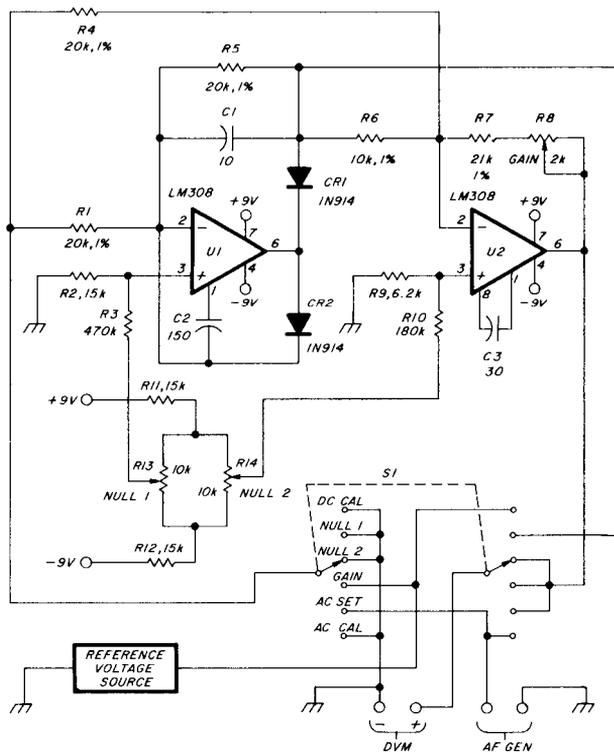


fig. 1. Schematic diagram of the semi-precision voltage calibrator. Switch S1 must be a nonshorting type rotary switch. The reference voltage source is shown in fig. 3.

Note that there are no coupling capacitors within the rectifier circuit, allowing it to operate as a dc amplifier having a gain of 1.111. Therefore a known dc voltage, E , applied to the input will produce an output equal to $1.111E$. It is this characteristic which permits ac calibration using a dc source.

The input offset voltages of U1 and U2 are nulled by means of potentiometers R13 and R14, respectively, in order to establish zero output voltage for zero input. The entire circuit may be powered by two inexpensive 9-volt transistor batteries, or a dual-voltage power supply can be used; total current drain is less than 2 milliamperes.

The sequence of operation is as follows: The DVM is calibrated on its dc range, as described previously, when switch S1 is set to DC CAL. When S1 is set to NULL 1, the output of U1 is nulled by means of R13, and when S1 is set to NULL 2, the output of U2 is nulled with R14. In the GAIN position of S1, the dc reference voltage source is applied to the rectifier input and GAIN control R8 is adjusted for a dc output, indicated on the DVM, of 1.111 times the reference voltage. Next, S1 is switched to its AC SET position, which connects the external af generator to the input of the rectifier. The amplitude of the audio-generator signal is then adjusted to provide a dc reading on the DVM which is equal to the dc reference voltage;

therefore the rms value of this ac signal is now equal to the dc reference. Finally, S1 is set to AC CAL, which connects the audio generator directly to the DVM. Changing the DVM function switch to AC VOLTS then permits calibration of its ac range.

reference voltage source

As you may have deduced from the discussion so far, the accuracy of the voltage calibrator is dependent upon the accuracy of the reference voltage source. Fortunately, the source utilizes an inexpensive and commonly available 1.35-volt mercury cell. Suitable types, the most expensive of which costs about \$2, are listed below. All are 1.35-volt reference cells. Mercury cells are also made with a 1.4-volt emf; they are not usable in this application.

Mallory	Eveready	Burgess
RM12R	E12N	HG12R
RM401R	E401N	HG401R
RM502R	E502	HG502R
RM601R	E601	HG601R

Although the battery voltage is specified to only two decimal places, it's been determined by a large equipment manufacturer that the terminal voltage of a new battery under about 0.1-milliampere load is within a millivolt or two of 1.354 millivolts. This value has been used as a basis for the reference voltage.

Fig. 3A shows the circuit of the 100.0-millivolt source, and fig. 3B shows the 1.000-volt source. To establish a reference voltage with an accuracy of ± 0.2 per cent, it would be necessary to use selected or special values of 0.1 per cent resistors. To eliminate this obviously impractical requirement (for the average user), standard values of 1 per cent resistors have been specified. However, these must be bridged or measured on a calibrated digital ohmmeter which is accurate to within ± 0.1 per cent. Since it is the ratio of the two sections of the voltage divider across the battery, (rather than the absolute resistances) which determines the reference voltage, it is possible to calculate a value for resistor R when the values of R101 and R102 are accurately known. Selecting a standard 1 per cent value which is closest to the calculated value of R will always result in a divider within the required accuracy.*

For example, assume that you have borrowed a Wheatstone bridge and have determined that the values of R101 and R102 for a 100.0-millivolt source are 9984.7 and 803.5 ohms, respectively. From fig. 3A,

$$R_p = \frac{R101}{12.540} = \frac{9984.7}{12.540} = 796.2 \text{ ohms}$$

*Refer to the end of this article for a source of these resistors.

Also from fig. 3A,

$$\frac{1}{R} = \frac{1}{R_p} - \frac{1}{R102} = \frac{1}{796.2} - \frac{1}{803.5} = 0.0001141$$

$$R = 87\,637 \text{ ohms}$$

The nearest standard 1 per cent resistance value is 86.6k ohms, which is then connected in parallel with R102. If you recalculate R_p using the limiting values of an 86.6k, 1 per cent resistor (i.e., 85 734 and 87 466 ohms), and then calculate the ratios of $R101/R_p$, you will find that they are well within 0.1 per cent of 12.540. This holds true for any value of R required for the ranges of R101 and R102.

To simplify the selection of a value for R, R_p has been plotted against the resistance of R101 in fig. 4 for both the 100.0-millivolt and 1.000-volt reference voltage sources. If the measured value of R102 is within 0.1 per cent of R_p , R will not be required. If the value of R102 is higher than R_p , R can be selected as described above. If R102 has a measured resistance less than R_p , it cannot be used, necessitating a change in either R101 or R102.

When the voltage calibrator is in use, either the DVM or the precision rectifier input is connected in parallel with the bottom section of the voltage divider. The effect of this on the 100.0-millivolt source is negligible because of the low value of R_p . However, the effect on the 1.000-volt source cannot be disregarded because of the relatively high value of R101 in that circuit. Thus the equation for the selection of R_p in fig. 3B, as well as the graphical representation in fig. 4, has taken into account the shunting effect of a typical 10-megohm DVM input resistance.

The equivalent input resistance of the precision rectifier circuit (including the effect of the op amps'

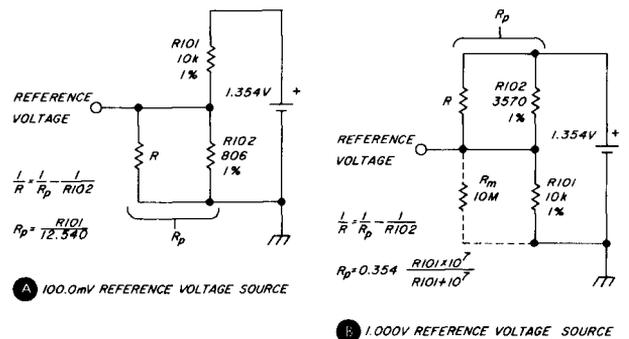


fig. 3. Reference voltage sources for use with the voltage calibrator. In each circuit, R is selected so that it, in parallel with R102, will satisfy the equations shown next to each circuit, within ± 0.2 per cent. R_m in circuit B is the input resistance of the DVM. Refer to the text for suitable battery types.

input-bias currents) will be between 5 and 65 megohms, depending on the individual devices which are used. When connected to the reference voltage source, the input voltage to the rectifier circuit will differ from that applied to the DVM by less than 0.1 per cent, which is insignificant compared to the 1 per cent accuracy of the precision rectifier.

Little need be said about building the voltage calibrator. The parts layout and arrangement are not at all critical, allowing any type of construction to be used. Point-to-point wiring on perf or copper-clad board will suffice, as will a printed-circuit board.

It is recommended that potentiometers R13 and R14 be multiturn types, because the nulling adjustments are critical. Since the calibrator will be used infrequently, these controls as well as GAIN potentiometer R8 can be multiturn, screwdriver-adjust trimmer pots.

One precaution — do not solder the mercury cell into the circuit. Rather, buy or build a suitable holder. A fresh battery should always be used for calibration; the battery must be easily replaceable.

DVM calibration

Connect the DVM to be calibrated to the DVM terminals of the voltage calibrator, and set it to the appropriate dc range. Connect an audio-frequency generator to the AF GEN terminals of the calibrator, with the generator output turned down to zero. Apply +9 and -9 volts dc to the calibrator, and allow the calibrator and DVM to warm up for at least 15 minutes.

The complete calibration procedure is as follows. The numerical values given in each step are based on using the 100.0-millivolt reference; equivalent values for the 1.000-volt reference are in parentheses.

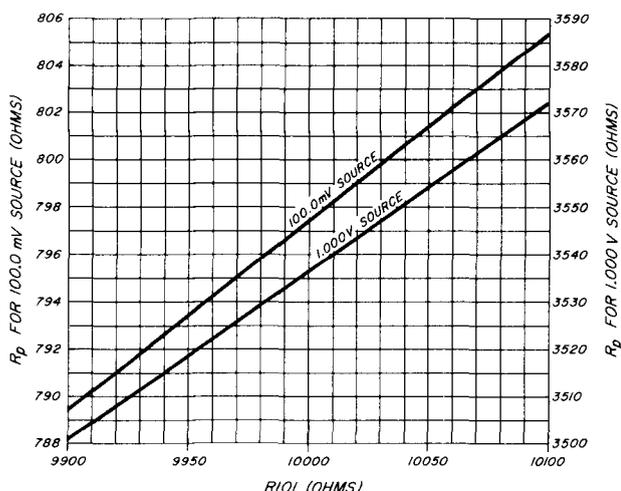


fig. 4. Values of R_p plotted against the measured resistance of R101, from the formulas given in fig. 3. R102 must be equal to or greater than R_p , as explained in the text.

1. Set S1 to DC CAL and adjust the DVM dc calibration control for a reading of 100.0 millivolts (1.000 volt).

2. Set S1 to NULL 1 and adjust NULL 1 potentiometer R13 for a reading on the DVM as close to zero as possible.

3. Set S1 to NULL 2 and adjust NULL 2 potentiometer R14 for a reading on the DVM as close as possible to zero.

4. Repeat steps 2 and 3; the nulls should be within 0.1 millivolt (1 millivolt) of zero.

5. Set S1 to GAIN and adjust GAIN control R8 for a reading of 111.1 millivolts (1.111 volts) on the DVM.

6. Set S1 to AC SET and adjust the af generator output for a reading of exactly 100.0 millivolts (1.000 volt) on the DVM. Note that the DVM is still set to read dc voltage.

7. Set the DVM to read ac volts and set S1 to AC CAL. Adjust the DVM ac calibration control for a reading of 100.0 millivolts (1.000 volt).

The DVM is now calibrated within normally acceptable standards for most amateur work. As the title of this article indicates, the calibrator is only a semi-precision substitute for a voltage calibration standard. But for most ham and experimental applications, its accuracy should suffice.

source of precision resistors

Precision resistors for the voltage calibrator can be supplied by Melvin Sales Company, Post Office Box 5283, San Mateo, California 94402. The following sets will be made available:

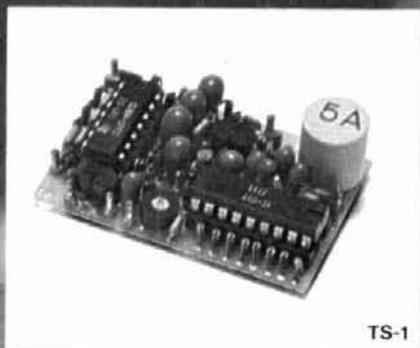
description	postpaid cost, USA & Canada	postpaid cost via air, other countries
Matched set ($\pm 0.2\%$) of voltage-divider resistors for 100.0-mV source	\$2.50	\$3.00
Matched set ($\pm 0.2\%$) of voltage-divider resistors for 1.000-V source	2.50	3.00
Set of five 1% resistors for precision rectifier	1.50	2.00
*Set of five resistors (nominally 100, 1k, 10k, 100k, and 1M ohms) individually calibrated to $\pm 0.1\%$ for use as resistance standards	4.00	4.50

*Not required for voltage calibration, but are suitable for checking or adjusting resistance ranges of a digital volt-ohmmeter.

reference

1. Robert C. Dobkin, "Precision AC/DC Converters," Linear Brief LB-8, National Semiconductor Corporation, Santa Clara, California 95051, August, 1969.

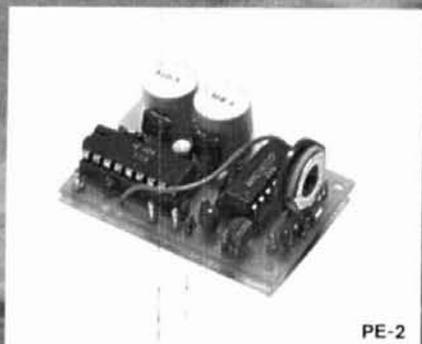
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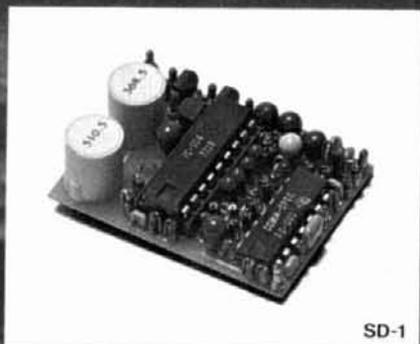
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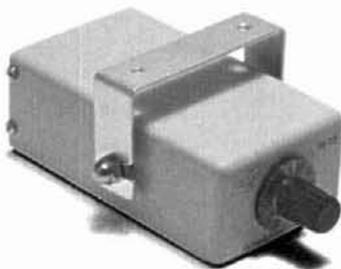
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J antennas

The J antenna has been around a long time. It is primarily a ham antenna, never having gained acceptance in the commercial world, although it has been sporadically manufactured by several companies. The J antenna is a half-wavelength radiator mounted vertically and end-fed with a quarter-wavelength resonant transmission line (basically an end-fed Zep). Its angle of radiation and gain are the same as a half-wavelength antenna; electrically it is similar to the coaxial or sleeve antenna. The horizontal radiation

pattern of the J antenna is almost a perfect circle, even when mounted on a metal roof of a car, because it doesn't require a ground plane and can be mounted well above the vehicle's roof.

The J antenna is fed at the appropriate point on the quarter-wavelength resonant line portion of the antenna (see **fig. 1**). Since the impedance on a resonant quarter-wavelength line goes from zero at the shorted end to infinite at the open end, it's possible to provide a good match to almost any transmission line. For best isolation from the supporting mast, however, it's necessary to feed the antenna with a balanced feedline; if you use coax, this requires a balun. If a balun is too much bother, the antenna can be fed directly with coax at the proper impedance matching point. In this case there will be some interaction between the antenna and the mast.

multibanding

After noting that the J antenna doesn't care what's below the shorting block at the bottom of the quarter-wave section, and it doesn't care about the diameter of the radiator (within reason), I concluded that a 450-MHz J antenna could be located on top of a 144 MHz J as shown in **fig. 2**. Chris Bushman, WB6EEQ, took me seriously and built just such a system. That was two years ago and it's still working very well.

Expanding this concept to include other bands would give you a single antenna structure, to install in that critical spot on the top of your tower, that in-

By Bob Thornburg, WB6JPI, 13135 Ventura Boulevard, Studio City, California 91604

cludes 10 meters, 6 meters, 2 meters, 220 MHz, and 450 MHz. Dimensions are given in **table 1** for all these although only a 144-220-450 mobile version has actually been built and tested by WA6VSK.

design

The design of the J antenna is very simple. The length of the quarter-wavelength matching section is given by

$$L = \frac{1134}{f_{\text{MHz}}} \text{ (cm)} = \frac{2880}{f_{\text{MHz}}} \text{ (inches)}$$

The half-wavelength radiator is approximately twice this length or

$$L = \frac{2181}{f_{\text{MHz}}} \text{ (cm)} = \frac{5540}{f_{\text{MHz}}} \text{ (inches)}$$

The spacing between the mast portion of the quarter-wave section and the quarter-wavelength rod is not critical as long as the two are parallel.

construction

A mobile 144-220-450 MHz J antenna can be easily built from a CB whip. A base station vhf/uhf J that includes 10 and 6 meters should be built with aluminum tubing. Since the antenna should not be insulated, a push-up mast would work well for the 10 meter and up antenna.

The mobile J antenna is made from an ordinary CB whip. For strength cut 60 cm (2 feet) from the top of the antenna and use this piece for the 144-MHz stub. The shorting blocks are aluminum stock drilled as shown in **fig. 3**. The spacers to maintain the stub spacing for 144 and 220 MHz are made from plexiglass, drilled the same as the corresponding shorting block. No plexiglass spacer is needed for the 450-MHz section. Cut the quarter-wavelength stubs

the quarter-wave 200-ohm point is an arbitrary length of TV twinlead. Other forms of feedline can also be used.

The 50-ohm coax may be directly fed to the stub at the appropriate 50-ohm point with the shield tied to

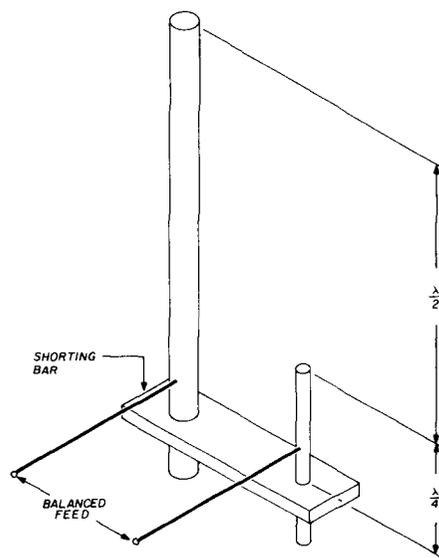


fig. 1. Basic J antenna consists of a half-wavelength vertical radiator and quarter-wavelength matching section. The balanced feedpoint (use balun with coax feedline) is moved along the quarter-wave section until a match is obtained. Unbalanced feed may also be used, as described in the text. The base of the antenna (shorting bar) may be grounded.

the opposite side. The stub could also be insulated from the shorting block and fed like a quarter-wave antenna (which it isn't). In general, the J antenna is relatively tolerant of the actual feedpoint or technique.

table 1. Construction dimensions for the half-wavelength J antenna.

frequency (MHz)	A	B	C	D	E	F
29.0	7.52 m	2.51 m	15 cm	31.8 cm	6.4 cm	3.42 m
52.5	4.16 m	1.38 m	10 cm	19.1 cm	3.8 cm	1.89 m
147.0	1.49 m	49.5 cm	5 cm	6.4 cm	1.9 cm	67.3 cm
223.5	97.2 cm	32.1 cm	3.8 cm	4.1 cm	1.3 cm	44.1 cm
440.0	48.9 cm	16.5 cm	2.5 cm	2.1 cm	1.3 cm	22.4 cm
29.0	296.5"	98.75"	6"	12.5"	2.5"	134.5"
52.5	163.75"	54.50"	4"	7.5"	1.5"	74.25"
147.0	58.50"	19.50"	2"	2.5"	0.75"	26.50"
223.5	38.25"	12.63"	1.5"	1.63"	0.5"	17.38"
440.0	19.25"	6.50"	1.0"	0.83"	0.5"	8.83"

about 8-10 cm (3-4 inches) longer than necessary to allow tuning and as a convenient place to store the balun (under the shorting block).

The balun is simply a half-wavelength section of 50-ohm coaxial cable as shown in **fig. 4**. The feed to

The feedlines for the higher-frequency antennas actually become part of the lower-frequency antenna system. For this reason they must be tied closely to the main antenna pole. This provides capacitive coupling through the coax jacket and ensures that

the outer shield electrically follows the fields on the main antenna; for all practical purposes it becomes part of that structure.

tuning

As noted previously, the specific location of the feedpoints is not too critical. However, the length of the stub is very critical and by adjusting the stub

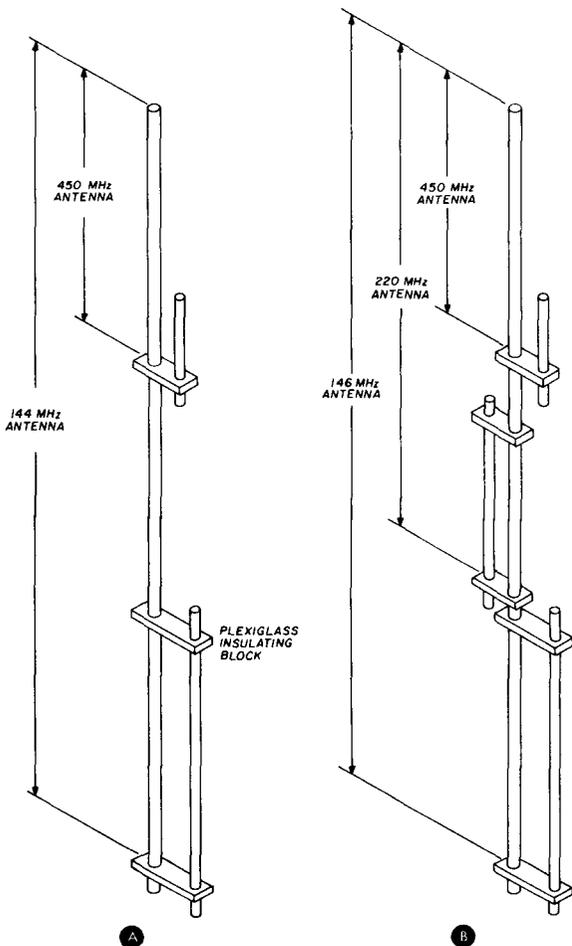


fig. 2. Multiband J antennas for 146 and 450 MHz, A, and for 146, 223, and 450 MHz, B. Antennas built by the author were fed with separate coaxial feedlines and baluns.

length, nearly any antenna length or feedpoint can be matched. The recommended procedure is to fully assemble the antenna, connect an rf source to the highest frequency coax, and adjust the appropriate stub for minimum vswr. If the vswr at the edge of the band is less than 2:1 (and more-or-less symmetrical) move on to the next highest frequency/coax and tune the stub. After adjustment is completed, recheck the lower frequency antenna because there is some interaction.

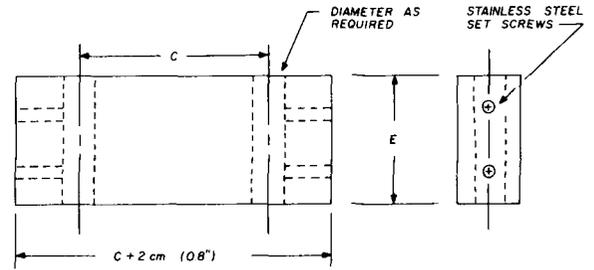


fig. 3. Layout of the aluminum shorting bars and plexiglass spacers. Dimensions C and E are given in table 1.

If the bandwidth is too narrow for your application, load the tuned section more by moving the feedpoint toward the antenna portion. The 2:1 swr bandwidth should be at least 10 MHz at 450 MHz, 4 MHz at 220 MHz, and 3 MHz at 144 MHz.

operation

The original design of the multiband J antenna was intended to allow operation on one band at a time, with the unused coax connectors left open. There is a reasonable amount of coupling between the sections so that if you are transmitting on 144 MHz, for example, some of this energy will be received by the 450-MHz antenna. Measurements have shown this coupled power to be 15-20 dB down, but that's still sufficient to damage a sensitive receiver frontend.

I don't recommend you try two-band duplex operation with this antenna system, although with the excellent preselectors in the Motorola *Motrac*, it has been done on 450 MHz and 146 MHz without damage.

ham radio

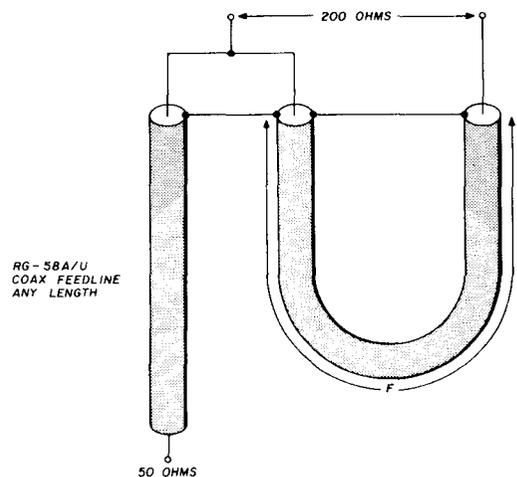


fig. 4. Construction of the half-wavelength coaxial balun. Length F is given in table 1.

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eliminates the need
for empirical design

Almost everyone interested in radio communications has at one time been confronted with the need for a high-frequency oscillator. Unfortunately, little information is available to help those individuals design their own oscillator circuits. This article describes a technique for the design of a Colpitts oscillator which is applicable to both crystal and LC oscillators. The technique to be described is simple, and yet has been proven to be more effective than empirical solutions. Basic oscillator design at the lower frequencies is covered in many textbooks and, therefore, will not be described in detail.

Although this article will deal exclusively with the Siliconix U310, the design equations are also applicable to bipolar transistors. For reasons to be men-

tioned later, bipolar transistors are not as desirable in some applications. Subjects such as oscillator noise and stability are covered elsewhere and will not be described in detail here. More specifically, what will be dealt with is what it takes to get the oscillator to *start* oscillating. The technique offered is simple enough so that you need not know how oscillators oscillate. Since scientific calculators are now selling for less than fifty dollars, it is justifiable to do away with some of the old "rule-of-thumb" solutions.

basic assumptions

The Colpitts oscillator is perhaps most commonly seen in the configuration shown in **fig. 1**. To simplify the analysis, those components necessary for dc operation have been omitted (**fig. 2**). To further simplify the ac analysis, the oscillator is redrawn in the configuration shown in **fig. 3**. This is the general form of a common-gate amplifier, with a feedback capacitor, C1, between the input and the output. The amplifier is considered with no signal applied from an external source; the input is shunted only by C2 and the source-bias resistor. Note that if the capacitors were replaced with inductors and vice versa, the circuit would be the common Hartley oscillator. An analysis is somewhat easier with the circuit in **fig. 3** since it now appears as a resonant tank circuit with an amplifier connected to it. However, one more circuit element needs to be added: the load resistance, R_L . This is the element which will accept power from

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the oscillator. For this circuit to oscillate, all that's necessary is enough energy be tapped from the tank circuit, amplified, and routed back to the tank circuit to compensate for the energy absorbed by R_L .

The first step is to determine what output power is required. This, of course, depends upon the application, but in most cases it will be relatively low, particularly when frequency stability is of prime importance. Crystal oscillators generally have relatively low output levels, mainly to prevent the crystal from fracturing. In one case, a crystal ordered from a prominent manufacturer had a rated power dissipation of one milliwatt. To obtain more power from an oscillator with a fixed-supply voltage, a lower load impedance is required. If a lower load impedance is applied, however, either the gain of the transistor has to be increased, or feedback has to be increased to maintain oscillation. Since the crystal is in series with the feedback signal, care must be taken when considering how much output power you can expect from the oscillator.

Another consideration regarding output power is that inductors and capacitors do consume some power. They always have some associated series resistance which can be minimized by using higher-quality components. When rf current passes through these components, heat is generated from the power dissipated across the resistance. This heat causes a change in the values of the inductor or capacitor and, hence, a change in frequency. The effects of these changes and the changes associated with various components are covered in reference 1.

The transistor also has power-limiting characteristics. From turn-on of the oscillator, until a steady-state condition has been reached, the transistor parameters will change. The amount of change must

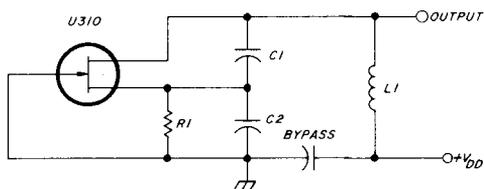


fig. 1. Schematic diagram of a basic common-gate Colpitts oscillator.

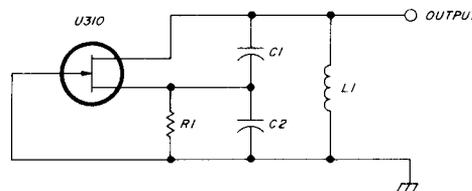


fig. 2. Ac model of the common-gate Colpitts oscillator shown in fig. 1. For analysis, the dc components have been eliminated.

generally be determined empirically, but can be minimized by operating the transistor at relatively low power, limiting output from the oscillator.

In this article, the Siliconix U310 jfet is characterized at 9 volts drain-to-gate voltage and 2 mA of drain current in the common-gate configuration. This is thought to be a fair compromise between output power and parameter changes. Consider a class-A oscillator in which the drain current does not appreciably change from the oscillating to not-oscillating condition. If optimized, it will be less than 50 per cent efficient, with a maximum output power of 9 milliwatts ($9\text{ V} \cdot 2\text{ mA}/2$). Oscillator design almost always requires some compromises, so there is nothing binding regarding the 9 volt, 2 mA operating point; it can be changed to meet the needs of your application.

oscillator design

The first step to consider in oscillator design is the required output power; the second is to determine the load resistance necessary to obtain the required output. The load the oscillator requires is almost always different than the amplifier, buffer, or mixer it must drive. Initially, only the oscillator load will be considered.

Only class-A oscillators will be discussed in this article because the transistor parameter changes are more easily defined. If the supply voltage and drain current are known, the solution for the load resistance, R_L , is

$$R_L = \frac{(V_{DS} - V_{DS(sat)})^2}{2 P_{out}}$$

$V_{DS(sat)}$ can be obtained from the data sheet; in this

case it is 2 volts. Earlier it was determined that 9 milliwatts would be a desirable output. Given the operating point of 9 volts:

$$R_L = \frac{(9 - 2)^2}{2(0.009)} = 2722 \text{ ohms}$$

Since $V_{DS(sat)}$ is only an approximation, rounding R_L to 2700 ohms for use in the following calculations is justified.

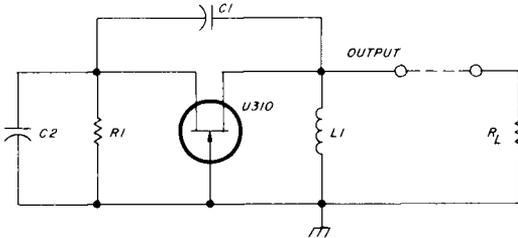


fig. 3. To further simplify analysis of the oscillator, this diagram shows the circuit redrawn into the general form of a common-gate amplifier.

The third item to consider is transistor selection. The Siliconix U310 has been demonstrated to operate as an oscillator up to 900 MHz.² It has an advantage over more commonly available vhf and uhf jfets because of its high zero-bias drain current and g_m (forward transconductance). This means that the U310 has potentially higher output power and more stable characteristics when operated at a lower-drain current. Another distinct advantage of the U310 is that the gate lead, and in this case, the intrinsic bulk of the transistor, is connected to the metal case of the transistor package. Therefore, the U310 can be used with a heatsink, minimizing the change in transistor parameters. Although not recommended by the factory, an alternative heatsink method is to solder the U310's case to the chassis. This is convenient only when used in the common-gate configuration (be careful when soldering, as the U310 can be destroyed by overheating).

Perhaps the most critical parameter in *amplifier* design is the gain stability factor of the transistor itself.³ The Siliconix U310, when operated in the common-gate configuration at 20 volts and 10 mA, is unconditionally stable at almost all frequencies. In fact, when loaded with practical, somewhat lossy external components, the U310 could be considered unconditionally stable at all frequencies. Unconditional stability, by definition, means the transistor will not oscillate when presented with any positive real source or load impedance; stability, therefore, is a measure of the transistor's ability to oscillate.

It is desirable that the stability factor of the transistor is such that it will not oscillate. This may seem

contradictory to the design goal, but is used to emphasize that the source of feedback necessary to sustain oscillation should be the option of the *designer* and not the transistor. In the common-gate configuration, the U310 has been optimized to the point where the intrinsic feedback elements of the transistor are so small that the values of feedback necessary to make the U310 oscillate *are* the choice of the designer.

Most rf bipolar transistors, in the equivalent common-base configuration, are only conditionally stable and tend to be so at many frequencies. Special precautions must be taken to insure oscillation at only the desired frequencies. This requires more components, which increases cost and decreases reliability. Also, very few bipolars, *designed* for the common-base configuration, are available.

The next item to consider is the operating or loaded Q (Q_L) of the resonant tank circuit. The value chosen for Q_L is very much dependent upon the application. In addition, there are upper and lower restrictions on Q_L . In an attempt to make Q_L low, the transistor's reactive components would assume a higher percentage of the total circuit values. This is undesirable in many applications since stability would be sacrificed as the transistor parameters change with temperature. Additionally, the harmonic content of the output increases as Q_L decreases. There are many applications where Q_L should be very high. One example would be when a low-noise oscillator is required to drive a sensitive mixer or product detector.

The upper limit for Q_L is established by the unloaded Q (Q_u) of the inductor and how critical the tuning adjustment can be; this is particularly true for crystal oscillators.

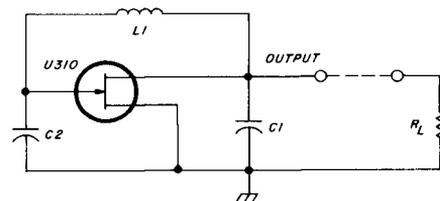


fig. 4. A common-source oscillator shown as an ac model.

At very high frequencies, the Q_u of the capacitor must also be considered, especially when using capacitors generally designed for use at lower frequencies. While Q_L is used as a starting point for calculations, the final value of Q_L will be different than the initial assigned value. In general, the small change will not affect circuit performance.

The Q_u of a reactive element is determined by the intrinsic resistive element. In circuits where higher-quality capacitors are used, this resistance generally can be disregarded. Inductors, however, with only a few exceptions, must always be considered as having some associated resistance. The unloaded Q is defined as:

$$Q_u = \frac{X_L}{R_S}$$

where R_S is the series resistance of the inductor, and X_L is the inductive reactance at the operating frequency f_o . Q_u can be measured, but in those instances where it can't, assuming a Q_u between 150 to 250 is safe, if standard inductor winding techniques are used.

In those cases where a Q_L of 10 or greater is chosen, it is beneficial to include Q_u in the calculations necessary to determine total R_L , since the circuit performance will be altered. When time permits, it is advisable to include Q_u for all designs. The effects of Q_u can best be emphasized by the following:

$$\text{Insertion loss} = 20 \log_{10} \left[1 - \frac{Q_L}{Q_u} \text{ (dB)} \right]$$

For a Q_L of 10 and a Q_u of 100, the insertion loss would equal 0.92 dB. This means that 0.92 dB of the total power is dissipated across the inductor rather than delivered to R_L . A side effect is that as Q_u gets smaller, the change in frequency from initial turn-on increases because the power dissipated across the inductor causes it to change reactance. This is especially true in those cases where a ferrite core is used for tuning.¹ Many of the commercially-available inductors, particularly the smaller molded variety, are not designed for use in frequency-determining applications where stability is of concern. The Q_u of many of the smaller molded inductors is as low as 50, which means with a $Q_L = 10$, the insertion loss of the inductor is 1.94 dB. The point to keep in mind is that wherever frequency stability is the prime consideration, Q_u is *very important*, even though its effects may have to be determined empirically.

y parameters

This design technique will use what is commonly referred to as admittance, or y parameters. The oscillator designs presented here require the use of algebra and trigonometry with the y parameters. Review material can be found in any of the standard references.^{4,5}

One item which should be understood concerning y parameters is their relationship to ohms and reac-

tance. Resistance is a measure of the opposition to current flow. Its counterpart in terms of admittance is conductance, which is a measure of the acceptability to current flow. They are simply related; one thousand ohms is equal to 1/1000 mho (the mho is designated as the *siemens*).

Reactance is a measure of the opposition to changing current flow and its counterpart in terms of admittance is susceptance; these are also similarly related: 1000 ohms reactance is equal to 1/1000 siemens of susceptance. Just as inductors and capacitors have different signs attached to them in terms of reactance, they also have different signs in terms of susceptance.

Inductors in terms of susceptance have negative signs placed in front of them and capacitors have plus signs in front of them. Therefore, a capacitive reactance of 1000 ohms is equal to +1/1000 siemens, and an inductive reactance of 1000 ohms is equal to -1/1000 siemens. Another simple relationship is that 1000 ohms equals 1 millisiemens or mS, ($1000 \text{ ohms} = 1/1000 \text{ siemens} = 0.001 \text{ siemens} = 1 \text{ millisiemens}$). This article will use the term millisiemens or mS often so it is important to be familiar with the relationships. The relationships previously described regarding reactance and susceptance are correct and usable as relates to this article.

Referring to **figs. 3 and 4**, the oscillator designs will determine the component values necessary to *start* the initial oscillations. For oscillations to start, the ac resistance measured at the junction of C1, L1, R_L , and the drain of Q1 must be infinite.⁵ But, in terms of conductance, it would be zero. Therefore, the conditions necessary to start oscillations are:

$$y_t = y_{22} - \frac{y_{21}y_{12}}{y_{11}} = \text{zero}$$

where

- y_t = terminal conductance (junction of C1, L1, R_L , and the drain of Q1)
- y_{22} = output admittance of the circuit
- y_{21} = forward transconductance of the circuit
- y_{12} = reverse transconductance of the circuit
- y_{11} = input admittance of the circuit

These parameters also include the y parameter data of the transistors, which in most cases are available from their data sheets. The parameters might be listed as shown in **table 1**.

Listed below are the equations which relate common source, gate, and drain parameters to each other. These equations also relate to bipolar transistors, except the source becomes the emitter, the gate becomes the base, and the drain becomes the collector. To convert common-gate parameters to

common-source parameters:

$$y_{11s} = (g_{ig} + g_{fg} + g_{rg} + g_{og}) \pm j(b_{ig} + b_{fg} + b_{rg} + b_{og})$$

$$y_{21s} = -(g_{fg} + g_{og}) \pm j(b_{fg} + b_{og})$$

$$y_{12s} = -(g_{rg} + g_{og}) \pm j(b_{rg} + b_{og})$$

$$y_{22s} = g_{og} \pm b_{og}$$

where the non-subscript g represents conductance and the non-subscript b represents susceptance. To convert from common source to common gate, exchange s and g subscript values. That is

$$y_{21g} = -(g_{fs} + g_{os}) \pm j(b_{fs} + b_{os})$$

From this point on, unless stated otherwise, *all* numerical terms will be in millisiemens (mS); instead of using the term 1000 ohms or 1/1000 siemens, 1 mS will be used. Further, in many cases the mS will be *implied* and, therefore, just the number with its sign will be used. If no sign is given, it is assumed to be positive. As an example: $y_{11} = +7.66 \times 10^{-3}$ siemens $+ j 1.59 \times 10^{-3}$ siemens will be shown as $y_{11} = 7.66 + j 1.59$, where the first term is the conductance and the second term the susceptance.

common source design

This design example will use the common-source oscillator shown in **fig. 4**. Earlier calculations for the common-gate oscillator determined the R_L to be 2700 ohms, which will also be used in the common-source example in addition to an f_o of 100 MHz. A Q_L of 10 has been selected since it yields practical component values. Referring to **fig. 5**,

$$Q_L = \frac{R_L}{X_{LI}} \text{ or } X_{LI} = \frac{2700}{10}$$

therefore, X_{LI} equals 270 ohms or -3.7 mS. At 100 MHz, this is an inductance of 430 nH, or approximately 10 turns of no. 22 AWG (0.6 mm) enameled wire wound on a 6.5 mm (1/4-inch) drill bit. An actual inductor was wound and found to be 422 nH for an X_{LI} of 265 ohms. The number used for the calculations is therefore -3.77 mS. The Q_u of the inductor was measured to be 188, so the resistance in shunt with the inductor is equal to 0.02 mS, [$(X_{LI})(Q_u) = R_p$, $(265)(188) = 50k$ ohms or 0.02 mS]. The inductive susceptance (b_L) and conductance (g_L) can be added to the transistor's y parameters in the following manner:

$$y_{11} = [(g_{11s} + g_L) \pm j](b_{11s} + b_L)$$

$$y_{21} = [(g_{21s} - g_L) \pm j](b_{21s} - b_L)$$

$$y_{12} = [(g_{12s} - g_L) \pm j](b_{12s} - b_L)$$

$$y_{22} = [(g_{22s} + g_L) \pm j](b_{22s} + b_L)$$

The 100 MHz y parameters, from **table 1**, for the Siliconix U310 are:

$$y_{11s} = -0.919 + j 3.34$$

$$y_{21s} = 8.5 - j 1.8$$

$$y_{12s} = -0.0455 - j 1.13$$

$$y_{22s} = 0.129 + j 1.18$$

Adding the inductive susceptance ($b_L = -3.77$) and shunt conductance ($g_L = 0.02$) yields:

$$y_{11} = [(-0.919 + 0.02) \pm j](3.34 + (-3.77))$$

$$y_{21} = [(8.5) - (0.02)] \pm j[(-1.8) - (-3.77)]$$

$$y_{12} = [(-0.0455) - (0.02)] \pm j[(-1.13) - (-3.77)]$$

$$y_{22} = [(0.129 + 0.02)] \pm j[(1.18) + (-3.77)]$$

or

$$y_{11} = -0.0899 - j 0.435$$

$$y_{21} = 8.48 + j 1.97$$

$$y_{12} = -0.0655 + j 2.64$$

$$y_{22} = 0.149 - j 2.59$$

Any slight discrepancies noted can be attributed to round-off errors. All of these calculations were done with an HP67 programmable calculator with automatic rounding.

table 1. y-parameter listing for the Siliconix U310 operated at 9 volts and 2 mA. The unit, in all cases, is mS.

	common gate		common source	
	$y_{11} = Y_{ig}$ input admittance		$y_{11} = Y_{is}$ input admittance	
	$y_{21} = Y_{fg}$ forward transconductance		$y_{21} = Y_{fs}$ forward transconductance	
	$y_{12} = Y_{rg}$ reverse transconductance		$y_{12} = Y_{rs}$ reverse transconductance	
	$y_{22} = Y_{og}$ output admittance		$y_{22} = Y_{os}$ output admittance	

common gate				
freq. (MHz)	y_{11}	y_{21}	y_{12}	y_{22}
100	7.66 + j 1.59	-8.62 + j 0.615	-0.0831 - j 0.0512	0.129 + j 1.18
125	8.75 + j 1.92	-8.56 + j 0.824	-0.0914 - j 0.0714	0.111 + j 1.49
150	8.98 + j 2.34	-8.47 + j 0.816	-0.0797 - j 0.0861	0.129 + j 1.81
175	9.06 + j 2.80	-8.55 + j 0.646	-0.0669 - j 0.0972	0.126 + j 2.13
200	9.07 + j 3.24	-8.84 - j 0.112	-0.0602 - j 0.119	0.124 + j 2.42

common source				
freq. (MHz)	y_{11}	y_{21}	y_{12}	y_{22}
100	-0.919 + j 3.34	8.50 - j 1.8	-0.0455 - j 1.13	0.129 + j 1.18
125	0.209 + j 4.16	8.45 - j 2.31	-0.0197 - j 1.41	0.111 + j 1.49
150	0.563 + j 4.88	8.34 - j 2.62	-0.0490 - j 1.72	0.129 + j 1.81
175	0.576 + j 5.48	8.42 - j 2.77	-0.0589 - j 2.03	0.126 + j 2.13
200	0.294 + j 5.43	8.71 - j 2.31	-0.0633 - j 2.3	0.124 + j 2.42

The next step is to add the load conductance (0.370 mS) to g_{22} , the *real* part of y_{22} .

$$y_{22} = (0.149 + 0.37) - j 2.59 \\ = 0.519 - j 2.59$$

Change the sign of the susceptance and record it as $b_{22} = 2.59$ mS. This is the starting value of susceptance which will be used to tune the circuit to resonance, whereupon b_{22} equals zero. The new set of y parameters from the previous operations are:

$$y_{11} = -0.0899 - j 0.435 \\ y_{21} = 8.48 + j 1.97 \\ y_{12} = -0.0655 + j 2.64 \\ y_{22} = 0.519 \pm j 0$$

The next step is to find the value which when added to the input of the transistor will make the circuit's output conductance zero. This can be found by solving:

$$y_{11t} = \frac{y_{21}y_{12}}{g_{22}} - y_{11} \\ = \frac{(g_{21}g_{12} - b_{21}b_{12}) + j(g_{21}b_{12} + g_{12}b_{21})}{g_{22}} - (g_{11} + j b_{11}) \\ = \frac{-0.00576 + j 0.0222}{0.519} - (-0.899 - j 0.433) \\ = \frac{0.023 \angle 105^\circ}{0.519} - (-0.899 - j 0.433)^* \\ = 0.0442 \angle 105^\circ - (-0.899 - j 0.433) \\ = (-11.1 + j 42.8) - (-0.899 - j 0.433) \\ = -10.2 + j 43.3 \text{ mS}$$

Since the solution produces a real part which is negative, the addition to the input of the transistor cannot be performed with passive components. The real part is not always negative; in many cases it can be a positive resistance, but if added to the circuit it increases the cost and adds complexity. I'll demonstrate that the real component of y_{11t} can be neglected, with little error added to the calculations.

*The numerator in the expression $y_{21}y_{12}/g_{22}$ must be converted to polar form before it can be divided by g_{22} . When a polar-to-rectangular conversion key is not provided on the calculator, the following rules must be applied: If the real part of the numerator is greater than zero, the angle is equal to $\tan^{-1}(Im/Re)$. If the real part is less than zero or negative, the angle is $180^\circ + \tan^{-1}(Im/Re)$, where Im is the imaginary part of $y_{21}y_{12}$ and Re is the real part. To divide, retain the angle of $y_{21}y_{12}$ and divide by 0.519. The polar form of this expression must now be converted back to rectangular form to subtract y_{11} . The imaginary part is $(\sin 105^\circ)(0.0442) = 42.8$ mS and the real part $(\cosine 105^\circ)(0.0442) = -11.1$ mS.

In this design, the real part will be disregarded and only the imaginary part (43.3 mS) will be used.

The b_{11t} value of 43.3 mS is the susceptance of the input shunt capacitance, C2. At 100 MHz, 43.3 mS

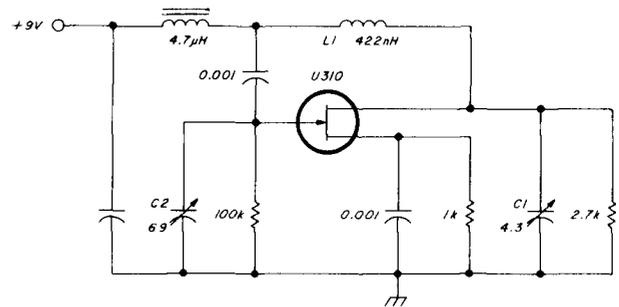


fig. 5. A common-source Colpitts oscillator designed using the procedure described in the text.

susceptance is equal to an X_{C2} of 23.11 ohms, which is 69 pF. Add 43.3 mS (b_{11t}) to b_{11} for a new total b_{11} of 42.8 mS. Subtract the load conductance, 0.370 mS, from g_{22} for a new g_{22} of 0.149 mS. And finally, solve the equation:

$$y_{out} = y_{22} - \frac{y_{21}y_{12}}{y_{11}} \\ = 0.149 - \left[\frac{(g_{21}g_{12} - b_{21}b_{12}) + j(g_{21}b_{12} + g_{12}b_{21})}{-0.899 + 42.8} \right] \\ = 0.149 - \left[\frac{-0.00576 + j 0.0222}{-0.899 + 42.8} \right] \\ = 0.149 - \left[\frac{0.023 \angle 104.5^\circ}{42.81 \angle 91.2^\circ} \right] \\ = 0.149 - (5.357 \times 10^{-4} \quad 13.3) \\ = 0.149 - (0.521 + j 0.123)$$

$$y_{out} = -0.373 - j 0.123 \text{ millisiemens}$$

To tune to resonance, change the sign of the imaginary part and add this value to the first recorded value of b_{22} ($2.59 + 0.123 = 2.71$ mS). This is the total susceptance of C1 at 100 MHz. The reactance of C1 is 369 ohms, or 4.31 pF at 100 MHz. Note that the real part of y_{out} is equal to -0.373 mS or -2681 ohms. The desired value was -2700 ohms, an error of less than 1 per cent. This is the error introduced by not using the real part of y_{11t} .

After all the calculations were performed, the circuit in fig. 5 was constructed and tested. The value of C2 required to start oscillations was found to be 45 pF instead of the calculated value of 69 pF. The feedback network consisting of L1 and the dc blocking

capacitor were removed, and the circuit was measured again. The lead lengths, added value of the dc blocking capacitor, and the fact that the inductor was slightly distorted from its original shape, changed the total feedback circuit to be equivalent to

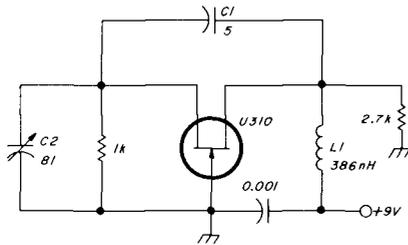


fig. 6. A common-gate oscillator designed with the generalized design technique.

497 nH or -3.2 mS. This changes the calculated value of C2 from 69 pF to 53 pF. Further, if the value of g_m (forward transconductance) were reduced by a factor of 10 per cent, a conceivable situation, the value of C2 would change to 48 pF, not far from the measured 45 pF.

The previous calculations assume that all measurements are absolutely accurate. Since this is not possible in practice, C1 and C2 should be variable to compensate for inaccuracies in measurements, as well as changes in transistor parameters.

The values calculated for the passive components in the circuit are those values required to start oscillation. As the circuit oscillates, the net parameters of the transistor change and consequently, the values of C1 and C2 will change. The equations presented in this article only provide a starting point, but are a preferred alternative to the empirical approach.

common gate design

The common-gate oscillator in fig. 3 is easily designed by starting with the values of C1 and L1 obtained from the common-source oscillator problem and adding them to the common-gate parameters. The common-gate y parameters for the U310 at 9 volts and 2 mA are:

$$\begin{aligned}
 y_{11g} &= 7.66 + j 1.59 \\
 y_{21g} &= -8.62 + j .615 \\
 y_{12g} &= -0.0831 - j 0.0512 \\
 y_{22g} &= 0.129 + j 1.18
 \end{aligned}$$

Add the source resistor (in this case 1000 ohms or 1 mS) to g_{11} ; y_{11} then equals $8.66 + j 1.59$. Using the value of capacitance for C1 obtained in the first design, select the closest standard value, fixed

capacitor available. Since a 5-pF capacitor is generally more available, this value is used for C1 in fig. 6. Since both terminals of C1 are at some rf potential, a tuning tool will generally change the total value of the feedback capacitance by some unknown amount, therefore necessitating the use of a fixed value, instead of a variable capacitor. The 5-pF value is equal to 3.14 mS at 100 MHz, and can be added to the new y parameters in the following manner:

$$\begin{aligned}
 y_{11} &= g_{11} \pm j (b_{11} + b_f) \\
 y_{21} &= g_{21} \pm j (b_{21} - b_f) \\
 y_{12} &= g_{12} \pm j (b_{12} - b_f) \\
 y_{22} &= g_{22} \pm j (b_{22} + b_f)
 \end{aligned}$$

which yields

$$\begin{aligned}
 y_{11} &= 8.66 + j 4.73 \\
 y_{21} &= -8.62 - j 2.53 \\
 y_{12} &= -0.0831 - j 3.19 \\
 y_{22} &= 0.129 + j 4.32
 \end{aligned}$$

To the new value of y_{22} , you should add the value of R_L (0.370 mS) to g_L and also X_{L1} (-3.77 mS) to b_{L1} , yielding $y_{22} = 0.499 + j 0.551$. The value of b_{22} should be recorded for future use. As in the common-source example, set b_{22} equal to zero and calculate y_{11i} . The calculated value of y_{11i} ($-23.4 + j 50.9$) serves as a starting value for C2. The next step consists of adding 50.9 mS to b_{11} , subtracting the g_L of 0.370 mS from g_{22} , and solving for y_{out} , which equals $-0.338 - j 0.205$ millisiemens. The final value for L1 can be determined by solving the following equation:

$$\begin{aligned}
 X_{L1} &= (-1)(b_{out}) + (-1)(b_{22}) + (b_{L1}) \\
 &= (-1)(-0.205) + (-1)(0.551) + (-3.77) \\
 &= -4.12 \text{ mS}
 \end{aligned}$$

which equals 243 ohms or 386 nH.

Note that $g_{out} = -0.338$ mS ($-2.95k$ ohms), which is not the desired $-2.7k$ ohms. The reason for the er-

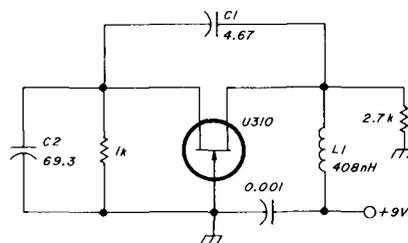


fig. 7. The same oscillator shown in fig. 6, except designed with the more accurate method.

ror is the same as before — the real part of y_{11t} was disregarded. This can easily be compensated for by decreasing the susceptance of C2. As an example, by reducing the susceptance of C2 by 10 per cent, g_{out} becomes -2626 ohms which is sufficient to start oscillation. The circuit for the common-gate configuration, before adjustment of C2 for the necessary g_{out} , is shown in **fig. 6**. In this solution, the inductor should be variable, or the inductance decreased and a variable shunt capacitor added for adjustment.

For those cases when a more accurate result is necessary, the following procedure may be used. Use those component values determined in the design for **fig. 5** as a starting point and proceed with the following steps.

1. Start with the U310 common-gate parameters.
2. Add the source resistor to g_{11} (in this case, 1 mS)
3. Add the susceptance of C2 to b_{11}
 $y_{11} = 8.66 + j 44.9$
4. Add the susceptance of C1 to b_{11} and b_{22} .
5. Subtract the susceptance of C1 from b_{21} and b_{12} , which produces new y parameters of:

$$\begin{aligned} y_{11} &= 8.66 + j 47.6 \\ y_{21} &= -8.62 - j 2.1 \\ y_{12} &= -0.0831 - j 2.76 \\ y_{22} &= 0.129 + j 3.89 \end{aligned}$$

6. Add the susceptance of L1 to b_{22}
7. Add the load conductance to g_{22}
 $y_{22} = 0.499 + j 0.124$
8. Record the value of $(-1)(b_{22})$ or -0.124 mS
9. Set b_{22} equal to zero
10. Solve the following equation for g_f and b_f

$$\begin{aligned} g_f \pm j b_f &= \frac{A + jB}{C + jD} \\ &= \frac{0.0094 - j 0.255}{-0.451 - j 42.7} \\ &= 0.00365 + j 0.220 \end{aligned}$$

where

$$\begin{aligned} A &= (g_{22}g_{11} - g_{12}g_{21} + b_{21}b_{12} - b_{11}b_{22}) \\ B &= (g_{22}b_{11} + b_{22}g_{11} - g_{12}b_{21} - g_{21}b_{12}) \\ C &= (-1)(g_{12} + g_{21} + g_{11} + g_{22}) \\ D &= (-1)(b_{12} + b_{21} + b_{22} + b_{11}) \end{aligned}$$

11. Record the value of b_f
12. Add b_f (0.220 mS) to b_{11} and b_{22}

13. Subtract b_f from b_{21} and b_{12} to obtain

$$\begin{aligned} y_{11} &= 8.66 + j 47.8 \\ y_{21} &= -8.62 - j 2.32 \\ y_{12} &= -0.0831 - j 2.98 \\ y_{22} &= 0.499 + j 0.220 \end{aligned}$$

14. Solve for y_{11t} , which yields $0.118 + j 0.260$
15. Record b_{11t} (0.260 mS)
16. Add b_{11t} to b_{11} , $y_{11} = 8.66 + j 48.1$
17. Subtract the load conductance 0.370 mS from g_{22} , $y_{22} = 0.129 + j 0.220$
18. Solve for y_{out} , which equals $-0.371 + j 0.00109$ mS
19. Record the value $(-1)(b_{out}) = -j 0.00109$ mS
20. Final values for C1, C2, and L1 are determined as follows:

C1 Starting value of C1 from fig. 5	2.71	mS
value of b_f	0.22	mS
	2.93	mS

C1 final value = 2.93 mS or 4.67 pF

C2 Starting value of C2 from fig. 5	43.3	mS
value of b_{11t}	0.26	mS
	43.6	mS

C2 final value = 43.6 mS or 69.3 pF

L1 Starting value of L1 from fig. 5	-3.77	mS
plus $(-1)(b_{22})$ from step 8	-0.124	mS
plus $(-1)(b_{out})$ from step 19	-0.00109	mS
	-3.9	mS

L1 final value = -3.9 mS or 409 nH

The final circuit diagram, with component values, is shown in **fig. 7**. Note that g_{out} from **step 18** is within 0.32 per cent of the desired g_{out} of -0.37 mS or $-2.7k$ ohms. Again, the reason for the slight

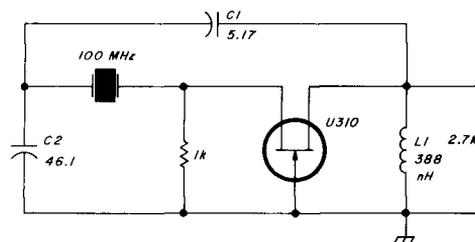


fig. 8. A 100-MHz crystal oscillator which was designed using the same methods as the Colpitts oscillators.

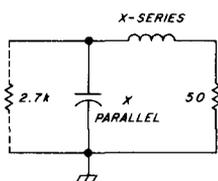
discrepancy is caused by disregarding the real parts of y_{11t} and g_f , and is of no consequence.

crystal oscillators

This oscillator design technique can be modified for use with crystal oscillators, particularly overtone crystals (fig. 8). When operated in the series-resonant mode, the crystal has some series resistance, which must be added to the transistor's common-gate parameters, along with the 1000-ohm source resistor. Given the common-gate y parameters:

1. Add the source resistor to g_{11} .
2. Convert the y parameters to z parameters and add the series resistance of the crystal directly to the real part of z_{11} . A typical value of crystal series resistance is 80 ohms for a seventh overtone crystal; this data is available from the crystal manufacturer.
3. Convert the z parameters back to y parameters.
4. Convert the new set of common-gate parameters to common-source parameters.
5. Design the circuit using the technique described for fig. 5.
6. Using the common-gate parameters from step 3, and the component values determined from step 5, complete the design by using those steps outlined for fig. 7.

fig. 9. A lowpass L network for matching between the high source resistance of the oscillator and a 50-ohm load.



The necessary equations to convert y parameters to z parameters are:

$$z_{11} = \frac{y_{22}}{\Delta y} \quad z_{21} = \frac{-y_{21}}{\Delta y}$$

$$z_{12} = \frac{-y_{12}}{\Delta y} \quad z_{22} = \frac{y_{11}}{\Delta y}$$

where

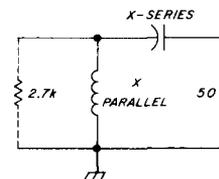
$$\Delta y = y_{11}y_{22} - y_{21}y_{12}$$

To convert z parameters back to y parameters, interchange the y and z values.

It is helpful to know what effects the three circuit components have on the oscillator circuit. In general,

when the input shunt capacitor is increased in value, the frequency of the oscillation is decreased, while the negative output resistance increases. Increasing the feedback capacitance lowers the frequency and also the negative output resistance. Changing the output reactance, theoretically, only changes the frequency and does not effect the output conductance.

fig. 10. A highpass L network, in addition to impedance matching, also provides dc isolation for the load.



The tuning procedure for the oscillator is quite easy (refer to figs. 3 and 4). Apply dc power to the oscillator. If it immediately oscillates, tune the output shunt element to the desired frequency. If the oscillator does not start, or ceases oscillation when tuning, decrease the capacitance of C2, which adjusts the output conductance. When it has been determined that the oscillator is tuned to frequency and oscillating, the input shunt element (C2) can be adjusted for the desired output level and the output shunt element tuned for the correct frequency.

The crystal oscillator is tuned in the same manner. The exception is when the parallel capacitance of the crystal is relatively high. In that case the procedure is to increase the capacitance of C2 until oscillation ceases and then decrease the capacitance in small increments until the circuit oscillates again. This procedure should be followed since in some cases, the parallel capacitance of the crystal provides enough feedback to allow the circuit to oscillate at frequencies very close to f_o , but not as a function of the series-resonant mode of the crystal. For breadboard designs, it is easiest to insert a resistor, equal to the series resistance of the crystal, in place of the crystal; the same tuneup procedure is used except that the resistor is replaced with the crystal during the last stages of tuning.

As an added advantage, when using the U310 in the crystal oscillator circuit, the inductor sometimes shunted across the crystal to prevent spurious oscillations is not necessary. This inductor is almost always necessary when using a bipolar transistor as a Colpitts overtone oscillator.

impedance matching

Throughout this article, a load impedance of 2700 ohms was used. The actual load will generally be some other value. Quite often this load will be 50 ohms resistive. In this case, the 2700-ohm load

resistance necessary for oscillation will have to be matched to the 50-ohm load. The simplest solution is the L network shown in **figs. 9 and 10**. The value for the series reactive element can be found with the following equation:

$$X_{series} = R_s \sqrt{\frac{R_p}{R_s} - 1}$$

where

R_s = series resistance of load

R_p = parallel resistance to be transformed

In this example

$$X_{series} = 50 \sqrt{\frac{2700}{50} - 1} = 364 \text{ ohms}$$

The value for the shunt element can be determined from:

$$\begin{aligned} X_{parallel} &= \frac{R_s R_p}{X_{series}} \\ &= \frac{(50)(2700)}{364} \\ &= 371 \text{ ohms} \end{aligned}$$

Fig. 11 shows the highpass matching network (**fig. 10**) applied to the common-gate oscillator shown in **fig. 6**. The 364-ohm reactance at 100 MHz is equivalent to 4.37 pF. Any error introduced by using the nearest standard capacitance value (5 pF) can be compensated for by L1 and C2. The shunt inductance of the matching network must be added to L1 for a new total inductance for L1. This is easily accomplished by adding the susceptance of $X_{parallel}$ and L1 [$-4.1 \text{ mS} + (-2.7 \text{ mS}) = -6.8 \text{ mS}$ or 147 ohms inductive reactance which at 100 MHz equals 234 nH]. The circuit in **fig. 11** was constructed and the test results for 100 MHz are: Value of C2 to start oscillation = 43 pF; final values after tuning are C1 = 5 pF; C2 = 25 pF; and L1 = 215 nH. Maximum power output = 9.2 milliwatts. With C2 set to 12 pF, the circuit oscillated with crystals ranging from 95 MHz to

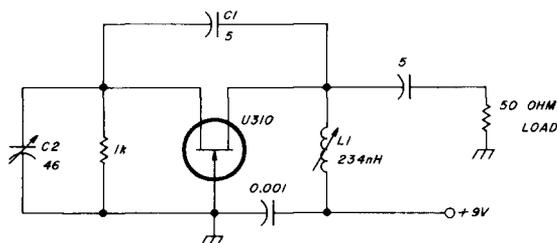


fig. 11. When the highpass L network is used for matching, the shunt inductor is combined with inductance in the oscillator. In this case, L1 is a slug-tuned 6.5 mm (1/4 inch) coil, wound with 6 turns of no. 18 AWG (1 mm) wire. The turns are spaced to occupy 12.5 mm (1/2 inch).

116 MHz. The value of L1 was changed to accommodate the different frequencies.

concluding comments

This article is not intended as a construction project. However, those circuits shown with inductance and capacitance values have been built and tested, and performed very close to predictions. This design technique has been used for many oscillator designs and has been found to be superior to the empirical approach, particularly if a programmable scientific calculator is available.

The techniques presented here appear to be equally valid at the lower frequencies. The data necessary are g_m , c_{rss} , c_{oss} , and c_{iss} . This information is almost always obtainable from the transistor data sheet. These parameters can be substituted in the common-source y parameters by assuming the input and output resistance of the fet is very high and can, therefore, be disregarded.

$$y_{11s} = 0 + j \left(\frac{1}{X_{c_{iss}}} + \frac{1}{X_{c_{rss}}} \right)$$

$$y_{21s} = g_m - j \frac{1}{X_{c_{rss}}}$$

$$y_{12s} = 0 - j \frac{1}{X_{c_{rss}}}$$

$$y_{22s} = 0 + j \frac{1}{X_{c_{oss}}} + \frac{1}{X_{c_{rss}}}$$

Low-frequency oscillator design is available from many sources. This technique might not be as usable as others, but it does allow a close approximation for low-frequency design.

Expressing an idea is often difficult for me. I am gratified by the many personal, and professional friends who have helped expand this idea and also provided the additional enthusiasm and necessary technical expertise, especially Ed Oxner, manager of Special Projects Engineering at Siliconix, Will Alexander, WA6RDZ, Earl McCune, WA6SUH, and Bonnie (The Boss).

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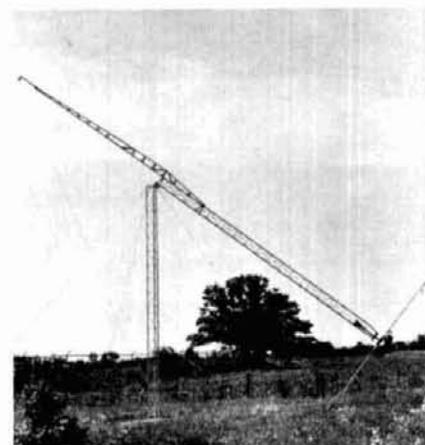
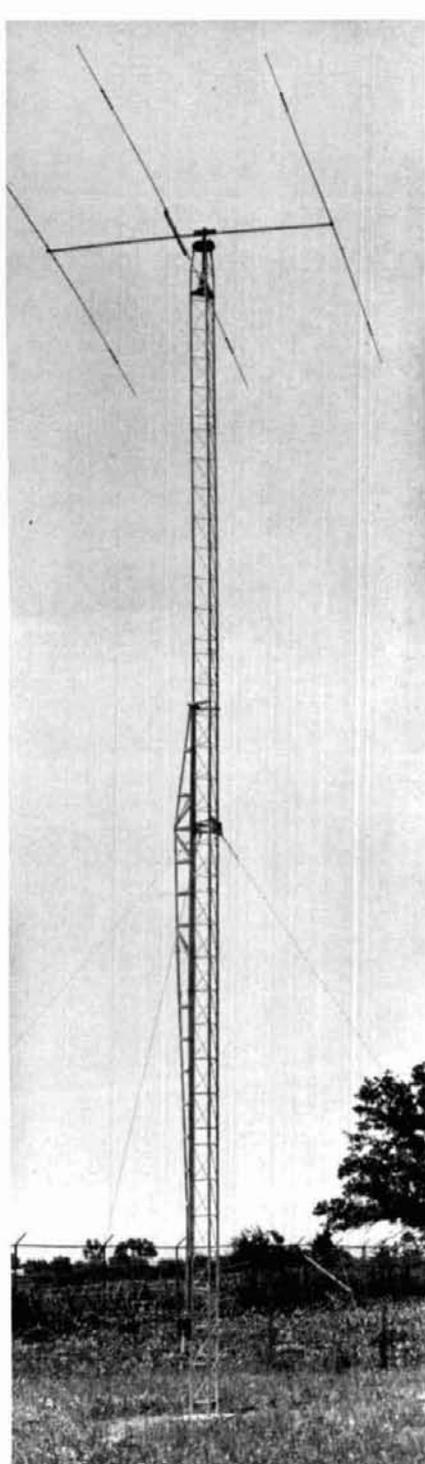
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visual aids

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Devices are getting smaller and smaller — consider these visual aids before you give up on a construction project with today's ICs

With the introduction of transistors and miniaturized circuits most everyone dealing with them has probably experienced difficulty seeing components and circuits because of their small size. Today, with integrated-circuit devices requiring even smaller printed-circuit-board design, the visual or seeing requirement is even greater.

the problem

When you consider the visual anomalies found in the general population such as nearsightedness, farsightedness, astigmatism, or combinations of these, and when you include problems of binocular-

ity (the ability or inability of the eyes teaming together, such as one eye receiving a larger image than the other), poor eye-muscle coordination, cataracts, and any ongoing or past history of ocular diseases, it's no wonder many of our IC projects end up with solder bridges, improper or unsoldered connections, and components mounted improperly. The fact is that many simply can't see well enough to avoid these pitfalls.

I recently completed a project that used 29 ICs, three of which were LSI devices, all mounted on four PC boards, three of which had circuit paths on both sides. Even though I have 20-20 vision at distance and near, and have never had any pathological problem causing visual impairment, I experienced great difficulty seeing my work. I therefore came to the conclusion that if I were having these kinds of problems with what's generally considered normal vision, there must be many electronic hobbyists who are having even greater problems. With this in mind, I offer the following observations to help in understanding these visual tasks and to allow most to see with better efficiency.

lighting and magnification

First, realize that vision requires light. But it has to be *useful* light; that is, not too dim (below the visual threshold) or too bright (above the threshold and therefore saturating the visual system, especially the retina). Second, visual acuity (how well you see) depends on the image size impinged onto the retina. Visual acuity is directly related to object size. The conclusion is that, if you have the proper lighting

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By Dr. Robert Sullivan, Optometrist, K9SRL,
410 NE 7th Street, Linton, Indiana 47441

conditions and magnification, visual acuity generally improves. However, there are limiting factors. For example, as the image size on the retina gets larger, the field of view gets smaller.

optical aids

If you're nearsighted (can't see well at a distance without glasses or contact lenses), you might do better visually, at the near distances used in electronic construction, without your glasses or contacts. Since a nearsighted eye without correction in place is in effect too strong, removing the spectacles has the same effect as looking through a magnifying lens. You'll notice that near objects (within about 20 cm [8 inches] of your eyes depending upon your prescription) look larger. However, this might not work if you have astigmatism, as your vision could be distorted.

If you're farsighted (can't see nearby objects well without glasses or contacts), you should wear your correction at all times for near electronic work. Since a farsighted eye is a weak eye without correction, wearing glasses or contact lenses, in effect, makes your eyes stronger. Also, far sightedness involves a problem with the eye's focusing mechanism and without correction, eye fatigue and headaches are more common.

If you normally wear glasses full time, and if they're the bifocal or trifocal type, you should wear them for near electronic work.

work glasses

You might consider having a special pair of glasses made especially for electronic work. I made a pair with one lens having a +16 diopter power and the other lens opaque, which forces me to use one eye only, since such a large prescription for both eyes creates a condition that makes binocularity impossible. This is a common problem with some of the available optical aids recommended for near visual tasks. These devices are usually binocular in nature (both eyes are used). To maintain this binocularity, weaker lenses are used and the resulting magnification is less.

I put the +16 lens on my right eye since I'm right-eye dominant. With these glasses, approximately 4X magnification is achieved and the field of view is about 7 cm (3 inches). The focal point is 6.25 cm (2.5 inches), so I must hold things close. I found that with these glasses and a handheld penlight (I use the popular disposable type), I could examine PC boards for errors in component mounting, find solder bridges, and perform general inspection with great ease. Just one thing more: these suggestions should be implemented *after* you've checked with your eye doctor.

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- **Frequency Coverage**— 1.8-54 MHz • **Line Impedance**— 50 ohm resistive • **Power Capability**— 2000 watts continuous
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solving RFI problems in home-entertainment devices

How to dispose
of RFI problems
quickly and easily —
ideas from an
overseas amateur
that will work
in your station

Radio-frequency interference (RFI) from amateur transmitters to television, broadcast, and hi-fi sets is still a problem. Most such cases of RFI can be attributed to *fundamental overload* of the home-entertainment device from the transmitter, especially if the transmitter is running high power (500 watts or more).

The remedies are well known for this type of RFI, but I'd like to summarize some of the cures:

1. Install a filter in the ac line to the device.
2. Install a 0.01- μ F capacitor across the speaker terminals.
3. Install an rf choke in the speaker leads. Such a choke can be made by winding the speaker leads onto a ferrite rod.
4. Make sure the home-entertainment device has a good rf ground.
5. Use shielded cable for the speaker leads, and ground the shield.
6. Install a good-quality 0.01- μ F capacitor between the device chassis and ground.
7. Ground the device's antenna coaxial-cable shield.
8. Wind 10-15 turns of the device's antenna coax cable onto a ferrite rod. This will form an effective rf choke.
9. Install a good highpass filter directly at the device's antenna terminals. Make sure the filter is shielded, and ground the shield.

By **John DeVoldere, ON4UN**, 215 Poelstraat,
B-9220 Merelbeke, OV Belgium

In all cases of reported RFI, I've eliminated the problem by one or more of the methods mentioned above without having to make any changes inside the home-entertainment device.

There's nothing original here so far, but I offer some other hints that you may find useful in your own RFI problem:

I always keep a selection of filters available for any RFI complaint. The set of filters includes a good-quality ac filter for the power line, a highpass filter for twin lead or coax-cable, ferrite-rod filters for speakers, an rf choke for coax cable, and some wires and clip leads for grounding purposes in tests.

responding to RFI complaints

Almost all interference complaints I receive are by telephone. Here's what I do. I ask the complainant to leave the phone off the hook and tell him I'll be right over to check the problem. Then I switch on my phone patch, turn down the receiver gain control so the receiver won't trip the VOX in my transmitter, and adjust my transmitter for VOX operation through the telephone. Then I grab my assortment of filters, hop into my car, and ring the complainant's door bell within minutes after his telephone call.

Now here's where you need some diplomacy. Don't be aggressive, but explain that you're genuinely interested in resolving the RFI problem as a mutual endeavor. Tell the complainant you're there to investigate the problem and want to use his telephone to put your transmitter on the air. He may raise an eye-

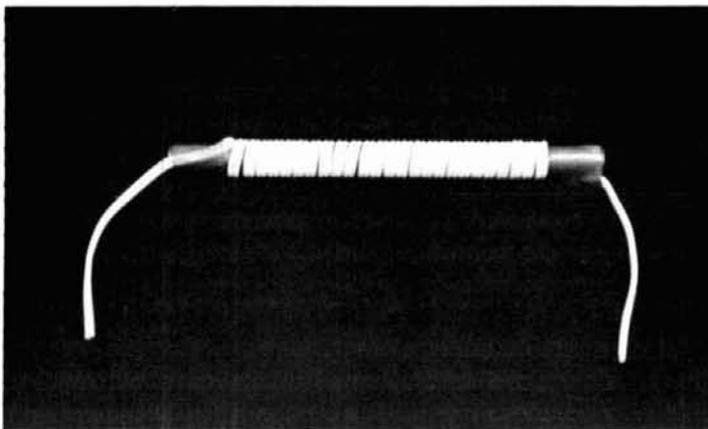


Ferrite or powdered-iron toroid can also be used to form in-line rf chokes with loudspeaker leads.

TVI suggestions

In cases of TVI complaints, proceed as follows:

1. Disconnect the device's antenna feedline. If TVI continues, install a filter in the device's ac line. If this doesn't help, you're probably in big trouble, because you have a case of direct pickup by the device.
2. If disconnecting the antenna feedline stops the TVI, try the following:
 - A. Install a highpass filter
 - B. Install a coaxial-cable rf choke (10-15 turns on a ferrite rod).
 - C. Ground the antenna coax-cable shield at the TV-set chassis (through a 0.01- μ F capacitor if necessary) to a good rf ground.



Loudspeaker lead can be wound into an rf choke by winding the lead around a ferrite antenna rod.

brow at this remark, but pick up his phone and put your transmitter on the air. Talking into the complainant's phone will trip your VOX, which makes it possible for *you* to make the diagnosis of the RFI problem, if any. No help from the outside, and you can do it right away.

suggestions for hi-fi RFI cures

If you have to deal with a complete hi-fi system including tuner, amplifier, turntable, and recorder-deck, there's only one logical approach that can be used.

Freedom of Choice

Disconnect all units from each other and from the ac line — except for the amplifier (just leave the ac cord and speaker leads connected). If interference still persists, disconnect the speakers and try the unit on headphones. If the interference disappears, the problem probably lies in pickup through the speaker cables (the RFI is being fed to the preamplifier stages through the audio feedback circuit).

Install the ferrite-rod filters in the speaker leads (10-15 turns). If necessary, connect 0.01- μ F capacitors across the speaker terminals to ground. In very stubborn cases you may have to go all the way and use shielded speaker cable and ground the cable shield to a good rf ground.



In some cases interference can be reduced by winding external audio input leads on toroidal cores.

If disconnecting the speaker cables doesn't kill the RFI, the pickup must be coming through the ac line, so you must install a good filter between the set ac input and the house wiring. Again, a good rf ground may have to be connected to the ac filter.

If the hi-fi amplifier plus the speakers (by themselves) don't show any interference, connect all other pieces of equipment, one-by-one, to determine where the RFI is appearing. If connecting the tuner brings up the RFI, try highpass (or coax rf-choke) filters on the antenna lead; then try an ac line filter. A similar approach can be used when connecting other pieces of equipment, such as tape decks.

summary

Using your phone patch to solve the RFI complaint quickly and independently; using a logical approach when checking the complainant's set; having a ready assortment of anti-RFI filters available at all times — all these will go a long way toward maintaining a good relationship with your neighbors. This approach will also ensure keeping your amateur station on the air at all times.

ham radio

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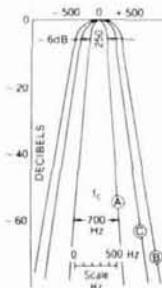
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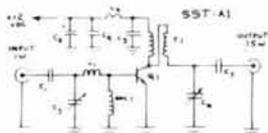
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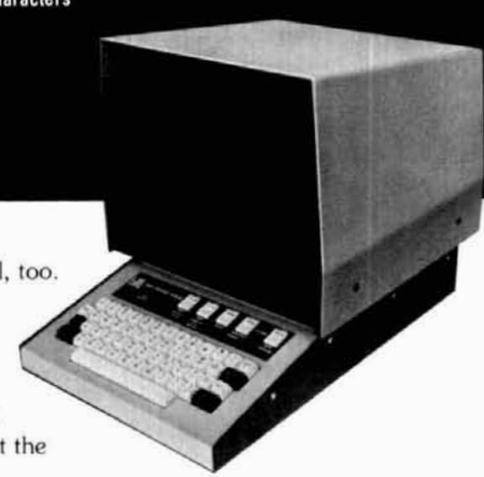
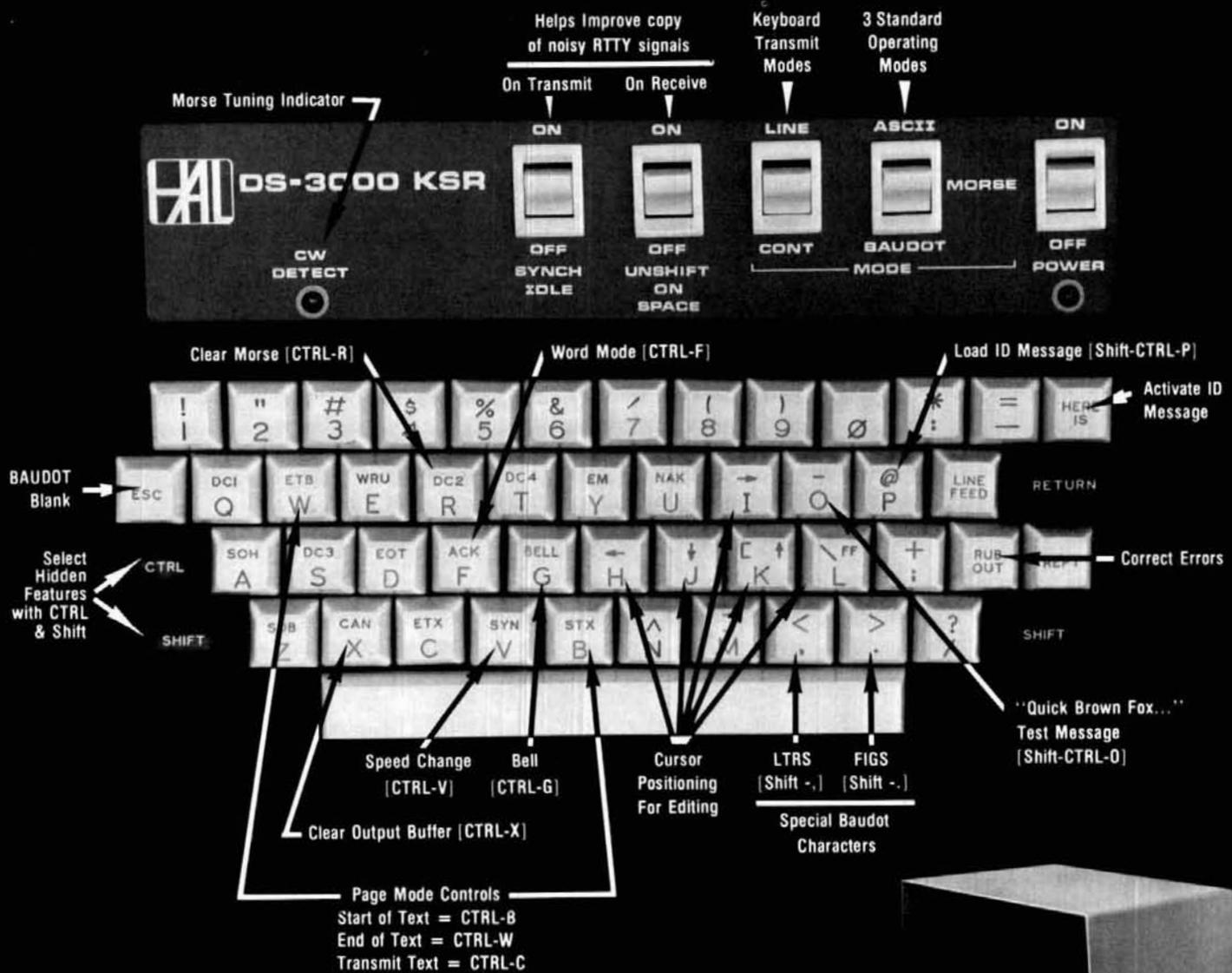
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illumination for lever action switch

Having problems reading thumbwheel or lever action switches in the dark? The new Heath HW2036 is a perfect example of a fine synthesized receiver at a price anyone can afford. One of its shortcomings is the lack of illumination on the lever action switches. Material used to make the light bar in **fig. 2** came from the junk box of my model railroad. However, the brass and lamps can be obtained at any hobby shop for less than \$2.00.

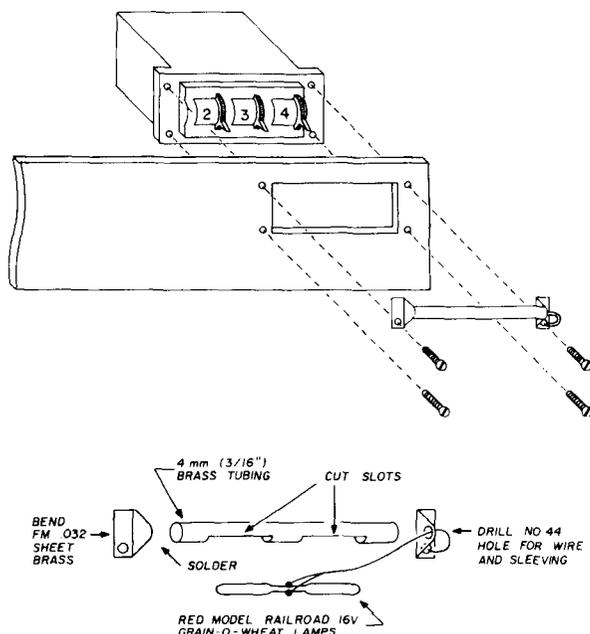


fig. 2. Adding illumination so you can see the HW2036 thumbwheel switches in the dark. Brass stock is available in most hobby shops.

Use a 3/16-inch (4mm) diameter brass tube cut to the width of the switch assembly so that the brackets, when soldered to the tube, will fit snugly against the side of the switch bezel and mount under the existing screw. Cut two slots for the lamps, large enough for the lamps to pass through after soldering and painting. Rough trim the brackets, then solder so that the slots aim down and the tube is spaced away from the panel the thickness of the protruding bezel,

approximately 2 mm (1/16 inch) or so. Bend brackets, drill holes for mounting screws and for the lamp wires in sleeving to pass through the right hand bracket; file smooth and paint.

The brackets should hold the tube high enough above the top of the switch to clear the upper most position of the lever switch. One screw is sufficient to mount, and the wires in sleeving are passed through the hole in the panel and connected to ground and the meter lamp wire. Use grain or wheat lamps and carefully insulate the wires at the base with five-minute

epoxy. I used red lamps to cut the reflected glare from the switch.

Another modification for the HW2036 which improves the operation is to change the back panel and replace the RCA phone jack with an SO239 connector. Enlarge the hole in the printed-circuit board to take the stub of the SO239 and replace the remote speaker RCA phone connector with a miniature phone jack.

Fred W. Snow, W2IFR

re-forming the oxide layer in electrolytic capacitors

Electrolytic capacitors, including computer grades, which have not been used for any length of time should not be subjected to full voltage without first re-forming the internal oxide layer. If this is not done, they may have high leakage which will result in rapid failure due to internal heating.

The oxide layer may be re-formed in the following manner: Connect the capacitor to a power supply set to the dc voltage rating of the capacitor with a series resistor to limit the short circuit current to around 10 mA. For example, for a 200-volt capacitor, the power supply would be set to 200 volts and the series resistor would be 20,000 ohms. If several capacitors are being re-formed simultaneously, they should not be directly paralleled. Instead, each capacitor should have its own current limiting resistor. This allows each capacitor to charge independently, at a rate dependent on its internal leakage. It also allows the voltage on each capacitor to be measured separately as an indication of its condition. It has been my experience when re-forming large numbers of surplus computer grade electrolytics in this manner that most of them will charge to close to full voltage in just a few minutes. A few will stabilize at a considerably lower voltage, indicating that they have a high leakage and really need the re-forming procedure.

I have not yet encountered a capacitor that would not charge to the same level as the others if left connected long enough, although this has taken as much as several days. When this finally happens, it means that the oxide layer is totally re-formed, and the leakage has dropped to a normal level.

John Becker, K9MM

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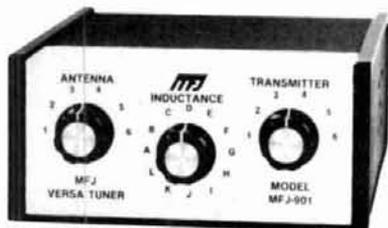


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MFJ antenna tuners

MFJ Enterprises has introduced a series of three new antenna tuners, using efficient, air wound coils, producing less loss than a tapped toroid.

The versatile, top-of-the-line MFJ-941 Versa Tuner II features built-in swr and dual range wattmeter (300 and 30 watts full scale), antenna switch for selecting two coax-fed antennas, random wire or balanced line, and tuner bypass, and a 1:4 balun for balanced lines. It handles up to 300 watts of rf power and matches dipoles, inverted Vees, random wire, verticals, mobile whips, beams, balanced lines, and coax lines from 1.8 through 30 MHz.



This beautiful little tuner is housed in a deluxe, eggshell white Ten Tec enclosure with walnut grain sides and is a compact 20 x 5 x 15 cm (8 x 2 x 6 inches). SO-239 coax connectors are provided for the transmitter input and all coax fed antennas, while quality, five-way binding posts are used for balanced line, random wire, and ground connections.

The MFJ-941 Versa Tuner II sells for \$79.95 and comes complete with mobile mounting brackets.

The MFJ-901 Versa Tuner also uses an efficient airwound coil, handles up to 200 watts, and has a built-in 1:4 balun for balanced lines. It matches all types of transmission lines (coax, balanced lines, random wire) are virtually all types of antennas from 1.8 through 30 MHz. It is an ultra compact 12.7 x 5 x 15 cm (5 x 2 x 6 inches) and uses SO-239 coax connectors and quality, five-way binding posts. The MFJ-901 Versa Tuner sells for \$59.95.



The MFJ-900 Econo Tuner is the same as the MFJ-901 Versa Tuner except that it does not have the built-in 1:4 balun for balanced lines. Price is \$49.95.

The MFJ-941 Versa Tuner II, MFJ-901 Versa Tuner, and the MFJ-900 Econo Tuner are all available from MFJ Enterprises, and each has a 30-day money back trial period. MFJ also provides a one year, unconditional warranty.

To order, call toll free 800-647-8660 or write to MFJ Enterprises, Box 494, Mississippi State, Mississippi 39762.

mobile antenna matcher

Barker & Williamson offers their new model AT-200 antenna matcher for matching two-meter amateur mobile transceivers to automobile a-m/fm broadcast receiver antennas. The model AT-200 is intended to provide the theft-foiling benefits of disguised and "hideaway" antennas at lower cost, and to eliminate the nuisance of constantly putting up and taking down a second antenna.

The new unit consists of a tunable matching network, an output indicator, and a selector switch, in a compact case. It is furnished with two

coaxial cables for connections to the entertainment radio and the two-meter rig. The connecting cables have a Motorola-type connector for the entertainment radio and a PL-259 connector for the two-meter rig. The front panel contains a tuning knob, a two-position (a-m/fm or two-meter) selector switch, and an output indicator light. The unit is supplied with a mounting bracket and installation instructions and will handle 100 watts from the transceiver.

The model AT-200 antenna matcher is available through Barker & Williamson distributors and dealers at an introductory price of \$22.50. For additional information, call or write Barker & Williamson, Inc., 10 Canal Street, Bristol, Pennsylvania 19007.

power/swr meter

A new power/swr meter, with an swr accuracy 100 times better than typical meters, is now available from Communications Power. The CPI Model WM-7000 uses a 30 dB directional coupler which provides



accurate swr readings to a ratio of 1.1:1. Most other swr bridges use 10 dB directional couplers, accurately reading to only 2.0:1, and even some of the more expensive meters only use 20 dB direction couplers.

The CPI WM-7000 reads peak or average power for accurate ssb power measurements, facilitating adjustment of microphone and speech compressor controls. The big 9 cm (3-1/2-inch) meter allows easy reading on three scales: 20, 200, and

SCR 1000

Standard of Comparison IN VHF FM REPEATERS!

2M & NOW 220 MHz!



Optional Cabinet \$130.00

Now Available with Autopatch!

Now Spec Comm has taken the hassle out of putting an autopatch repeater on the air! The SCR1000/SCAP is a fully self-contained 30 watt repeater with built-in autopatch and land line control. You simply plug in the phone line, hook up the duplexer, and you're on the air! The usual months of problems are eliminated! The SCR1000/SCAP has been meticulously engineered to provide the smoothest performing patch together with a positive land line control of the repeater. Just look at all these features:

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- Normal patch, or secure "reverse" patch
- 3 digit anti-falsing access — single digit disconnect
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- Patch access and repeater control — either over the air or over the land line

The SCR1000/SCAP is a complete Autopatch Repeater — fully assembled, set-up and checked-out in our lab. As with all Spec Comm products, all workmanship and components are of the very highest quality. The price? A very reasonable \$1585.00 — complete! (\$2080.00 w/WP641 Duplexer.) (For Rptr., w/o FL-6 Rcvr. Preselector, deduct \$85.00.) Get your order in A.S.A.P.

- The SCR 1000 — the finest repeater available on the amateur market . . . often compared to "commercial" units selling for 3-4 times the price! This is a 30Wt. unit, with a very sensitive & selective receiver. Included is a built-in AC Supply, CW IDer, full metering and lighted status indicators/control push buttons, crystals, local mic, etc. Also, jacks for emergency power, remote control, autopatch, etc.
- Custom options available: Duplexers, Cable, 'PL', HI/LO Power, Touch Tone Control, Racks, etc. Inquire.
- The Spec Comm Repeater System . . . a sound investment . . . available only by direct factory order. SCR1000 w/FL-6 Rcvr. Preselector \$1035.00. (\$950.00 w/o FL-6.) Commercial price somewhat higher.
- A Full Line of Repeater Boards & Assemblies Are Also Available: Inquire.

Call or write today and get the details! Send for Data Sheets!

SPECTRUM COMMUNICATIONS
1055 W. Germantown Pk., Norristown PA 19401 (215) 631-1710

1000 watts. The unit covers the 1.8 MHz-30 MHz frequency range (160 through 10 meters).

Further information on the CPI WM-7000, and the company's complete line of high technology American-made communications gear, is available from Robert Artigo, Communications Power, Inc., 2407 Charleston Road, Mountain View, California 94043.

lightweight headphones from Telex



To overcome extraneous noise, and at the same time provide extra comfort and performance, Telex Communications, one of the world's leading manufacturers of professional communications equipment, offers the radio amateur three lightweight headphone models. All three models have magnetic transducers with rising frequency response, making them particularly well suited for communications-quality audio use.

Headphones concentrate the signal at the ear, overcoming ambient noise and room acoustics. Also, they allow the operator to work without disturbing others, which is not possible with a conventional speaker. Additionally, the operator is able to more easily hear and understand weak signals than when using a speaker.

The HTC-2 is an over-the-head unit. This is Telex's lightest twin receiver unit, widely favored by pilots,



*our little boxes
replace a lot of cable!*

- select any of five antennas at the turn of a knob, with just one feedline and a control cable to the remote switching unit.
- saves coax, simplifies station layout.
- handles 4 kw p.e.p.
- other models to nine positions.
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- Independent receive selectivity. Built-in VSWR/watt/S-meter.

MATCHING ACCESSORIES:

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| ■ Remote VFO | AM filter | Mobile mount |
| ■ Matching Speaker | RTTY filter | Antenna tuner/wattmeter |
| ■ Digital read-out board | 2 CW filters, 300 Hz, 500 Hz | Wattmeter/VSWR bridge |
| ■ 120/240-volt wide-range power supply | Fan (for RTTY) | Aux-7 Expanded receive Capability |
| | Noise blanker | |

*includes capability for MARS, Embassy, and Government frequencies, and possible future Amateur-Band expansion. Receiver coverage continuous from 1.5 - 30 MHz, and 0 - 1.5 MHz with Aux-7.

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broadcasters, and hams. The dual magnetic receivers rest on the operator's temples with the sound fed to the ears through adjustable, ball-and-socket mounted tubular arms. This system permits either or both sound arms to be turned away for conversation without removing the entire headphone.

Two of the Telex units are under-the-chin style: the Model HMC-2 featherweight offers a magnetic driver-element positioned between the adjustable anodized-aluminum tone arms for optimum sound reproduction. The plastic eartips are removable. The Model HFC-91 provides a millisecond delay between ears by means of a magnetic element that channels the signal through the acoustic tubes, resulting in greater depth and clarity of the signal. The comfortable foam ear cushions are easily removed for cleaning and replacement. For single ear use, the driver element snaps onto the Nylon earloop.

For more information and a catalog of the Telex line of equipment, contact Otto Janssen at Telex Communications, 9600 Aldrich Avenue South, Minneapolis, Minnesota 55420.

two-meter transverter

Hamtronics has announced the VX2, a new 2-meter ssb transverter which you can use for Oscar mode J operation. Of course, it may be used for Mode A and simplex activity as well. The new model VX2 transverter is constructed on a pc board, as shown in the photo. The kit is easy to build and align, with convenient test points at each stage.

The kit intended for use with 10-meter ssb exciters, but some have been used with recycled 11-meter ssb units for inexpensive Oscar operation. Various frequency schemes are available to accommodate different types of exciters. The transverter requires only 5 mW of drive to

Loop Antenna

provide 2W PEP output. Many of the newer exciters have a low power output connection, and older ones can either be modified or used with an attenuator to provide the required drive. Perhaps the best feature of this new transverter kit is the economical price — only \$59.59.

Two linear power amplifiers are available for higher power output: A model LPA2-15 provides 15 watt PEP; and model LPA2-70 provides 70 watt PEP output. A cyclac case is also available for the transverter and PA as an option.

For more information write for a free catalog on these and other vhf and uhf kits, including preamps and converters for Oscar frequencies. Hamtronics, Inc., 182-F Belmont Road, Rochester, New York 14612.

high-power VHF mobile antennas

A line of mobile antennas with high power ratings, covering the six and two-meter frequency ranges, has been introduced by Antenna Incorporated. The six-meter antennas feature 200-watt loading coils; the two-meter antennas are available with either 150 or 200-watt loading coils. They are available with 3/4-inch toggle mounts, cowl mounts, and no-hole trunk-lip mounts. Also available are 100-watt models with either the same mounts, or with 3/8-inch snap-in, spring-clip gutter, and magnet mounts. Loading coils are tuned at the factory to achieve a standing wave ratio of 1.5:1 or less, and each antenna includes a cutting chart so the whips can be field trimmed to exact frequencies.

Each antenna features a plated stainless steel whip for low resistivity to combat skin effect, stainless steel impact spring, shock-resistant and weatherproofed PVC-wrapped loading coil, and 17 feet of coaxial cable with a soldered PL-259-type connector.

All Palomar Engineers products are made in U.S.A. Since 1965, manufacturers of Amateur Radio equipment only.



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Here is an exciting new device to improve your reception on 160, 80, the broadcast band, and on VLF.

It is well known that loops pick up far less noise than most other antennas. And they can null out interference. Now Palomar Engineers brings you these features and more in a compact, carefully engineered, attractive desktop package.

Unlike ordinary direction-finder loops, it tilts to match the incoming wave front. The result: Deep nulls up to 70 db. You have to listen to believe it!

Does the Loran on 160 give you a headache? The loop practically eliminates it. Broadcast station 2nd harmonic ruining your DX? Turn and tilt the loop and it's gone. Does your friend in the next block with his kilowatt block those weak ones? Use the loop and hear him fade out.

Loop nulls are very sharp on local and ground wave signals but usually are broad or nonexistent on distant skywave signals. This allows local interference to be eliminated while DX stations can still be heard from all directions.

The loops are Litz-wire wound on RF ferrite rods. They plug into the Loop Amplifier which boosts the loop signal 20 db and isolates and preserves the high Q of the loop. The tuning control peaks the loop and gives extra preselection to your receiver.

Plug-in loops are available for these bands:

- 150-550 KHz (VLF)
- 540-1600 KHz (Broadcast)
- 1600-5000 KHz (160 & 80 meters)

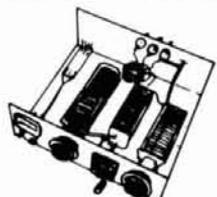
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Order direct. Loop Amplifier \$67.50; Plug-in Loop Antennas \$47.50 each [specify frequency band]. Add \$2 packing/shipping. Calif. residents add sales tax.

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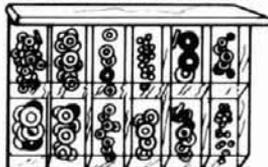
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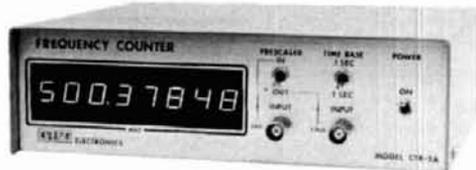
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500 MHz Kit CTR-2A-500K	\$249.95
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DAVIS ELECTRONICS 636 Sheridan Dr., Tona., N.Y. 14150 716/874-5848

The 200 watt two-meter antennas also feature Antenna Incorporated's new high-power coaxial cable. While the 150 watt high-band and 200 watt low-band antennas include RG/58-U cable, this cable cannot safely handle 200 watts of power on two meters. Antenna Incorporated's high-power cable has performance characteristics similar to RG-8/U, but in a smaller size, thus eliminating the problems of using the larger cable in mobile applications.

"These antennas also are part of Antenna Incorporated's professional land mobile line and have been designed to meet the needs of high power communications users," sales manager Randall Friedberg said. "They offer the amateur the best in antenna quality and dependability."

For further information on the company's complete line of communications antennas and accessories, contact Randall J. Friedberg, Antenna Incorporated, 23850 Commerce Park Road, Cleveland, Ohio 44122. Phone (216) 464-7075.

nye viking master key



The William M. Nye Company announces a new addition to their NYE VIKING line of products with the introduction of the MASTER KEY. Called the first major design change in telegraph keys in over 50 years, it is designed for the expert, yet is perfect for the beginner.

A prime feature of the *Master Key* is a contact assembly that is electrically isolated to keep the keying circuit separated from the base, the key arm assembly, and all exterior metallic parts. Thus, the shock hazard is greatly reduced. With its heavy

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Hi-Z \$15.00
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die-cast body and non-skid feet, the key does not need to be secured to the operating desk, nor does it require a sub-base. As with all NYE VIKING keys, the contacts are gold-plated silver for sharp, sure sending. The base of the *Master Key* has a black wrinkle finish with nickel plated exterior hardware. The adjustable action key arm is fitted with a Navy-style knob. The *Master Key* comes complete with 90 cm (3 feet) of two-conductor cord with attached plug. The list price is \$19.50

This new product joins the famous-for-quality NYE VIKING line, which includes Speed-X and Super-Squeeze Keys, Iambic Keyers, Low-Pass Filters, Antenna Impedance-Matching Networks, and Phone Patches. All are manufactured by Wm. M. Nye Company, Inc., 1614 - 130th Avenue NE, Bellevue, Washington 98005, and available at dealers nationwide. Write for more information.

new OSCAR book

What's OSCAR? It's a series of communications satellites designed and built by amateur radio operators. The best thing about them is their accessibility — anyone with an inexpensive receiver and simple antenna can hear the tiny satellites as they pass overhead, and persons holding amateur radio licenses can transmit voice and Morse-code signals up to them as well.

OSCAR is an acronym for Orbiting Satellite Carrying Amateur Radio. A new book published by The American Radio Relay League provides everything the interested electronics and space buff needs to know to track, listen to, and transmit through the spacecraft. This new book, *Getting to Know OSCAR — From the Ground Up* is a complete guide to the amateur satellites. Its 14 sections include an introduction to the exciting world of space communications, the equipment needed, a description of the brand-new OSCAR (and future ones now under construction), and a dis-

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cussion of their many practical uses, such as relaying of medical data over long distances. Each copy also contains an attractive four-color tracking device that makes finding the satellites a simple task.

Every amateur radio operator, short-wave listener, electronics enthusiast, and technician now has access to all he will need to know to enter the space age via OSCAR. *Getting to Know OSCAR — From the Ground Up* is available for \$3.00 plus 35¢ postage from Ham Radio's Communications Bookstore, Greenville, New Hampshire 03048. Order AR-OSC.

antenna center insulator from Hy-Gain



The engineering group at Hy-Gain Electronics has introduced a new center insulator unit for multi-band doublet antennas such as the Hy-Gain Model 380 (2BDQ). The new insulator, Model 157, has a built-in SO-239 for easy hook-up to a PL-259 on attaching coaxial cable. The insulator is molded from high-impact ABS plastic and all internal connections are fully weatherproofed and insulated with silicone for complete reliability under all environmental conditions.

All hardware is iridited to resist corrosion. Hardware is provided on each end for positive antenna attachment and an eyescrew is attached to the top of the insulator for stringing support wires for the antenna. The Model 157 will handle 1 kW average power and 2 kW PEP.

For further information on the new Hy-Gain doublet antenna center insulator or other Hy-Gain amateur products write: Hy-Gain Electronics, 8601 Northeast Highway Six, Lincoln, Nebraska 68505.

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CURTIS KEYSER CHIP now \$14.95

- 8043*; IC only (Qty pricing available) \$ 14.95
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*Apr 75 HR, Feb 76 QST, Radio Hdbk 75, ARRL Hdbk 77-78

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And it's rugged. The high-impact case protects a minimum number of component parts (47 in all), and they're all readily available from any of the worldwide Fluke service centers. Your 8020A is factory calibrated by NBS traceable equipment. And we guaran-

tee it'll live up to published specs for a full year.

The 8020A is a true field instrument, designed with a highly readable LCD display, and inexpensive 9V transistor battery power for continuous use up to 200 hours. Reliability, quality and value: *that's* Fluke tradition.

To get your hands on one, call (800) 426-0361, toll free. Give us your charge-card number and we'll ship an 8020A the same day. Or, we'll tell you the location of the closest Fluke office or distributor (where you can save by buying a ten-pack of 8020As for only \$1521*).

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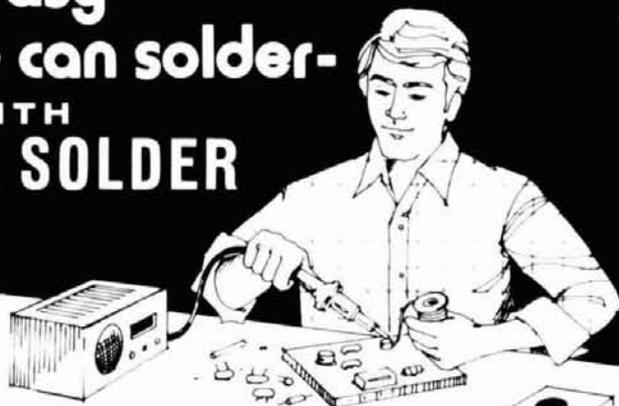
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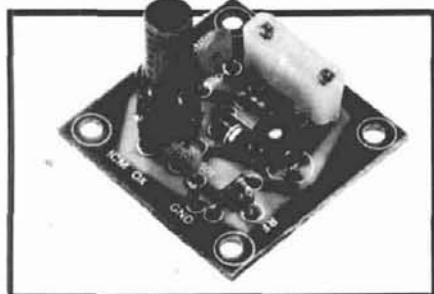


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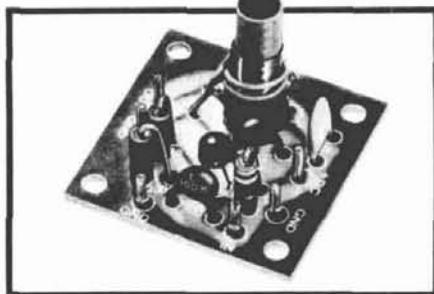


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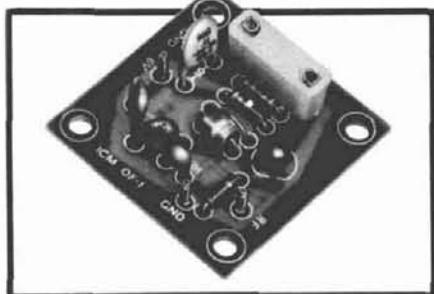


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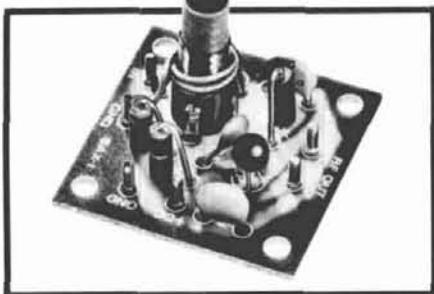


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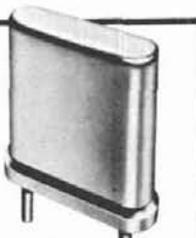
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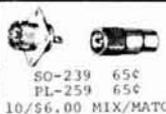
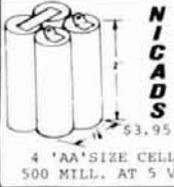
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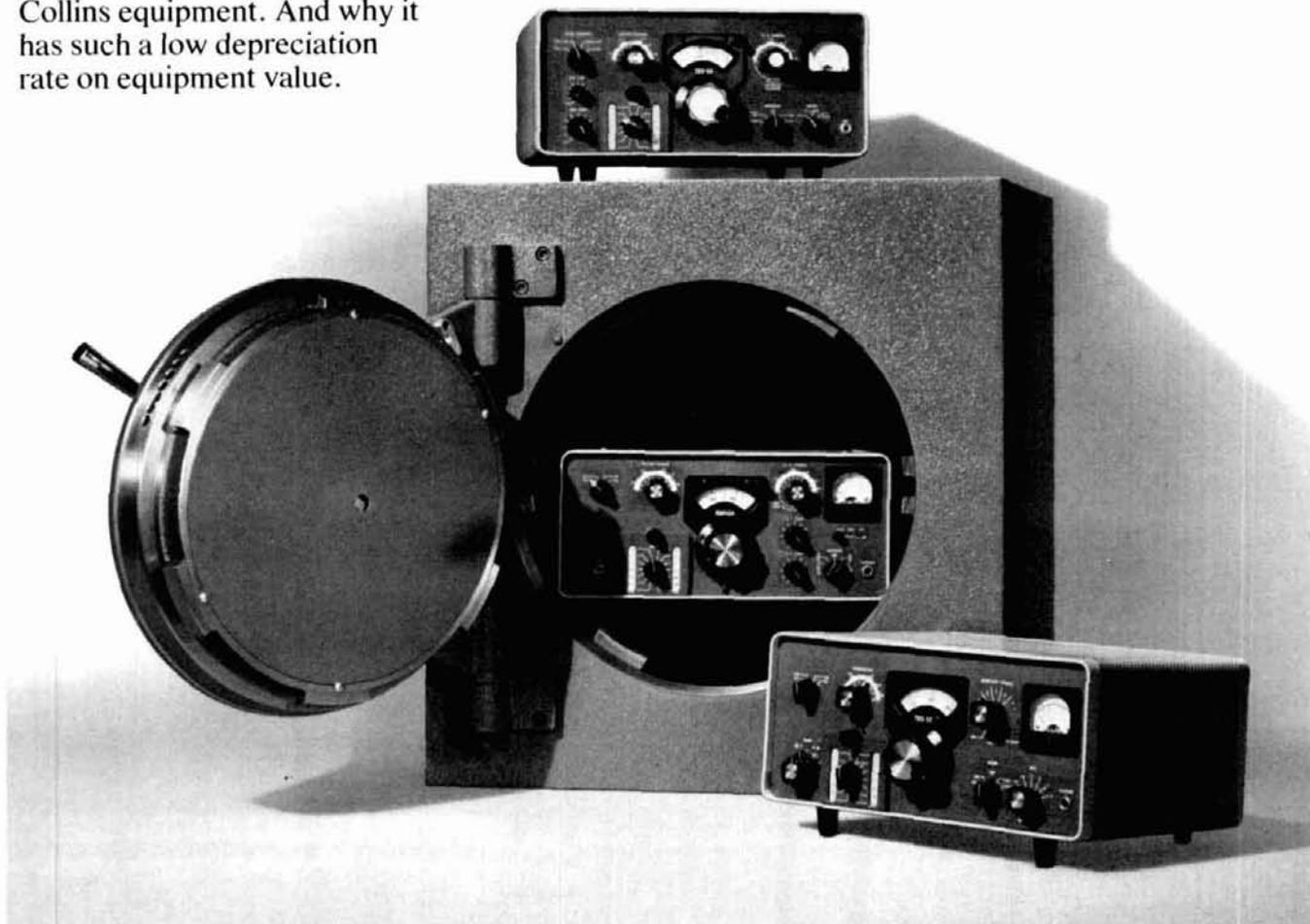
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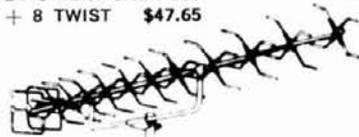
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FND510 69¢

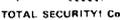
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Black plastic knob for above: **FREE**

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New!



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TOTAL SECURITY! Completely invisible ultrasonic (23KHz) sound beam works like a photoelectric beam but is unaffected by light, heat or noise. Separate Transmitter and Receiver can be used from 6 inches to 25 feet! A solid object breaking the beam causes an output to go low that will sink up to 150 MA to Drive a Relay, TRIAC, etc. Complete electronics are provided. Works on 12VDC (unregulated) and draws less than 100 MA. Use it for burglar alarms, object counters, automatic door openers, automatic door bells, electronic rat trap(?) and more.

COMPLETE KIT LESS CASES **21.50**

Optional entry delay and Alarm Timeout Circuit: **3.95**
will source or sink up to 200 MA DC.

LED'S JUMBO: RED 5/.89 GREEN 4/.89



MEDIUM: RED .15 MINI:
GREEN .16 RED .10
YELLOW .16

1.5V 10-30 ma

POWER SUPPLY KIT PS-14

- Better than 200MV load and line regulation
- Foldback Current Limiting
- Short Circuit Protected
- Thermal Shutdown
- Adjustable Current Limiting
- Less than 1% ripple.
- 15 amps 11.5 to 14.5V
- All parts supplied including heavy duty transformer.
- Quality plated fiberglass PC board.

Less Case, meters & jacks

39.95 UPS SHIPPING PAID!

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Provides cheap insurance for your expensive equipment. Trip voltage is adjustable from 3 to 30 volts. Overvoltage instantly fires a 25A SCR and shorts the output to protect equipment. Should be used on units that are fused. Directly compatible with the PS-12 and PS-14. All electronics supplied. Drilled and plated PC board. (Order OVP-1)

Public Notice!

THE PS-14 HIGH CURRENT POWER SUPPLY KIT HAS BEEN SELLING FOR OVER A YEAR. IN EXCESS OF ONE THOUSAND KITS HAVE BEEN SOLD AT THIS PRICE. WE WOULD LIKE TO GIVE OUR CUSTOMERS A CHANCE TO PURCHASE THE KIT AT THIS PRICE BEFORE A SCHEDULED INCREASE TO 43.00 IN SEPTEMBER.

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- Complete Electronics!
 - Chimes the hour (ie: 3 times for 3 o'clock)
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Custom case for above kit. Over 9" tall. **19.95**

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A fun EASY kit to assemble that emits an ear piercing 10 watt dual tone scream. Resembles European siren sound. Great for alarms or toys. Operates from 5-12VDC at up to 1 amp (using 12VDC x 8 ohm speaker). Over five thousand have been sold. All parts including PC board, less speaker. **ORDER WB-02**

2.50



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DEADLINE 15th of second preceding month.

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CQ and QST 1950-1975 ISSUES FOR SALE. Send SASE if ordering Ham Radio, 73, or other CQ and QST issues. One dollar minimum order and all issues cost 25¢ each, including USA shipping. Send chronological list and full payment to W6LS, 2814 Empire, Burbank, CA 91504.

RTTY — NS-1A PLL demodulator WT \$24.95 ppd. Price advance effective Sept. 1st. SASE for info. Nat Stinnette Electronics, Tavares, FL 32778.

WANTED: Crystal Impedance Meters, TS-683, TS-330, TSM-15 RFL Models 531, 459 or 1207. Glenn Kurzenkabe, K3SWZ, 403 Centerview Ave., New Cumberland, PA 17070.

HAM RADIO HORIZONS, a super new magazine for the Beginner, the Novice and anyone interested in Amateur Radio... What it's all about, How to get started, The fun of ham radio. It's all here and just \$10.00 per year. **HURRY! HURRY!** Ham Radio HOIRZONS, Greenville, NH 03048.

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WANTED: Davco DR230 receiver. Please state price and condition in first letter. Jim Fisk, W1HR, Ham Radio, Greenville, NH 03048.

SEE OUR AD in this issue, Pyramid Data, Page 140.

B&K TEST EQUIPMENT. Free catalog. Free shipping. Dinosaur discounts. Spacetron-CG, 948 Prospect, Elmhurst, IL 60126.

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

AUTHORIZED DEALER for DenTron, KLM, Larsen, Bearcat, etc., Big Catalog 201-962-4695 Narwid Electronics, 61 Bellot Road, Ringwood, N.J. 07456.

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

TELETYPEWRITER PARTS WANTED: for all machines manufactured by: Klenschmidt Corp., Teletype Corp. and Mita. Any quantity, top prices paid send list for my quote. Phil Rickson, W4LW, Rt. 6, Box 1103G2, Brooksville, FL 33512.

VERY in-ter-est-ing! Next 3 issues \$1. "The Ham Trader", Sycamore, IL 60178.

QSL CARDS 500/\$10. 400 illustrations, sample. Bowman Printing, Dept. HR, 743 Harvard, St. Louis, MO 63130.

HOMEBREWERS: Stamp brings component list. CPO Surplus, Box 189, Braintree, Mass. 02184.

WE MAY NOT HAVE a toll-free number, but we'll save you more \$\$\$ in the long run! This month's special: CDE Ham-III Heavy-Duty rotor for only \$114.95, prepaid anywhere in the Continental United States! We are also factory-authorized dealers for Yaesu, Drake, Kenwood, Ten-Tec, ICOM, DenTron, and many more. For the best deal around on the HF or VHF gear of your choice, write or call us today for our low quote. Try our personal, friendly Hoosier service and become one of our many happy and satisfied customers. **HOOSIER ELECTRONICS**, P.O. Box 2001, Terre Haute, Indiana 47802. (812) 238-1456.

HELP WANTED — Beehive Telephone Co. has need of a Ham/plot/FCC licensed technician telephone maintainer. We have, planned or under construction, 6 automatic dial switches spread out over 6,000 square miles of Utah from Lake Powell to the Idaho Border; four hops of microwave; 150 miles of toll line; all for just over 300 customers. We use low band/high band and UHF and have several private SSB HF circuits for our maintenance coordination. We do all our own construction, etc. Write telling W7NVY about yourself and your ambitions. ZIP: 84313.

WANTED TO BUY — Heath IT-121 Curve Tracer. Ernest Ruland, 69 Southworth St., Williamstown, MA 01267.

HRO7 COIL SETS WANTED below 1800 kHz, W600Q 985 W. 3rd Ave., Escondido, CA 92025. (714) 743-9822.

CLUB CALL PINS 3 lines 1-1/4 x 3-1/4 \$1.25 each. First Name, Call and Club. Colors Black, Red, or Blue with White letters. (Catalogue) Arnold Linzner, 2041 Linden St., Ridgewood, N.Y. 11227.

SUPER-BUY — 5000 mfd. @ 40 volt electrolytic cap. factory new and complete w/all hardware. \$1.35 ea. ppd.



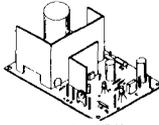
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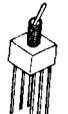
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WANTED — Millen I-F's 455 kc and 1600 kc — #61455 #61160 several needed. Also schematic for Misers Dream Rec. Owen Laughlin 1310 Pinecrest Dr., Ferndale, MI 48220.

CREATIVE QSL CARDS — New designs. Send 50¢, receive catalog, samples. WILKINS CREATIVE PRINTING, Box 787-2, Atascadero, CA 93422.

SELL — Swan Monobander 40A — Good — \$100.00. Henry Ankeny, 420 South 16, Clarinda, Iowa 51632. (712) 542-4715.

WANTED: Lorentz type 15 teleprinter 45.5 baud gear set. VE5KQ, D. Stephenson, #34 30 Summers Place, Saskatoon, Sask., S7H 3W4.

ELECTRONIC BARGAINS, CLOSEOUTS, SURPLUS! Parts, equipment, stereo, industrial, educational. Amazing values! Fascinating items unavailable in stores or catalogs anywhere. Unusual FREE catalog. ETCO-012, Box 762, Plattsburgh, N.Y. 12901. SURPLUS WANTED.

CODE PRACTICE TAPES — 100 High Frequency words which constitute 50% of all writing. For details write to: El Don Enterprises, P.O. Box 3404, Redondo Beach, CA 90277.

WANTED — Hallicrafters PS-150-120AC power supply. Hugh McKnight, BOQ 6-337, U.S. Naval Station, Charleston, S.C. 29408.

NICADS: 600mAh A size penlights, surplus, unused, excellent for portable, QRP. 6.5V five pack, \$3.75. Bramwell, 1330 SE Walnut, Hillsboro, OR 97123.

450 PROGLINE REPEATER including COR, Timers, Coupler. \$200.00. K7FE, Frank Exum, 1244 Granada, Casper, WY 82601. (307) 265-8083.

DRAKE "C LINE" STATION — R-4C, T-4XC, AC-4, MS-4, 4NB Blanker, all filters, nine Xtals, Fan — \$1300. Also Sherwood CF-125/8 Filter & "DX Engineering" RF Speech Processor, \$90 each. All excellent condition — one year old. WA6ERB, 451 Via Casitas #1, Greenbrae, CA 94904. (415) 461-3209.

CUSTOM printed and photo QSL's, very economical; free samples. Stamp appreciated. Stu, K2RPZ, Box 412, Rocky Point, NY 11778. (516) 744-6260.

WANTED: HP noise source 343A and 349A, Plug-ins for SINGER sweep-generator 650S, Tektronix Spectrum Analyzer 491. DL 7 QY, Claus Neie, Martin-Luther-Str. 121, D-1000 Berlin 62. Tel: 030/7824418.

NEW HIGH SCHOOL CLUB NEEDS EQUIPMENT DONATION, will provide tax deductible receipt, pay shipping. No Junk Please. Direct Response: Mark Cheek, Franklin County High School, Royston Road, Carnesville, Georgia 30521.

MINT CONDITION: Atlas 210 X or Triton II or IV. Also: Power supply. B. Levin, 24251 Oneida, Oak Park, Mich. 1-313-545-5939.

TELETYPE EQUIPMENT for beginners and experienced operators. RTTY machines, parts, supplies. Beginner's special: Model 15 Printer and demodulator \$139.00. Dozen black ribbons \$6.50; case 40 rolls 11/16 perf. tape \$17.00. Atlantic Surplus Sales, 3730 Nautilus Ave., Brooklyn, N.Y. 11224. Tel: (212) 372-0349.

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EXCLUSIVELY HAM TELETYPE 24th year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 35¢. \$3.50 per year. 1155 Arden Drive, Encinitas, Calif. 92024.

MANUALS for most ham gear made 1937/1970. Send only 25¢ coin for list of manuals, postpaid. HI, Inc., Box H864, Council Bluffs, Iowa 51501.

QSL FORWARDING SERVICE — 30 cards per dollar. Write: QSL Express, 30 Lockwood Lane, West Chester, PA. 19380.

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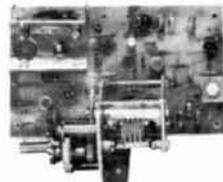
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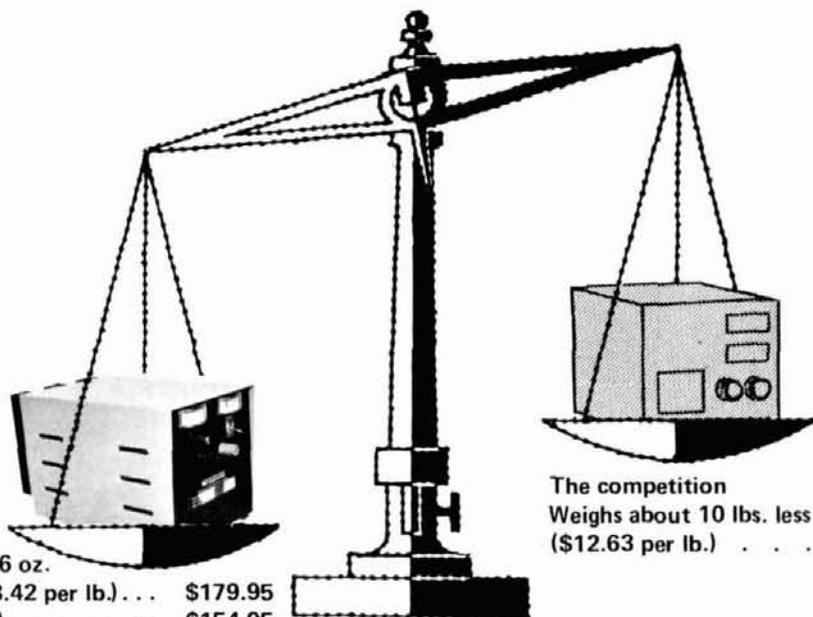
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Coming Events

ALASKA: ARRL Convention, Anchorage, August 26, 27. Write: ARRL Alaska Convention 78, Anchorage ARC, PO Box 1987, Anchorage, AK 99510.

WIMU (Wyoming, Idaho, Montana, Utah) The 46th Annual WIMU Hamfest is scheduled for August 4, 5, and 6, 1978 at Mack's Inn, Idaho; 25 miles South of West Yellowstone, Montana. Talk-in 146.34/94 and 3935. Advance registration: \$6.00 for adults and \$2.00 for children, before July 25th, 1978. Late/regular registration: \$7.00 and \$2.50. **SPECIAL PRIZE DRAWING FOR PRE-REGISTRATION.** Please send pre-registration to: WIMU Hamfest, 3645 Vaughn Street, Idaho Falls, Idaho 83401. Phone (208) 522-9568.

ANNUAL TEXAS VHF-FM SOCIETY SUMMER CONVENTION, hosted by the Houston Echo Society, August 4, 5, 6, 1978 at the Galleria Plaza Hotel off Interstate Loop 610 at Westheimer Road. Microprocessors/microcomputers, hidden transmitter hunt, OSCAR communications, VHF-FM activities. ARRL & FCC forums, open hospitality suite, ladies' activities, Astrodome-Astroworld tours for the kids, Exhibitors, and prizes. Saturday night banquet featuring Bill Tynan, W3XO, editor of QST's "World Above 50 MHz", as guest speaker. For information and reservations write FM Society Summer Convention, P.O. Box 717, Tomball, Texas 77375.

EYEBALL WITH FRIENDS July 16 Tri Club Hamfest — Talk in 34.94 & 52 Lookers \$2 Sellers \$3 — Police Pistol Range Allentown — Information SASE K3AI, RI Emmaus, Pennsylvania 18049.

AMATEUR COMPUTING 78, WASHINGTON, DC, July 22-23. Sheraton National Hotel, Columbia Pike and Washington Blvd., Arlington, VA. Registration at door \$5. Talk-in 147.81/21. AMRAD, PO Box 682, McLean, VA 22101.

VHF/UHF ANTENNA MEASURING CONTEST and free flea market July 23rd, 10:00 AM at Trenton State College, Trenton, NJ. Contact K2UYH or WAZZZF.

BYTE, Drink and be merry at the Tidewater Hamfest, Flea Market and Computer Show, Norfolk, Virginia. September 23-24. Over 60,000 sq. ft. of exhibit and flea market space. All indoors. All air-conditioned. Write TRCI, P.O. Box 9371, Norfolk, Virginia 23505.

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flea market

THE 15TH ANNUAL INTERNATIONAL HAMFEST will be held July 8th and 9th on the Canadian side of the International Peace Gardens. All hams and interested persons are invited to attend.

FOX RIVER RADIO LEAGUE HAMFEST New Location: Indoors — Kane Co. Fairgrounds, St. Charles, IL. — Sunday, August 27th. Tickets: \$2.00 at gate — \$1.50 advance. Contact: Don Berridge WB9PAC, 2303 Deerfield Way — Geneva, IL 60134.

MANSFIELD, PA. — The Tioga County PA ARC Hamfest will be held Saturday, August 26th starting at 9:00 AM at the Tioga Co. Fairgrounds on Rt. 6 between Wellsboro and Mansfield, PA. The \$2 admission is good for all special programs and the XYL and children are free. In addition to the usual Flea Market and displays a bingo table and other items of interest will be available for the ladies and the PA Grand Canyon is within a short distance. Talk-in on 1979, 52 Sim. and CB 5. For more information write to Denny Voorhes, WA3FWQ, RD #2 - Box 117A, Millerton, PA 16936.

WASHINGTON: Fourth Annual Spokane Hamfest Saturday, August 12th and Sunday, August 13th. Eastern Washington University, Cheney. Microwave seminar, Planetarium shows, activities for ladies. Exhibits & Flea Market. Preregistration prize: 60' Rohn 25 tower. Main prize: ICOM 22S Two-meter transceiver. Ladies prize: Microwave oven. Other prizes and raffles. Dorm housing and camper space available. Preregistration — \$9.00 per person includes both days & Banquet Saturday night. For information, write: Spokane Hamfest, P.O. Box 3606, Spokane, Washington 99220.

PENNSYLVANIA: Tri-Club Hamfest, Sunday, July 16, 8:00 AM. Allentown Police Academy, Lehigh Parkway South, Allentown. Lookers: \$2.00, Sellers: \$3.00. Refreshments. Door Prizes. Bring tables and power. Talk in .34/94 & .52. Local repeaters .10/70, open auto patch .145/745. Details SASE F. J. Herman, K3AI, R. 1, Box 104, Emmaus, PA 18049.

CHARLESTOWNE HAMFEST July 8-9, 1978. Gaillard Municipal Auditorium, 77 Calhoun Street. Doors open at 8:00 AM. \$3.00 per table per day. FCC Exams on Saturday 8:00 AM. Talk in Frequency 34/94. YL Activities planned. Social Room Saturday night at 7:30 at the Gaillard Municipal Auditorium. Hamfest registration required for entry — \$3.00. Free refreshments. P.O. Box 4555, Charleston, South Carolina 29405.

VIRGINIA: Amateur Computing 78 microcomputer festival. July 22-23. Sheraton National Motor Hotel, Columbia Pike and Washington Blvd., Arlington. Registration: Door — \$5.00/2 days. Advance: \$4.00. Spouse and children free. Saturday night banquet, \$14.00 Advance: \$12.00 per person. Sends checks to AMRAD, P.O. Box 682, McLean, VA 22101.

INDIANA: Wabash Valley ARA 32nd annual Turkey Run Hamfest. Vigo County Fairgrounds mile south I-70 on U.S. 41. Overnight campers ONLY open Saturday, July 15, 12 noon EST. Public, Sunday, July 16, 8:00 AM Free outdoor Flea Market, covered \$3.00. XYL Bingo. Refreshments. Shopping mall nearby. Advance tickets \$1.50, 4 for \$5.00. Gate: \$2.00, 3 for \$5.00. Under 12 free. For tickets and information: SASE to WVARA Hamfest, P.O. Box 81, Terre Haute, IN 47808.

TEN-TEN INTERNATIONAL NET SUMMER QSO PARTY Starts: 0000 GMT July 15, 1978 Ends: 2400 GMT July 16, 1978. Open to all amateurs but only members eligible for awards. All contacts to be made on 10 meters, any mode, a station to be counted only once. Exchange: Name, QTH, 10-10 number. (Be sure to log date and time of each contact.) Scoring: 1 point for each contact; add 1 point if with a member. Maximum: 2 points. No multipliers. Give the name of your Chapter. Awards: 1st and 2nd place certificates to each U.S. Dist., KL7, KH6 and U.S. Pacific Islands; VE Dists.; Central America and Caribbean; South America; Europe; Africa and S. Atlantic; Asia; Australia, New Zealand and S. Pacific. Members only: Send log to Grace Dunlap, K5MRU, Box 13, Rand, Colorado 80473, no later than August 30, 1978. Results will be published in the Net Fall Bulletin.

VHF SPACE NET CONTEST From 6PM Saturday, July 15th to 9PM Sunday, July 16th, local times. This event commemorates the 9th Anniversary of Apollo eleven, Man's First Landing and Walk on the Moon. Activity will be on 50, 144, 220 MHz etc., in all modes, EXCEPT REPEATERS. Categories: Class 1, 10 to 500 watts; Class 2, 25 to 100 watts; Class 3, 5 to 25 watts; Class 4, 1 to 5 watts; Class 5, C.W. only, with any power; Class 6, XYL only, with any power; Class 7, Club participation. For more info, write: VHF Space Center, K4AWS, Box 15, Sumterville, FL 33585. SPECIAL BONUS SURPRISE FOR ALL STATIONS WORKING SPACE NET CENTER.

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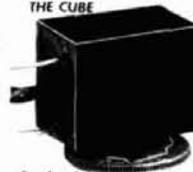
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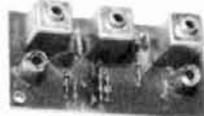
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P15 Kit \$18.95
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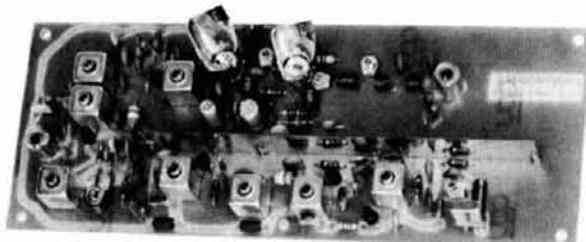
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T80-150; 140-175 MHz, 20-25W output \$79.95
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C144	144-146	28-30
C145	145-147 (OSCAR)	28-30
C146	146-148	28-30
C110	Aircraft	28-30
C220	220 band	28-30
Special	Other i-f & rf ranges available	

MODEL	RF RANGE (MHZ)	I-F RANGE
C432-2	432-434	28-30
C432-5	435-437 (OSCAR)	28-30
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flea market

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HAMFESTERS 44TH ANNUAL PICNIC AND HAMFEST, Sunday, August 13, 1978 at Santa Fe Park, 91st and Wolf Road, Willow Springs, Illinois, Southwest suburb of Chicago. Exhibits for OM's and XYL's, FAMOUS SWAP-PERS ROW. Tickets at gate \$2.00, Advance \$1.50. For Hamfest info or Advance Tickets (send check or money order — SASE appreciated) to Bob Hayes W9KXW, 18931 Cedar Ave., Country Club Hills, IL 60477.

OHIO — Wood County 14th annual Ham-A-Rama Sunday, July 16, Fairgrounds, Bowling Green off I-75. 10:00 AM. Admission/parking free. Tables \$3.00 or 8 ft. space \$2.00. Advance — dealers only. Tickets: \$1.50 advance — \$2.00 door. Main prize and door prizes. K8TH talk-in 146.52 simplex. Write: Wood County ARC, Erick Willman, 14118 Bishop Rd., Bowling Green, OH 43402.

MISSOURI: Indian Foothills ARC Third Annual Hamfest, Sunday, July 23, 1978 at Saline County Fairground, Marshall. Registration \$2.00 advance: \$2.50 door. Refreshments. Flea Markets for the OM and XYL. Tables for a small charge. For information and advance tickets contact Jim Little, WD0BPG, 405 East Rosehill, Marshall, Missouri 65340. (816) 886-8583.

ILLINOIS: The Big Thunder Hamfest, July 30, 1978, Boone County Fairgrounds, Belvidere. 8:00 AM and 3:00 PM. Tickets \$1.50 advance, \$2.00 gate. Campers \$2.00 additional. Talk in on 94 simplex. Write to Mike George K9ORU, 6159 Broadview Ave., Belvidere, IL 61008.

OHIO: NOARSFEST — Saturday, July 8, Lorain County Fairgrounds, Wellington. Prizes. Blacktopped flea market area, \$1.00 per space. Tickets: \$1.50 (before July 1), \$2.00 at gate. Info & Tickets: NOARSFEST, P.O. Box 354, Lorain, OH 44052. Directions: 146.10/70. Mobile check-in for prizes: 146.52.

OHIO: Kent State Salem hamfest July 23, 1978. Main door prize: Ten-Tec #540 transceiver. Hot air balloon, ramp for wheelchairs, free parking. Wives and children under 12 free. Recreation facilities. XYL drawing. Open 9 AM, Main drawing 3 PM. Admission \$2.00, fleamarket \$1.00, tables \$5.00. Talk-in: 146.10/70. Info: W8JPG, 147.27. Milhoan Electronics, 1128 W. State St., Salem, Ohio 44460. (216) 337-9275.

ILLINOIS: The Sangamon Valley Radio Club of Springfield, Third Annual Hamfest Sunday, September 24th, Sangamon County Fairgrounds in New Berlin, 16 miles west of Springfield. Hear Hugh Vandegrift WA4WME talk on the Clipperton DX-pedition! Various exhibits, kids activities and food. Camping. First Prize — Bearcat 210; Tickets: \$1.50 advance, \$2.00 at gate. Information — Al K9QFR; Tickets — Carole WB9QWR, write C/O 1025 S. Sixth, Springfield, Illinois 62703.

INDIANA — Indianapolis Hamfest, Sunday, July 9, 6:00 AM — 4:30 PM, Marion County Fairgrounds. Commercial Exhibiting, Flea Market, Camping with hookups. Write — Indianapolis Hamfest, P.O. Box 1002, Indianapolis, IN 46206.

RHODE ISLAND QSO PARTY Two periods (GMT) Sat. July 22 1700 — 0500 Sun. July 23; Sun. July 23 1300 — 0100 Mon. July 24. Sponsored by the East Bay Amateur Wireless Association. R.I. Stations work other R.I. Stations and the rest of the world. Others work R.I. Stations. The same station may be worked once per band and mode. Exchange: Send RS(T), QTH, County for R.I. stations, State, Province or Country for all others. CW: 1810, 3550, 3710, 7050, 7110, 14050, 21050, 21110, 28050, 28110. Phone: 3920, 7260, 14300, 21360, 28600, 50.3, 145.1. Use of FM simplex is encouraged. (No repeaters). For more info, write: East Bay Amateur Wireless Association, P.O. Box 392, Warren, Rhode Island 02885. Include SASE for results.

MARYLAND — Hamfest, Baltimore R.A. TV Soc. Sunday, July 30. 8:00 AM — Howard County Fairgrounds. Rt. 144, 15 mi. west of Baltimore. Off I-70@Rt. 32. Tickets \$2, Tailgating \$2, Tables — Advance \$3, Door \$4. Prizes, Flea Market, Refreshments, Contests. Talk in: 146.16/76, 147.63/03, 146.52, 52.525. XYL's & under 12 Free. Write: BRATS, Box 5915, Baltimore, MD 21208.

PENNSYLVANIA: Two Rivers ARC 14th Annual Hamfest July 23. Green Valley Fire Department Fairgrounds off US Route 30 near E. McKeesport. Flea Market. For information: Andrew Salitros, W3OFM, 2901 Stewart St., McKeesport, PA 15132.

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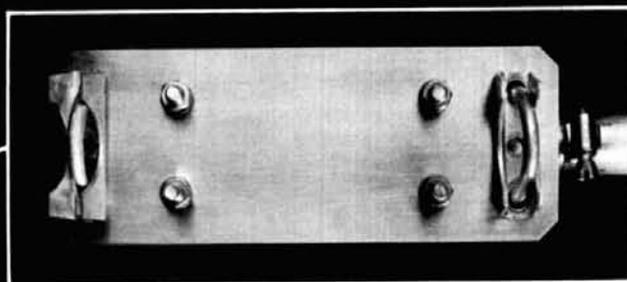
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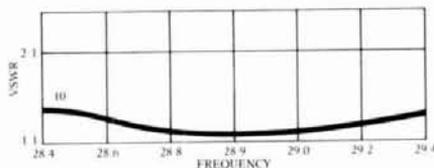
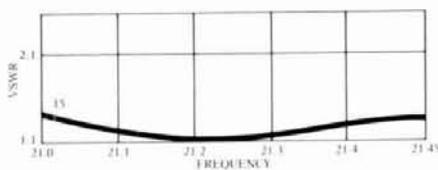
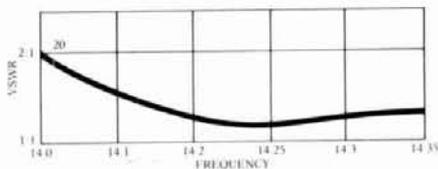
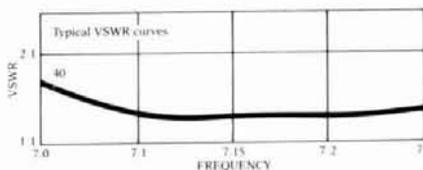
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flea market

MONTANA: International Glacier-Waterton Hamfest July 15 and 16. Three Forks Campground, 10 miles E. of Essex, Mt. on U.S. 2. 9:00 AM MST. For information: International Glacier-Waterton Hamfest, P.O. Box 2225, Missoula, MT 59806. (406) 543-5033.

MASS: Northern Berkshire Hamfest July 8th and 9th Cummington Fair Grounds, Cummington. Free overnight camping, tech talk & demos, dealers. Flea Market \$1 Admission \$4 with spouse \$6 Advanced \$3 and \$5. For info write: WA1ZNE Hildy Sheerin, 89 Greylock Terrace, Pittsfield, MA 01201.

OHIO: The Tusco Radio Club and Canton ARC's 4th Annual Hall of Fame Hamfest, Stark County Fairgrounds, Canton. Free display room for manufacturers and distributors of electronic equipment. Regular gate charge. Free camping. Write: Max Lebold, WA8SHP, Chairman, Hall of Fame Hamfest, Box 3, Sandyville, OH 44671. Phone: Business 216-821-5740 or home 216-866-3714.

MICHIGAN: 30th Annual U.P. Hamfest, Saturday, July 29 and Sunday, July 30. Dickinson County Armory, on M-95 Kingsford. 9:00 AM. Tickets: \$2.50 advance - \$3.00 door. Saturday banquet \$6.50. Reservations by July 1. Prizes galore. Talk in 148.25/85 and 3922. For information write UPHAMFEST 78, Box 2056, Kingsford, MI 49801.

CORA HAM HOLIDAY '78 OKLAHOMA: Central Oklahoma Radio Amateurs will present Ham Holiday '78, July 28-29-30. Location: Lincoln Plaza Forum, 4345 North Lincoln Boulevard, Oklahoma City. Pre-registration before July 14 \$3.00, \$400 at door. Non-commercial flea market tables FREE. Commercial Exhibitors contact K5MB, (405) 787-9545 or 787-9292. Many prizes, special pre-registration prize. Mail pre-registrations to HAM HOLIDAY '78, P.O. Box 14604, Oklahoma City, OK 73113.

MICHIGAN: 3rd annual Straits Area Radio Club Swap and Shop, Saturday, August 5. Emmet County Fairgrounds, Charlevoix Avenue, Petoskey. 9:00 AM - 3:00 PM. Talk-in 146.52. Admission: \$1.50. Food, prizes. Campsites. SARC - WB1ZS, Box 416, Pelston, MI 49769.

8TH S.A.R.T.G. WORLD-WIDE RTTY CONTEST 1978 We have the great pleasure to invite you to join the 8th W/W RTTY Contest run by the Scandinavian Amateur Radio Teletype Group. 1: 0000-0800 GMT Sat. Aug. 19; 2: 1600-2400 GMT Sat. Aug. 19; 3: 0800-1600 GMT Sun. Aug. 20. Use all bands 3, 5, 7, 14, 21, 28 MHz. a) Single Operator. b) Multi-operator, single transmitter. c) SWL's. RST and QSO no. To the top stations in each class, country, W/K, VE/VO and VK call district. For more information, write: OZ2CJ C.J. Jensen, Meisnersgade 5, 8900 Randers, Denmark.

THE ICHN-MARAC 10th ANNUAL CONVENTION will be held July 6, 7 and 8 at the Holiday Inn 4545 N. Lindburg Blvd., Bridgeton, MO. This convention will involve Amateur Radio operators from every part of the United States and several DX stations. All sharing a common interest in county hunting and mobile operation. The election of MARAC officers, meetings, workshops and some most important social activities will highlight this years convention. Of course there will be prizes, and awards to add spice to the festivities. Those interested in attending this years convention can send for information to Convention Director Jim Glascock WQFF, 3416 Manhattan Ave., St. Louis, MO 63143.

19TH ANNUAL NEW JERSEY QSO PARTY - July 29 - 31. Contest is from 2000 UTC Saturday, July 29 to 0700 UTC Sunday, July 30 and from 1300 UTC Sunday, July 30 to 0200 UTC Monday, July 31. For details on contest write: Englewood Amateur Radio Association, Inc., Post Office Box 528, Englewood, New Jersey 07631.

Stolen Equipment

STOLEN FROM AIRLINE BAGGAGE, probably either in Minneapolis/St. Paul or San Francisco, Wilson WE-800 s/n 12521811 with 10 white "no brand" Ni-Cad batteries inside, flex antenna with UHF ell connector and UHF to BNC connector. Also homebuilt battery charger with 723 IC. Mitt Nodacker, WA7TFE, Box 2632, Pocatello, ID 83201.

STOLEN EQUIPMENT: 1. KLM 160 watt amplifier, no I.D. 2. Black Heath 2036 with Micoder and several obvious modifications: hi-low power selector on squelch knob, variable power on internal potentiometer, RCA plug replaced with SO-239, Social Security No. 350-30-1717 etched in foil on transmitter board. Darrel Dorsett, K9JKZ, Kankakee Area Career Center, Rt. 2, Road 100-W, Bourbonnais, IL 60914.

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Collins 32S3, Transmitter, rnd., exc. \$850
Collins 75S3, Ham receiver, vy gd \$425
Collins 75S3B, Ham receiver, vy gd \$695
Collins 75A4, Ham receiver, vy gd \$425
Collins 51S1, 2-30MHz rcvr Special \$425
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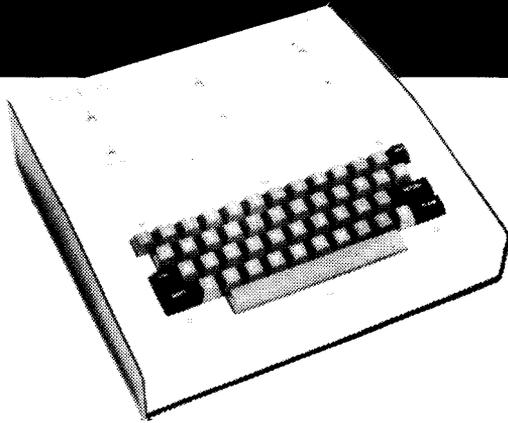
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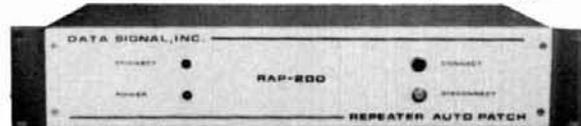


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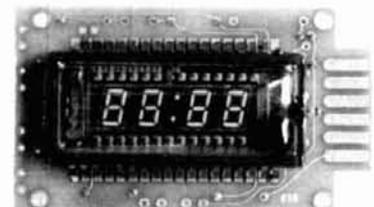
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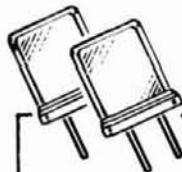
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146.04/146.64
146.07/146.67
146.10/146.70
146.13/146.73
146.16/146.76*
146.19/146.79
146.22/146.82*
146.25/146.85
146.28/146.88*
146.31/146.91
146.34/146.94
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146.46/146.46
146.49/146.49
146.55/146.55

147 MHz PAIRS

Transmit&Receive

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147.93/147.33
147.87/147.27
147.84/147.24
147.81/147.21
147.78/147.18
147.75/147.15
147.72/147.12
147.69/147.09
147.66/147.06*
147.63/147.03*
147.30/147.30
147.42/147.42*
147.48/147.48*
147.51/147.51*
147.54/147.54*
147.57/147.57*

220 MHz CRYSTAL PAIRS

Transmit/Receive

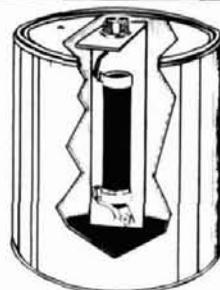
222.22/223.82 MHz
222.34/223.94 MHz
222.50/224.10 MHz
222.54/224.14 MHz
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223.18/224.78 MHz
223.22/224.82 MHz
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223.26/224.86 MHz
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Transmit/Receive

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Note: Items marked * stocked for Wilson only; * stocked for ICOM only

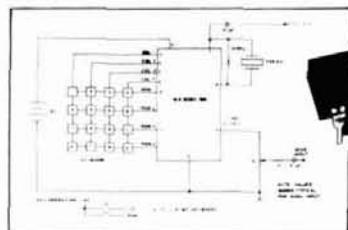


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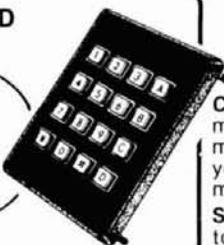
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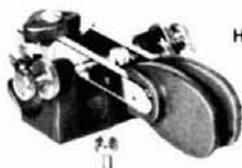
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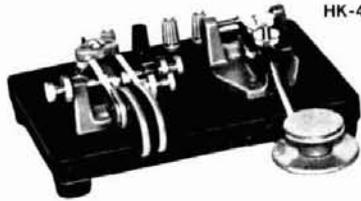
HK-1



HK-3



HK-2



HK-4

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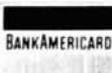
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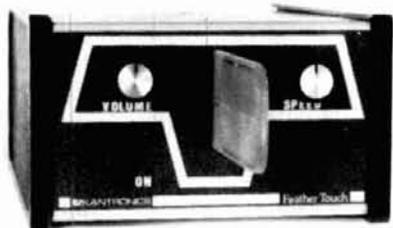
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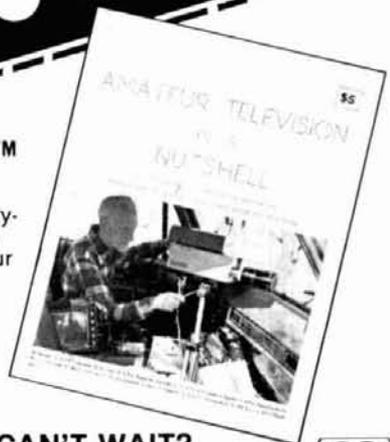
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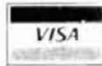
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7401	.15
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7403	.15
7404	.10
7405	.25
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7414	.75
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74LS90	.55
74LS93	.55
74LS107	.40
74LS123	1.00
74LS151	.75
74LS153	.75
74LS157	.75
74LS164	1.00
74LS193	.95
74LS367	.75
74LS368	.65

4069/74C04

4071	.25
4081	.30
4082	.30
MC 14409	14.50
MC 14419	4.85
4511	.95
74C151	1.90

MCT2	.95
8038	3.95
LM201	.75
LM301	.45
LM308 (Mini)	.95
LM309H	.65
LM309K (340K-5)	.85
LM310	.85
LM311D (Mini)	.75
LM318 (Mini)	1.75
LM320K5(7905)	1.65
LM320K12	1.65

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LM320T12	1.65
LM320T15	1.65
LM324N	1.25
LM339	.75
7805 (340T5)	.95
LM340T12	.95
LM340T15	.95
LM340T18	.95
LM340T24	.95
LM340K12	1.25

LM340K15	1.25
LM340K18	1.25
LM340K24	1.25
78L05	.75
78L12	.75
78L15	.75
78M05	.75
LM373	2.95
LM380 (8-14 PIN)	.95
LM709 (8,14 PIN)	.25
LM711	.45

LM723	.40
LM725N	2.50
LM739	1.50
LM741 (8-14)	.25
LM747	1.10
LM1307	1.25
LM1458	.65
LM3900	.50
LM75451	.65
NE555	.35
NE556	.85
NE556	.95
NE566	1.25
NE567	.95

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9301	.85	95H03	1.10
9309	.35	9601	.20
9322	.65	9602	.45

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1702A	4.50	8224	3.25
MM5314	3.00	8228	6.00
MM5316	3.50	8251	8.50
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2102L-1	1.75	8T13	1.50
2114	9.50	8T23	1.50
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146.07-67	146.28-88	147.65-06	147.87-27
146.10-70	146.31-91	147.65-09	147.90-30
146.13-73	146.34-94	147.72-12	147.93-33
146.16-76	146.37-97	147.75-15	147.96-36
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PLUS: The following simplex channels are also programmed in the switch:

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655
—dash indicating alternate channel means better legibility

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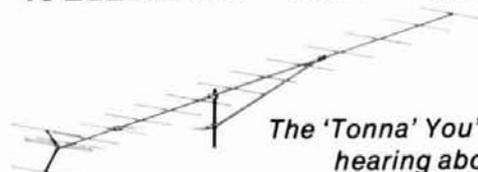
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Side lobe attenuation >60 dB	
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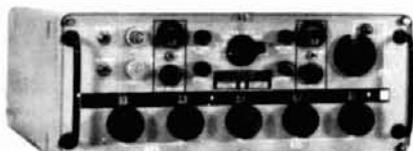
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924005-06R	40	6"	7.33 ea.
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923865-R	40	straight	1.94 ea.
923875-R	40	right angle	2.30 ea.
923866-R	50	straight	2.36 ea.
923876-R	50	right angle	2.82 ea.

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Mates with standard IC sockets. 24" length - Fully Assembled & Tested

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16 sgl end	924106-24	3.02
16 dpl end	924112-24	2.13
16 sgl end	924116-24	3.30
34 dpl end	924122-24	3.34
34 sgl end	924126-24	2.20
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Also Available in 12" and 36" lengths

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CY2 B	2.010MHz	HC33	1.95
CY2 C	2.500MHz	HC33	4.95
CY3 27	3.2768MHz	HC33	4.95
CY3 57	3.579545MHz	HC33	4.95
CY3A	4.000MHz	HC18	4.95
CY4 91	4.915MHz	HC18	4.95
CY7A	5.000MHz	HC18	4.95
CY5 18	5.185MHz	HC18	4.95
CY6 14	6.144MHz	HC18	4.95
CY6 40	6.400MHz	HC18	4.95
CY6 55	6.5536MHz	HC18	4.95
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10MM size trimmers - .394" Dia.

Part No.	1-9	10-24	25-49	100+
TR-11 (valve)	30	25	20	20

Resistance values - 100, 500, 1k, 2k, 5k, 10k, 20k, 50k, 100k, 200k, 1 meg

TRIMPOTS

Single-Turn - 1/2 Watt
Square - Top Adjust - 3/8" Size

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63P (valve)	99	89	80	70

Resistance Values - 50, 100, 500, 1k, 2k, 5k, 10k, 20k, 50k, 100k, 200k, 500k, 1 meg

15-Turn - 3/4 Watt

Rectangular Side Adjust 3/4" x 1/4" Size

Part No.	1-9	10-24	25-49	50-99
43P (valve)	1.35	1.25	1.20	1.15

Resistance Values - 50, 100, 500, 1k, 2k, 5k, 10k, 20k, 50k, 100k, 200k, 500k, 1 meg

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Part No.	0-1	1-9	10-19	Price
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GLASS	64P44 06CWE	4.50	8.50	2.56 2.31
	169P44 06CWE	4.50	17.00	5.04 4.53
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Model P180 includes 2-100' spools #28 AWG wire wrap wire

Supplies insulated wire from spool to wrap posts without pressing and precutting using "clay chain" method

Model P180 \$24.50

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For P180 Silt N Wrap

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Replacement wire wrap for P180 #28 AWG (pkgs of 3) \$2.75 each

W28 2 A green	W28 2 B red	W28 2 C clear	W28 2 D blue
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Injection molded unit. Complete with red bezel. 4 1/2" x 4" x 1-9/16"

\$3.49

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8251	Prog Comm. Interface	9.95
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1101	256 x 4 Static	5.95
1102	1024 x 1 Static	1.75
1107/5280	4096 x 1 Dynamic	4.95
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1112	256 x 4 Static	5.95
2114	4k x 1 Static 450ms	9.95
2114L	4k x 1 Static 450ms Low Power	10.95
2114L 3	4k x 1 Static 300ms Low Power	10.95
7489	16 x 4 Static	1.75
8101	256 x 4 Static	5.95
8101	Character Generator	6.95
8109	16 x 4 Static	3.49
21102	1024 x 1 Static	1.85
74200	256 x 1 Static	6.95
81021	256 x 1 Static	2.95
81022	2k x 1 Dynamic	\$17.00
MK4027 (UPD414)	4k DYNAMIC 16 PIN	5.95
MK4116 (UPD416)	16k DYNAMIC 16 PIN	39.95
1MS4044-45N	4k STATIC	14.95

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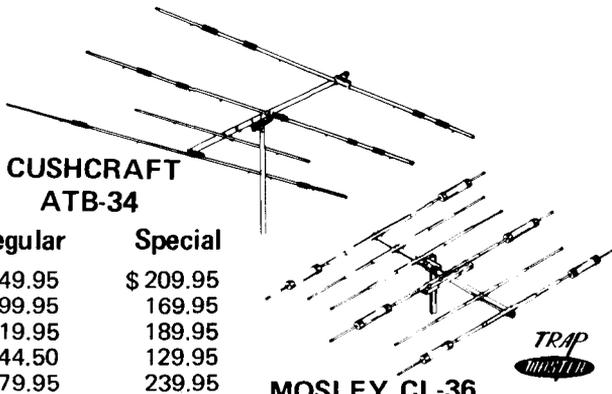
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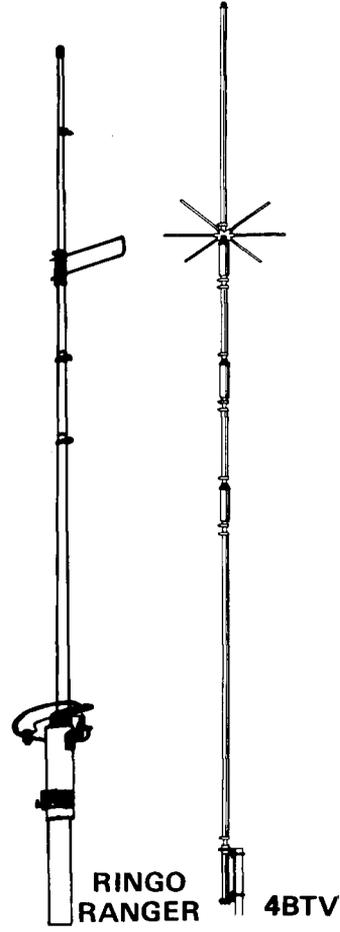
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 - Mounting bracket/handle, flex rubber antenna and 12 VDC power cord furnished.



GENERAL SPECIFICATIONS:

• **Frequency Range:** 144,000 - 147,995 MHz • **No. of Channels:** 799 @ 5 KHz or 399 @ 10 KHz • **Operating Mode:** Direct frequency modulation • **Type of Communication:** Simplex or transmitter offset ± 600 kHz • **Operating Voltage:** 13.6 VDC negative ground (10 to 15 VDC range) • **Current Drain:** Transmit: 290 mA @ 1 watt output, 2 amps @ 12 watts output. Receive: 45 mA squelched, 250 mA at full AF rated output • **Antenna impedance:** 50 ohms nominal • **Size:** 8-1/4 x 6-3/4 x 1-7/8 inches (209.6 x 171.5 x 47.6 mm) • **Weight:** 1 lb. 15 oz. (4.13 Kg): (3 lb. 11 oz. (8.16 Kg) with batteries) • **Frequency Determination Method:** C-MOS phase locked loop • **Offset Option:** Two optional offset TX positions also available.

PERFORMANCE SPECIFICATIONS:

TRANSMITTER: • **RF Output:** Hi 12W, Lo 1W • **Frequency Stability:** .001% - 10°C ~ +60°C • **Local Oscillator:** Simplex, 21.4 MHz; Offset +600 kHz, 22 MHz; -600 kHz, 20.8 MHz (Options for two more offsets other than 600 kHz) • **Harmonics & Spurious:** More than 60 dB below carrier • **Deviation:** ± 5 kHz • **Audio Response:** +1; -3 dB of 6 dB/Octave. Pre-emphasis characteristics from 300 to 3000 Hz.

RECEIVER: • **Receiving System:** Double conversion Superheterodyne • **First Local Oscillator:** PLL output of (F-21.4 MHz) • **First IF:** 21.4 MHz (with 2 pole monolithic filter) • **Second IF:** 455 kHz (with a ceramic filter) • **Stability:** .001% - 10°C ~ +60°C • **Sensitivity:** .3 μ V for 20 dB quieting • **Squelch Sensitivity:** .2 μ V • **Spurious & Image Rejection:** Better than 80 dB • **Intermodulation:** 60 dB • **Selectivity:** ± 6 kHz at 3 dB, ± 15 kHz at 80 dB • **Channel Spacing:** 15 kHz • **Audio Output:** 2 W (10% distortion to 4 ohm)

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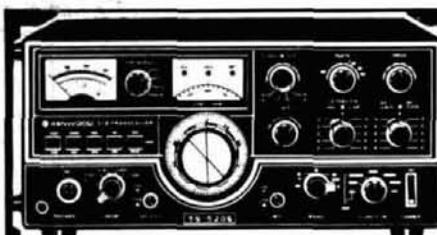
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KENWOOD TS-520S SSB transceiver

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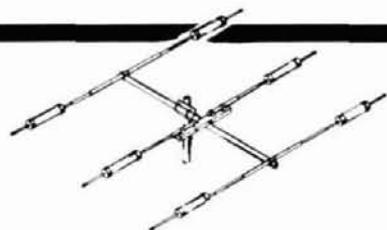
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EIMAC tubes win a place in Rockwell-Collins' HF-80 systems.

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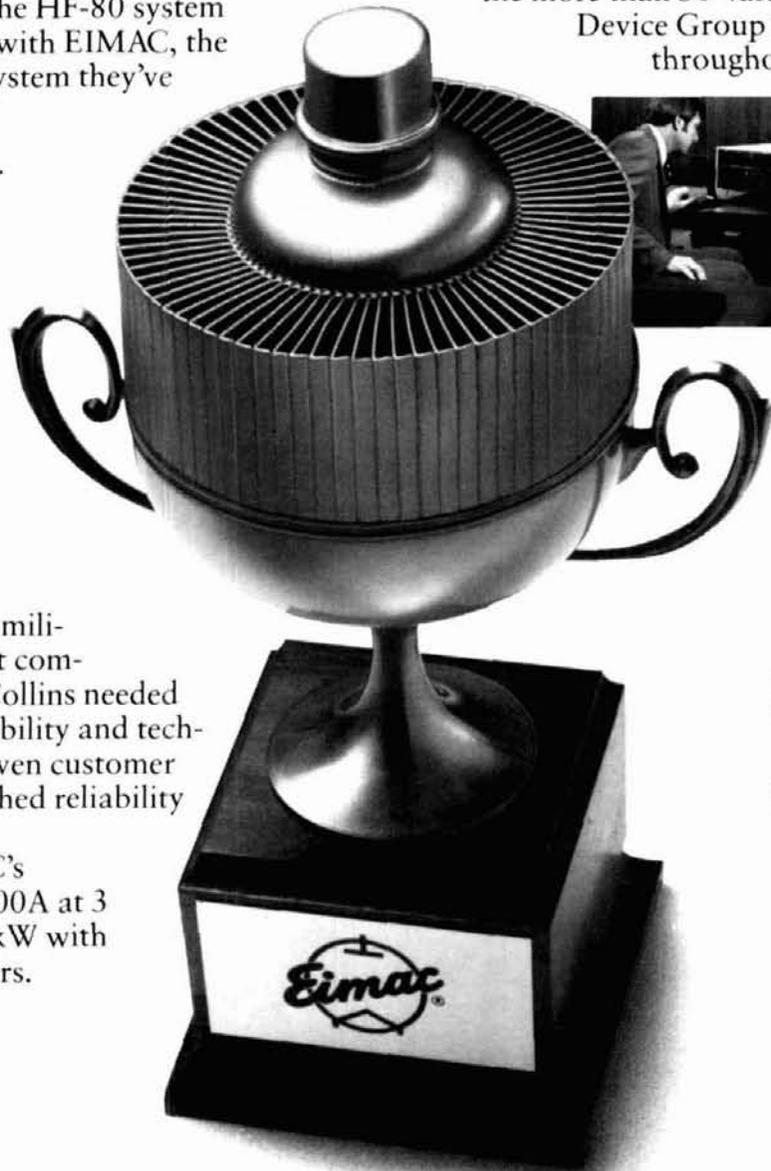
To power their new HF-80 family of 1 to 10 kW hf single sideband radio equipment, Rockwell-Collins needed tubes as well-constructed and reliable as the HF-80 system itself. That's why they went with EIMAC, the way they have for every hf system they've built since 1958.

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The HF-80 uses EIMAC's 4CX1500B at 1 kW, 4CX5000A at 3 kW, and 4CX15000A at 10 kW with EIMAC's 4CX350A as drivers.

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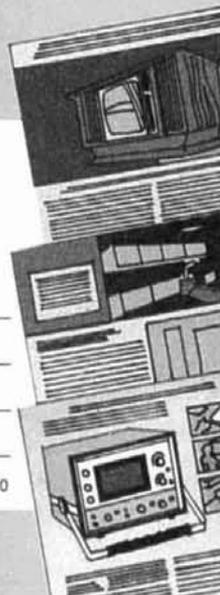
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