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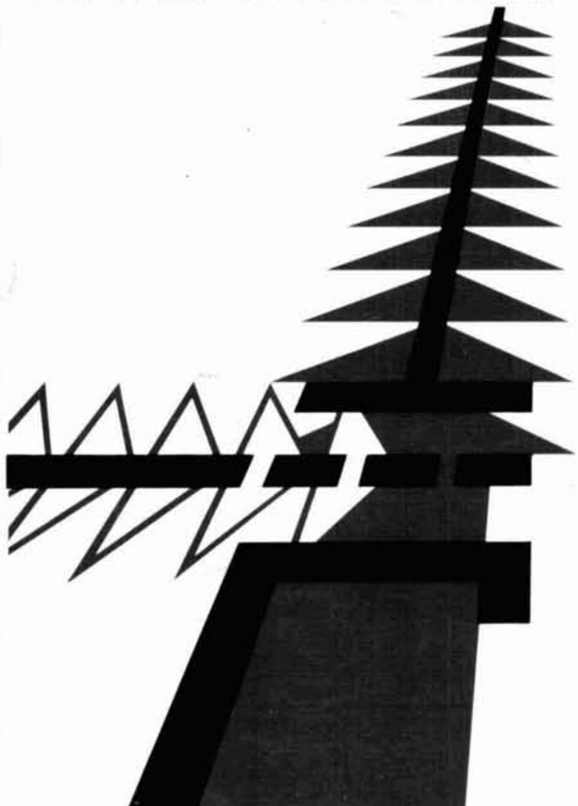
ham radio

magazine

JULY 1977

1296-MHz TRANSVERTER



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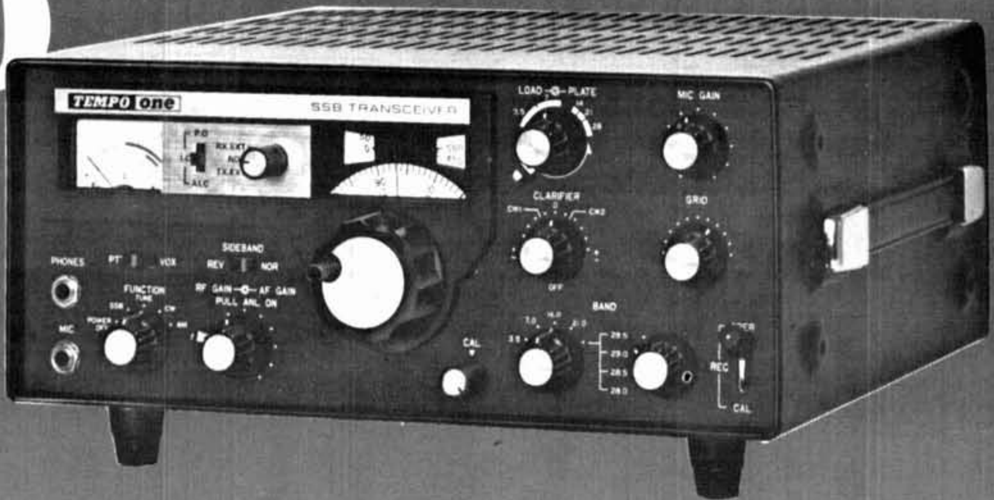

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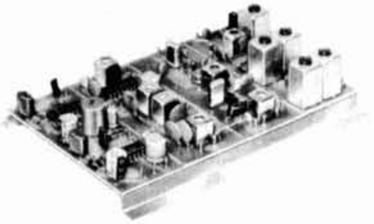

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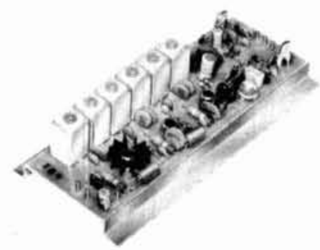

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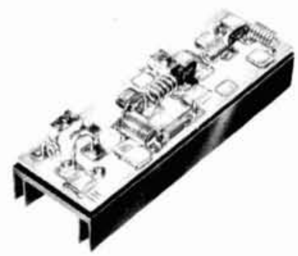

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To best take advantage of the DX as well as the extended range capabilities we recommend our optional tone squelch board TS-1 and TD-3 touch tone decoder.

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More Details? CHECK—OFF Page 126

ham radio magazine

JULY 1977

volume 10, number 7

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

subscription rates

U.S. and Canada: one year, $10.00
three years, $20.00
Worldwide: one year, $15.00
three years, $35.00

foreign subscription agents

Ham Radio Canada
Box 114, Goderich
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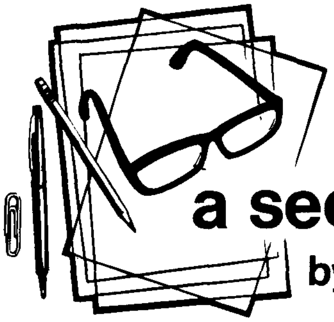
Post Office Box 63, Harrow
Middlesex HA3 6HS, England

Holland Radio, 143 Greenway
Greenside, Johannesburg
Republic of South Africa

Microfilm copies
are available from
University Microfilms, International
Ann Arbor, Michigan 48103

Second-class postage
paid at Greenville, N.H. 03048
and at additional mailing offices
Publication number 23340

contents



a second look

by Jim Fisk

The editor of a technically oriented magazine like *ham radio* has to wear several hats. I could use this whole page to describe the details that require attention to keep the magazine running smoothly, but I'd like to talk for a moment about one very important editorial task that means the difference between an interesting, accurate technical magazine, and one that isn't.

The articles in *ham radio* are written by authors not unlike you, the reader. They range from enthusiastic hams who want to share an idea, to amateurs with engineering backgrounds (who also want to share an idea). I welcome the output of anyone who is interested in contributing something which will benefit all hams.

Budding authors often ask, "What kind of articles are you looking for?" That question is difficult to answer since many new manuscripts arrive every day, but generally speaking, I am looking for simple construction projects that the average reader can put together in one or two weekends. Larger construction projects are also welcome, but the average *ham radio* reader must split his leisure time between amateur radio and other interests, so he doesn't have time to build a Chinese copy of a complex piece of gear.

When I read an article contributed to *ham radio*, the first thing I look for is interest value. If the manuscript passes this test, the next thing I look for is technical accuracy and attention to detail. The contributed article doesn't have to be a literary masterpiece. If you have a good idea; if it's well documented; if the illustrations and technical discussion are clear and accurate — you may have a winner!

Don't feel too badly if your article is not accepted. Since we publish about twelve feature articles in every issue, to keep the production pipeline full we purchase that same number of new manuscripts each month. This keeps the article backlog to a minimum and insures that the fare served up in each issue of *ham radio* represents the latest possible information on any given subject.

During an average month I receive 50 to 60 new manuscripts; at some point during the month I sit down and go through each of the articles rejecting those that are clearly not usable in *ham radio*; this narrows the stack of manuscripts down to 25 or so.

Now comes the most difficult part — deciding which of those remaining 25 articles are most desirable. Articles that are too long or need more polishing are returned to the author with suggestions for rewriting the article to our standards. The remaining material is further screened for technical accuracy and reader interest; this process continues, with comments from the staff (and usually arguments), until we have decided upon the articles which will be published in *ham radio*.

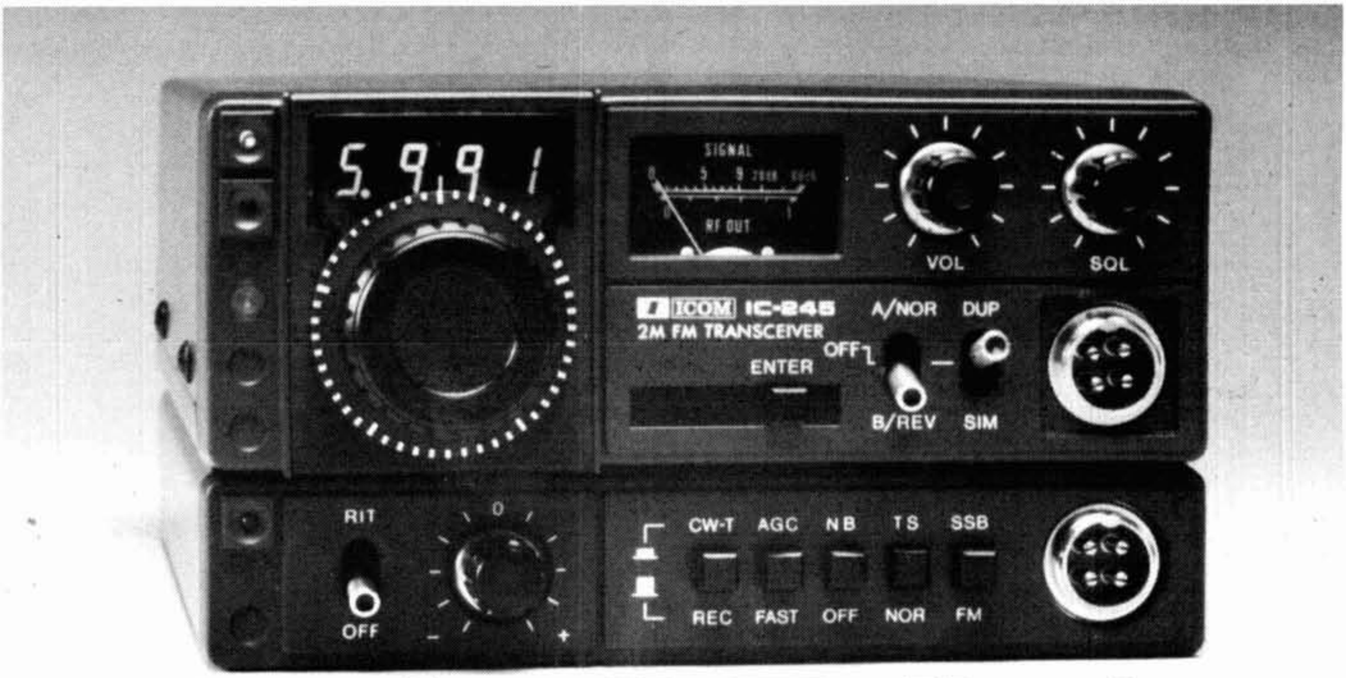
If you prefer to read *ham radio*, rather than write for it, you can help by telling me the kind of article you like. If you have a pet project in mind, or an old project that could be updated with transistors or ICs, let me know about it; I'll pass the idea along to one of our regular contributors.

1977 ham radio sweepstakes

The winner of the Grand Prize in the 1977 *ham radio sweepstakes*, a Kenwood TS-820 six-band transceiver, one of the most exciting of this year's new amateur transceivers, *and* the versatile Kenwood TR-7400A two-meter fm transceiver, is Clarence "Bick" Bickford, KØRHP. Other happy winners in this year's Sweepstakes are Bill Locke, W4RPU, and George Gruetzmacher, WB9QBA, who each won TS-820 transceivers; and Hal Tune, W8LZE, and Jack Langley, KØMER, who are proud new owners of TR-7400A fm rigs.

This year's sweepstakes was one of our biggest ever, and we want to thank all of you who took the time and effort to enter. Next year we'll be doing it again, so stay tuned . . .

Jim Fisk, W1HR
editor-in-chief



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All Mode Mobile is now yours in a superior ICOM radio that is a generation ahead of all others. The new, fully synthesized **IC-245/SSB** puts you into FM, SSB and CW operation with a very compact dash-mounted transceiver like none you've ever seen.

- **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 **IC-245/SSB** LSI chip provides for the input of programming digits from a remote key pad which can be combined with Touch Tone* circuitry to provide simultaneous remote program and tone. Computer control from a PIA interface is also possible.

* a registered trademark of AT&T.

- **FM stability on SSB and CW:** The **IC-245/SSB** synthesis of 100 Hz steps make mobile 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 discovered SSB.

The **IC-245/SSB** is the very best and most versatile mobile radio made: that's all. For more information and your own hands-on demonstration see your ICOM dealer. When you mount your **IC-245/SSB** you'll have all you need for All Mode Mobile.

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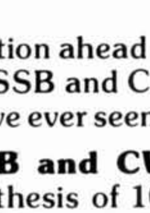

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More Details? CHECK — OFF Page 126

presstop

A NEW "13-METER" BAND, 15 meters back where it belongs, and 40 meters exclusively Amateur were the big pluses in the FCC's WARC Fifth Notice of Inquiry, released May 20th. On the negative side, the proposed 160-190 kHz band has been dropped entirely due to objections from the power-generating industry which uses those frequencies for carrier-current telemetry and control of their high-voltage transmission lines; they fear strong Amateur signals could cause interference problems.

Looking At The Latest proposals band-by-band and referencing them to last December's Third Notice of Inquiry (February HR) finds:

1600 Meters:	160-190 kHz dropped.
160 Meters:	1750-1800 kHz dropped; 1800-1900 kHz exclusive; 1900-2000 kHz restored, shared with existing services (Loran).
80 Meters:	3500-3900 kHz exclusive; 3900-4000 kHz shared (no change).
40 Meters:	6950-7250 kHz exclusive; 7250-7300 kHz dropped.
20 Meters:	13950-14400 kHz exclusive (no change).
15 Meters:	20700-20950 kHz dropped; 20950-21200 kHz exclusive; 21200-21450 kHz exclusive (restored).
13 Meters:	25.76-25.86 MHz new, exclusive.
10,6,2 and 1¼ Meters:	all remain as they are now and were proposed in December.
3/4 Meters:	420-450 MHz unchanged, except Amateur Satellite has been proposed worldwide for the 435-438-MHz slot.
35cm:	902-928 MHz shared unchanged.

In The 1215-MHz Band, Amateurs would lose 1215-1240 MHz with 1240-1290 MHz shared with radiolocation and Amateur satellite sharing 1290-1300 MHz. The only other change in the higher frequencies is a partial restoration of the previously dropped 48-50 GHz band to 49.8 to 50 GHz.

It Must Be Emphasized that these bands are only proposed and only for the U.S. position at the 1979 World Conference. They'll be reviewed by the Advisory Committee on Amateur Radio, tentatively scheduled to meet in Washington on July 12. Loss of the 160-190 kHz band will certainly be discussed, and perhaps another spot in that spectrum can be found for us. The new 25.76-MHz band, a useful addition, is not nearly so significant to Amateur needs as the 10- and 18-MHz bands we previously requested, so the ACAR should be focusing on that area, too.

Comments On This Fifth Notice of Inquiry are due August 1st, and Reply Comments August 22nd. We'll need lots of help if we're to preserve the relatively good position we're now in.

GETTYSBURG'S SPECIAL LICENSING CHIEF was indicted April 29th by a Harrisburg, Pennsylvania Federal Grand Jury on four counts of bribery. The indictment charged FCC official Richard C. Ziegler with the alleged solicitation of $100 each from four individuals in exchange for influencing the processing of their callsign applications last spring, according to the report in the April 29 Harrisburg *Patriot*. The indictment apparently was the result of an FBI investigation begun last fall.

220-MHZ CLASS-E CB still seems to be a live issue with the FCC, according to a report from the Commission's Office of Plans and Policy that was released in late April. The document, "Spectrum Alternatives of Personal Radio Service," was prepared by the Personal Radio Planning Group, which considered 17 alternative spots between 25 and 1215 MHz for personal radio expansion.

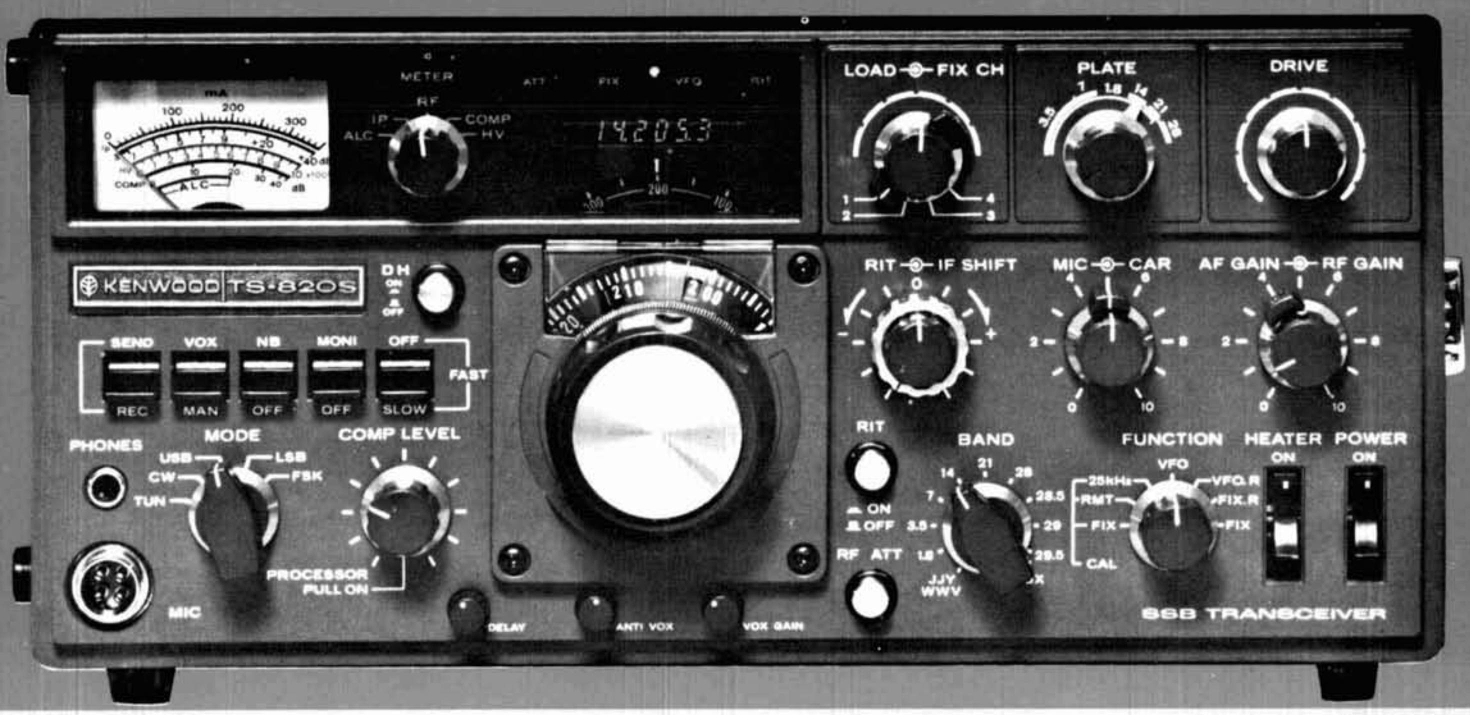
Based On Such Factors as user loading, costs of relocation, and TVI potential, seven segments were selected: 26.95-26.96, 27.54-28.00, 29.80-29.89, 29.91-30.00, 222-224, 894-902, and 928-947 MHz. Since many of the problems of the present Class-D band seem to rule out any of the slots in the 26-30-MHz region, we're back to that same old basic conflict between 220 and 900 MHz. However, despite the Planning Group's determination that its selections were the only likely current candidates, their report did not preclude consideration of some totally different spot, and even suggested that the Commission could decide that further CB expansion was not in the public interest.

EAST COAST VHF/UHF CONFERENCE in New Hampshire May 7th and 8th was rated a "huge success" by the nearly 100 attending from the U.S. and Canada. Noise figure competition winners were K2UYH with his V-244 based 432-MHz preamp at 0.95 dB, and W1JR's 144-MHz design with 0.8 dB.

Ham Radio's Annual Award for Technical Achievement went to the Mt. Airy VHF Club for the Pack Rats' 432-MHz EME DXpedition to Colombia last summer — W3HMU accepted the plaque for the club.

ARMY RESERVIST AMATEURS can get their two-week summer duty teaching teenagers Amateur Radio in the FAA's Career Interest Program. For details call Col. Colby or Lt. D'Angelo at (202)325-8483, or write 118 South Royal, Alexandria, Virginia 22314.

TS-820S



We told you that the TS-820 would be the best. In little more than a year our promise has become a fact. Now, in response to hundreds of requests from amateurs, Kenwood offers the TS-820S*... the same superb transceiver, but with the digital readout factory installed. The worldwide demand for the TS-820 far exceeded our initial production plans. However, production capacity has been substantially increased and our objective is to make the TS-820S more readily available to you. As an owner of this beautiful rig, you will have at your fingertips the combination of controls and features that even under the toughest operating conditions make the *TS-820S the Pacesetter that it is.*

Features

Following are a few of the TS-820S* many exciting features.

SPEECH PROCESSOR • An RF circuit provides quick time constant compression using a true RF compressor as opposed to an AF clipper. Amount of compression is adjustable to the desired level by a convenient front panel control.

IF SHIFT • The IF SHIFT control varies the IF passband without changing the receive frequency. Enables the operator to eliminate unwanted signals by moving them out of the passband of the receiver. This feature alone makes the TS-820S a pacesetter.

TV-506 TV-502 VFO-820 TS-820S SP-520

PLL • The TS-820S employs the latest phase lock loop circuitry. The single conversion receiver section performance offers superb protection against unwanted cross-modulation. And now, PLL allows the frequency to remain the same when switching sidebands (USB, LSB, CW) and eliminates having to recalibrate each time.

DIGITAL READOUT • The digital counter display is employed as an integral part of the VFO readout system. Counter mixes the carrier, VFO, and first heterodyne frequencies to give *exact* frequency. Figures the frequency down to 10 Hz and digital display reads out to 100 Hz. Both receive and transmit frequencies are displayed in easy to read, Kenwood Blue digits.

Specifications

FREQUENCY RANGE: 1.8-29.7 MHz (160 - 10 meters)
MODES: USB, LSB, CW, FSK
INPUT POWER: 200W PEP on SSB
160 W DC on CW
100 W DC on FSK
ANTENNA IMPEDANCE: 50-75 ohms, unbalanced
CARRIER SUPPRESSION: Better than -40 dB
SIDEBAND SUPPRESSION: Better than -50 dB
SPURIOUS RADIATION: Greater than -60 dB (Harmonics more than -40 dB)
RECEIVER SENSITIVITY: Better than 0.25uV
RECEIVER SELECTIVITY:
SSB 2.4 kHz (-6 dB)
4.4 kHz (-60 dB)
CW* 0.5 kHz (-6 dB)
1.8 kHz (-60 dB)
*(with optional CW filter installed)
IMAGE RATIO: 160-15 meters: Better than 60 dB 10 meters: Better than 50 dB
IF REJECTION: Better than 80 dB
POWER REQUIREMENTS: 120/220 VAC, 50/60 Hz, 13.8 VDC (with optional DS-1A DC-DC converter)
POWER CONSUMPTION: Transmit: 280 Watts Receive: 26 Watts (heaters off)
DIMENSIONS: 13-1/8" W x 6" H x 13-3/16" D
WEIGHT: 35.2 lbs (16 kg)

VFO-820
Function switch provides any combination of transmit/receive/transceive with the TS-820S. Both are equipped with VFO indicators showing which VFO is in use.

SP-520
Although the TS-820S has a built-in speaker, the addition of the SP-520 provides improved tonal quality. A perfect match in both design and performance.

TV-502
The TV-502 transverter puts you on 2-meters the easy way. Operates in the 144.0-145.7 MHz frequency range with a 145.0-146.0 MHz option. Completely compatible with the TS-820S, the TS-520S and most any HF transceiver.

TV-506
Similar to the TV-502 except that it opens up the 6-meter band (50.0-54.0 MHz) to your HF rig.

*The TS-820S and DG-1 are still available separately.

1296-MHz transverter

Complete construction details for a simple, inexpensive transverter for ssb and CW that will make a noticeable dent on the 1296-MHz amateur band

Recent issues of *ham radio* and other amateur publications have contained a wealth of construction articles on equipment for 1296 MHz, indicating the growth of interest and activity on that band. Conspicuously absent from the literature, however, is a simple, inexpensive way of generating reasonable amounts of stable transmitter output power — greater than 1 watt — for serious CW and ssb work on 1296 MHz.

Most long-haul, narrow-band DX work is conducted at or just above 1296 MHz, leaving the lower part of the band for wideband modes. Traditional transmitting schemes for this band usually involve tripling from 432 MHz using planar triodes in a cavity or stripline arrangements, and more recently, varactor diodes. These approaches yield CW or fm signals, but they are obviously unsuitable for single sideband.

Recent solid-state mixer designs, although suitable for producing clean ssb and CW signals have one principal drawback: low power output, typically in the dozens of milliwatts range. This requires considerable linear amplification to approach reasonable power levels. The circuitry associated with uhf mixers is difficult for amateurs unfamiliar with these devices and expensive in terms of dollars invested per watt of outpuut power obtained.

In an effort to overcome these drawbacks with minimum circuit complexity and financial investment, a high-level mixer was developed which uses the popular 2C39/7289/3CX100A5 family of planar triodes. These tubes are abundantly available surplus at extremely reasonable prices. Depending on the plate voltage applied to the tube, this transverter will deliver from 5 to 15 watts of clean, stable CW and ssb power output on 1296 MHz. This is more than enough for routine contacts up to a 100 miles (160km) or more. My 1296-MHz signals are regularly copied at +20 dB over S-9 over a 50 mile (80km) path using the transverter stage alone; for more serious DX work the unit will drive a single 7289 to 100 watts ssb and CW output!

theory of operation

The 1296-MHz transverter operates like a receiving converter or mixer in reverse, and at much higher power levels. As shown in **fig. 1**, an ssb signal from the output of a high-frequency or vhf transmitter (here considered to be the *intermediate frequency* or i-f) is mixed with a higher frequency carrier (the local oscillator or LO) to produce sum and difference frequencies, of which one is the desired uhf ssb signal. The remaining, undesired signals are eliminated with a selective filter.

I used the output of a 50-MHz ssb transceiver for my i-f. Obviously, other ssb source frequencies could be used, but it is desirable to use as high an i-f as possible to separate the desired mixer product from the unwanted LO and difference frequencies as much as possible, making it easier to eliminate the unwanted signals by filtering. Intermediate frequencies as low as 21 or 28 MHz can be used with little difficulty.

Transceivers in the 10-20 watt class are ideal for driving this transverter. They should, however, be

By Joe M. Cadwallader, K6ZMW, Star Route 2, Bosse Road, Jackson, California 95642

isolated from the transverter by a simple 3 dB attenuator[1] (such as a suitable length of RG-58/U coaxial cable) to make sure the transceiver is terminated in a matched, resistive load. The output of transceivers in the 100-watt class should be attenuated down to about 10 watts; don't just turn down the DRIVE or MIC GAIN control.

The transverter requires about 5 watts of local oscillator injection at 1296 MHz plus or minus the i-f. This signal can be derived in a number of ways. In my case, withh a 50-MHz i-f, a LO of either 1246 or 1346 MHz was needed. A crystal-controlled signal source providing about 10 watts output at 415.333 MHz was built; this was used to drive a tripler stage to 1246 MHz with about 5 watts output.

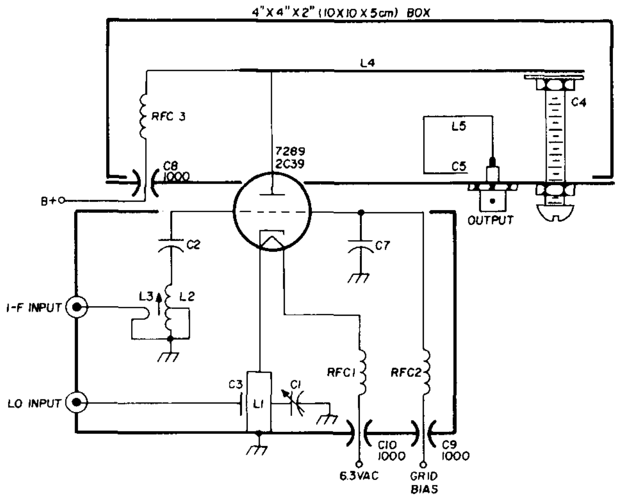
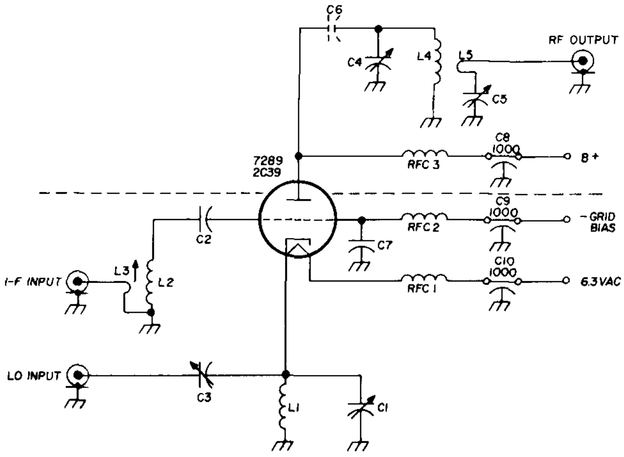


fig. 1. Circuit layout for the 1296-MHz transverter. Component details are listed under fig. 2.

There are a number of ways to generate the 415-MHz signal: the easiest is to modify and retune an existing transmitter which operates near this frequency. Many 432-MHz transmitters described in amateur publications can be easily retuned; transistorized transmitter kits advertised in amateur publications are very reasonably priced and should work well for this purpose.

Even old commercial 450-MHz fm transmitter strips, often available as junk, work nicely. If this approach is used, however, a few precautions are in order: turn the DEVIATION control off and, if possible, remove the speech-amplifier and phase-modulator tubes; substantially increase the power supply filtering to assure clean output with no ac hum which would otherwise appear on your transverter LO signal; voltage-regulate the oscillator and buffer stages with zener diodes or VR tubes to maintain oscillator stability; and use a good quality crystal with a *low temperature coefficient in a temperature-*controlled oven, or mount the crystal under the



C1, C3	part of cathode circuit (see fig. 5)
C2	150 pF silver-mica capacitor for 50-MHz i-f (three 47-pF dipped silver-mica capacitors in parallel)
C4	plate tuning capacitor (see fig. 7)
C5	part of L5 (see fig. 3) or 10 pF piston trimmers
C6	non-existent; represents dc open condition of this line configuration (see text)
C7	grid bypass capacitor (see fig. 4)
C8,C9,C10	1000 pF feedthrough capacitors. C8 must be rated for applied B + voltage
L1	part of cathode circuit (see fig. 5)
L2	4 turns no. 16 (1.3mm) enamelled copper wire on 1/4" (6.5mm) slug-tuned coil form (for 50-MHz i-f)
L3	1 turn no. 16 (1.3mm) around cold end of L2
L4	plate line (see fig. 6)
L5	1/4" (6.5mm) wide copper or brass strip, about 1/8" (3mm) away from plate line (see fig. 3)
RFC	15 turns no. 16 (1.3mm) copper wire, closewound on 1/16" (1.5mm) mandrel

fig. 2. Schematic diagram of the 1296-MHz transverter. The circuit layout is shown in fig. 1. Rf power output at 1296 MHz is 17 watts.

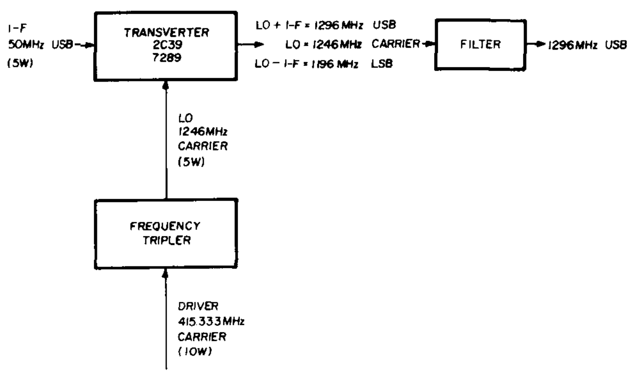
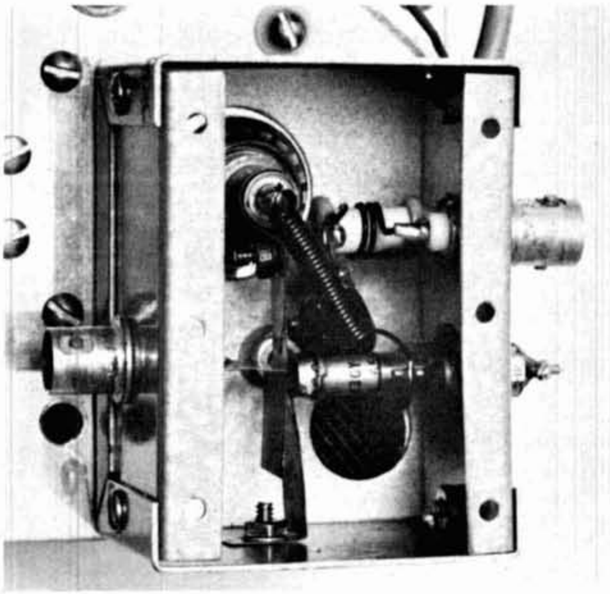


fig. 3. Block diagram of the 1296-MHz transverter system for CW and ssb operation. Although the author used a 50-MHz ssb/CW transmitter, 21 or 28 MHz could be used with equally good results. Frequencies below 21 MHz are not recommended because of the difficulty in separating the resulting mixer products.

chassis away from sources of heat, to reduce LO drift.

Regardless of which approach is used to generate the LO signal, a small amount can also be coupled off for LO injection to the receiving converter, thus reducing the total system equipment requirement and yielding true transceive operation on 1296 MHz. In my case, a small amount of 415.333-MHz energy was inductively coupled from the PA grid circuit of a 10-watt transmitter strip used as the LO source and applied to the multiplier diode of a popular trough-line receiving converter.[2]

The task of tripling up to the transverter's required LO injection frequency can be readily accomplished in a varactor multiplier — either commercial* or homebrew[3] — or a stripline or cavity multiplier stage[4,5] can be built around a 2C39/7289 triode. It has been suggested that cavity assemblies from surplus uhf equipment such as the UPX-6 would serve this purpose well. As is, these beautiful cavities tune from roughly 1000 to 1200 MHz.

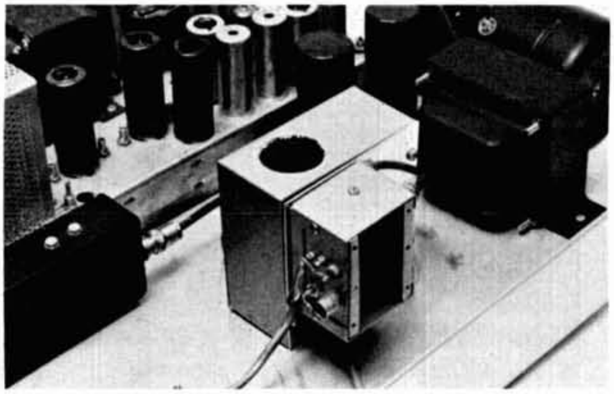


Close-up of the cathode compartment. The BNC connector on the left is used for the local oscillator input, while the one on the right is for the i-f input. The filament and grid voltages are brought into the compartment by feed-through capacitors.

Table 1 lists the required local-oscillator frequencies for various intermediate frequencies. Keep in mind that using an LO *above* 1296 MHz causes inversion of the sideband in the transverter: for example, a 1346-MHz LO minus a 50-MHz *upper* sideband signal equals 1296-MHz *lower* sideband.

Referring to the schematic diagram, **fig. 2**, the

*The MMv-1296 tripler available from Spectrum International, Box 1084, Concord, Massachusetts 01742.



The complete 1296-MHz ssb/CW transverter, mounted on a pressurized chassis. The local-oscillator chain is at the upper left, the varactor tripler is to the left, and the high-voltage power supply and blower are at the right.

transverter circuit is remarkably simple, using a 7289 (2C39) or equivalent in the familiar grounded-grid configuration. As is common practice in this application, the tube grid is not actually grounded directly but rather is bypassed, through capacitor C7 so the grid is grounded at the signal frequency while remaining above ground to dc. This provides a convenient way to apply grid bias through RFC2 and C9 without affecting the rf behavior of the grid circuit.

table 1. Possible i-f/local oscillator combinations for the 1296-MHz transverter. LO frequencies above 1296 MHz invert the sideband.

intermediate frequency	local oscillator	driver (LO ÷ 3)
21 MHz	1275 MHz	425.000 MHz
21 MHz	1317 MHz	439.000 MHz
28 MHz	1268 MHz	422.666 MHz
28 MHz	1324 MHz	441.333 MHz
50 MHz	1246 MHz	415.333 MHz
50 MHz	1346 MHz	448.666 MHz
144 MHz	1152 MHz	384.000 MHz
144 MHz	1440 MHz	480.000 MHz

The grid bypass capacitor, C7, consists of a flat, concentric brass or copper plate connected by finger stock to the tube's grid collar and insulated from the chassis with a thin mica or Teflon sheet. This type of bypass plate is standard equipment on military surplus vhf communications gear and can often be scavenged. The bypass plate can also be home built by soldering finger-stock material* around an appropriately sized hole centered on a flat brass or copper sheet (**fig. 4**). Thin mica for the dielectric material is available in most hardware or plumbing supply stores as replacement material for gas furnace pilot-light inspection holes. It can be easily cut with scissors or an *Exacto* knife.

The bypass plate is insulated from its mounting

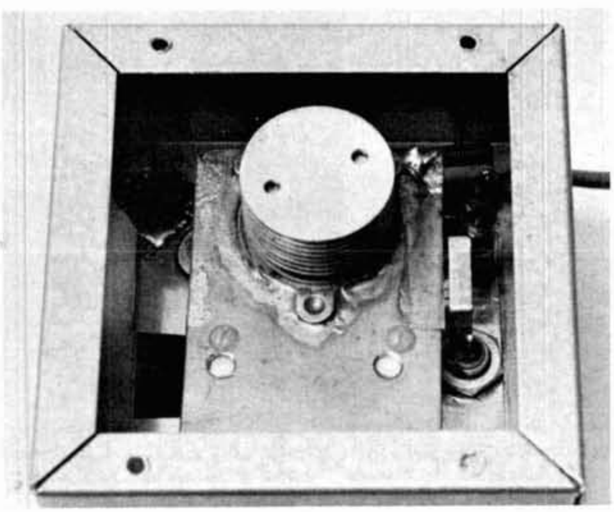
*Instrument Specialties Company, Little Falls, New Jersey 07424.

screws with the same type of nylon bushings which are used to mount and insulate power transistors. Once assembled, a typical value for this bypass capacitor is about 100 pF; this represents about 1 ohm of capacitive reactance — essentially a dead short — at 1296 MHz, and effectively grounds the grid at that frequency. However, at lower frequencies the grid is definitely *not* at rf ground: at 50 MHz the reactance of the grid bypass is about 30 ohms, and at 28 MHz it is about 55 ohms. Thus, the grid can be driven by the low-frequency ssb i-f signal while the cathode is driven by the high frequency LO signal: the sum and difference of these two signals appear in the plate circuit.

This simple approach can be applied to numerous uhf tubes in various grounded-grid configurations, stripline or cavity, commercial or military surplus, with equally good results.

The cathode circuit, **fig. 5**, driven at the LO frequency of 1246 MHz in my case, consists of a shorted quarter-wave line section L1, made of thin brass or copper sheet 1/4 inch (6.5mm) wide, wrapped around the tube cathode sleeve and running 1-3/4 inch (44mm) to chassis ground. The line is tuned with C1 which may be a low loss, high quality glass or ceramic piston trimmer or, better yet, a metal tab bent up near the line, or a brass machine screw with a brass disc about 1/2 inch (13mm) in diameter soldered to its end (similar to C4). The LO energy is capacitively coupled to the middle of this line by C3, a small brass or copper tab soldered to the LO input connector and bent up near the cathode line. Spacing can be adjusted for maximum LO drive.

The ssb i-f signal is coupled to the grid circuit by L3, a one-turn link wound around the cold end of L2. The L2-C2 circuit must resonate at the intermediate frequency. Since this circuit will be driven with about



The plate-line enclosure for the 1296-MHz transverter. The output coupling network, L5-C5, is to the right.

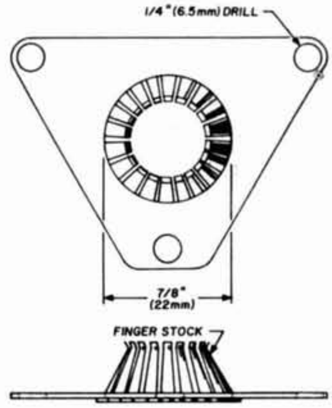
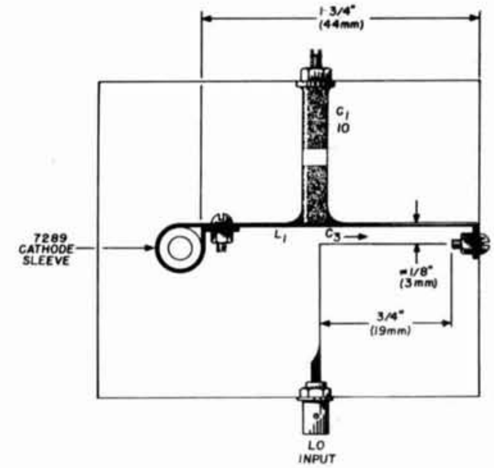


fig. 4. Construction of the grid bypass capacitor, C7.

5 watts, these coils should be either air-core or wound on a low-loss slug-tuned ceramic coil form of moderate diameter (1/4 inch [6.5mm] minimum, larger preferred) using no. 14 (1.6mm) to no. 18 (1mm) enameled copper wire. C2 should be a good quality mica or silver-mica capacitor to minimize losses. Two or three capacitors may be paralleled to handle the required rf current and still resonate with L2 at the i-f. The C2-L2 combination should be pre-

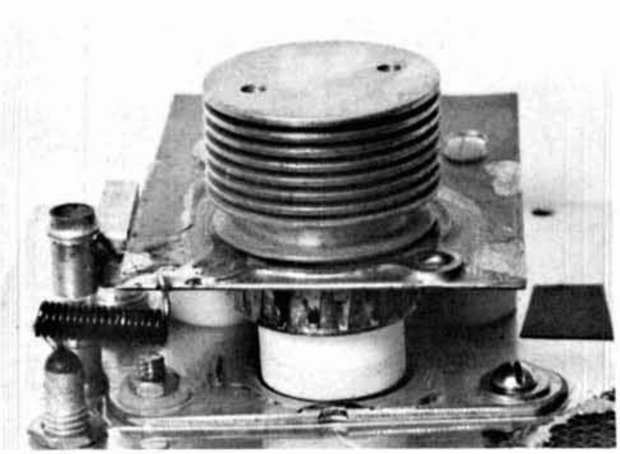


C2	10 pF piston trimmer capacitor or built as C4 (fig. 7)
C3	1/4" (6.5mm) wide copper or brass strap
L1	1/4" (6.5mm) wide copper or brass strap

fig. 5. Construction of the cathode circuit for the 1296-MHz transverter. This circuit is installed in a small aluminum minibox which is mounted on the plate enclosure (see fig. 1 and the photographs).

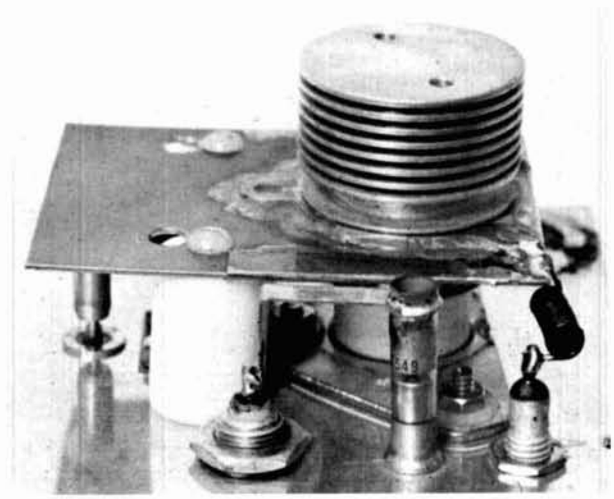
tuned to the intermediate frequency with a grid-dipper before installation.

The plate circuit (**fig. 1**) consists of an open half-wave line section, L4, tuned to 1296 MHz by capacitor C4. The line section is made of sheet copper or brass with finger stock connections to the tube anode ring and supported by insulating columns of Teflon, Rexolite, ceramic, or other good uhf dielec-



Construction details of the plate line, output network, and mounting of the grid-bypass capacitor, capacitor C7. Note that the mounting screw on the left is connected to the grid bypass plate for grid bias (from the cathode enclosure, below).

tric. The anode finger stock assembly can be found in the same military surplus vhf communication gear as the grid bypass assembly. Alternatively, commercial finger stock may be formed to the appropriate size and soldered around a hole in one end of the



Another view of the plate line showing the grid bypass capacitor, C7; output link, L5-C5; plate rf choke, RFC3; and bypass capacitor, C8.

plate line, L4, large enough to accommodate the tube anode ring as shown in **fig. 6**.

Capacitor C6 does not really exist, but merely represents the dc-open condition of this line configuration. In fact, the entire line and tube anode have B+ applied through RFC3 and C8. This feedthrough bypass capacitor must be rated to withstand the B+ voltage applied to the tube. If a commercial or surplus bulkhead capacitor cannot be found, this capacitor can be home-made using a 1-1/2-inch (38mm) square of sheet metal and sheet mica.

The plate tuning capacitor, C4, requires about 10 pF and must not break down at full applied B+. A low loss, high quality glass or ceramic piston trimmer (rated accordingly) would do nicely, or this capacitor may be constructed using 3/4-inch (19mm) wide, sheet brass strap bent up near the plate line for about 3/4 inch (19mm) and insulated from it with a layer or two of sheet mica (see **fig. 7**). The spacing between this tab and the plate line can be varied by any convenient mechanical means to tune the line. An alternate approach is to drill and tap the transverter top-plate to pass a no. 8 or no. 10 (M4-M5) brass machine screw with captive lock nut. Cut the screw so it is just short of touching the plate line (by at least 1/16 inch or 1.5mm) and solder a 3/4 inch (19mm) diameter brass disc to its end. Insulate the disc from the plate line with mica sheet.

Output power is inductively coupled from the plate line via L5 which is made from 1/4-inch (6.5mm) wide copper or brass strap soldered to the output connector (N or BNC type) and run parallel to the plate line about 1/8 inch (3mm) away. The end of this strap can then be run down to, and parallel with, the chassis to form the matching capacitor C5. A low-loss glass or ceramic piston trimmer of about 10 pF capable of handling some moderately large rf currents, could also be used.

construction

The grid i-f and cathode LO circuitry plus filament and grid bias wiring are all contained within (and shielded by) a small aluminum minibox mounted on top of the plate line enclosure. This minibox is secured by the same hardware which is used to mount the grid bypass plate, C7. Two screws are insulated with nylon bushings from this grid bypass plate; the third screw makes contact with C7 but is still insulated from the chassis by nylon bushings. Bias and i-f connections are then made to this screw.

The plate circuitry and output coupling loop are

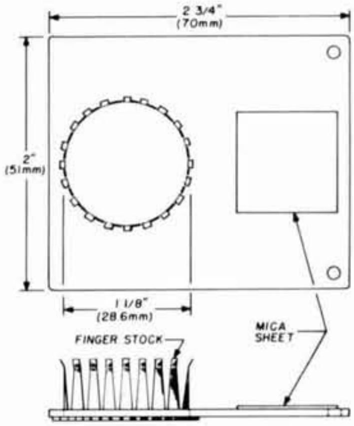
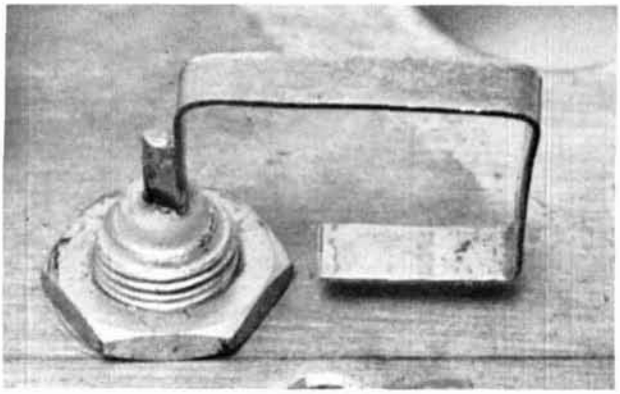


fig. 6. Plate line (L4 in figs. 1 and 2) for the 1296-MHz transverter. Finger stock can be obtained from Instrument Specialties, Little Falls, New Jersey.



An alternate arrangement for the output circuit, L5-C5.

contained within a 4 x 4 x 2 inch (10x10x5cm) minibox which serves as the chassis base. Screened air holes provide a path for forced air cooling of the tube anode structure, which is absolutely essential at reasonable power levels. These holes must be rf tight, so the shielding screen must be well grounded around its periphery.

Copper screen can be soldered to the aluminum

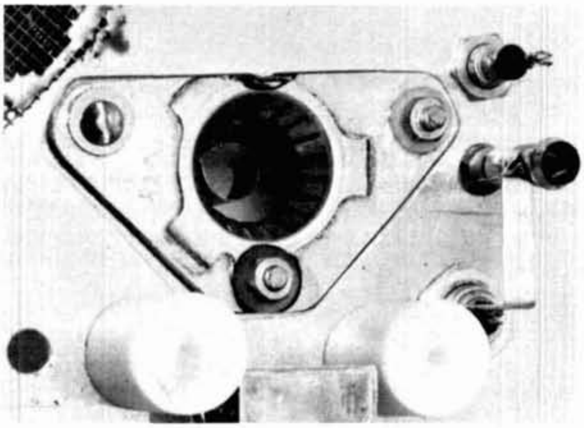


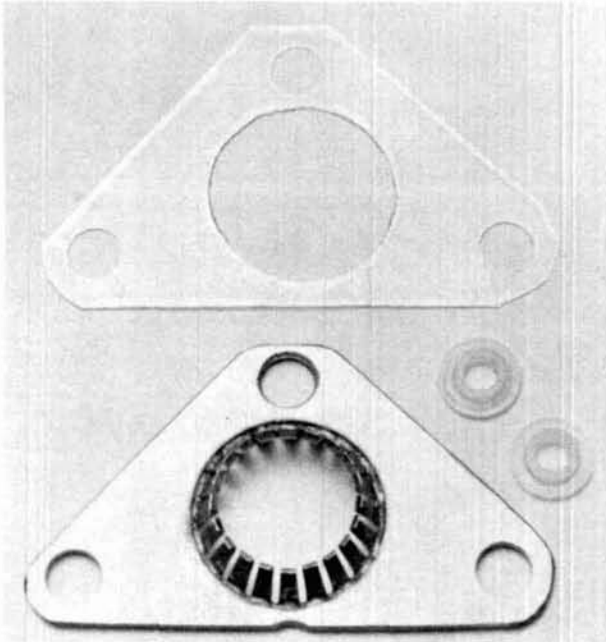
Plate line removed to show grid bypass plate. The plate line is mounted on the two Teflon pillars at bottom.

chassis by first tinning the aluminum with a large soldering gun or iron. Generously coat the periphery of the air hole with a heavy oil (clean auto engine oil works fine) to keep the aluminum from being exposed to air, then sand the surface clean using ordinary sandpaper or emery paper. Once cleaned, a hot iron (200 watts minimum) and rosin-core solder will tin the area beautifully. Then the copper screen can be soldered to the aluminum chassis. Practice this procedure first on some small aluminum scrap!

A convenient way to provide cooling air is to mount the transverter on edge on a large, air-tight pressurized chassis. The air hole in the transverter plate-line enclosure should be placed over an equal-sized hole in the pressurized chassis.

This chassis can also serve as a mounting platform and air source for the power supplies and LO chain components.

Common vhf construction practice should be followed throughout including short component leads, quality component selection, and rigid mechanical assembly. In one case, a 1/16-inch (1.5mm) thick, 4-inch (10cm) square brass plate was



Construction details of the grid bypass plate showing the sheet mica insulator cut out with an Exacto knife, and the insulating washers.

used as the top plate of the plate line enclosure with good results. It is a good idea to sand bare the adjoining surfaces of the plate line enclosure, top plate, and bottom plate, and use additional screws to assemble the top and bottom to assure good

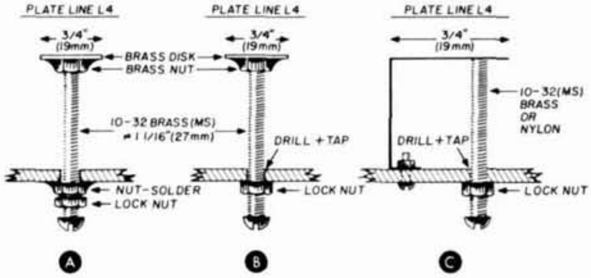


fig. 7. Three methods for building the plate tuning capacitor, C4.

shielding. When assembly is complete, a close visual inspection and a VOM continuity check should reveal any obvious problems before applying power.

tuneup and adjustment

Initial testing can best be done using variable-

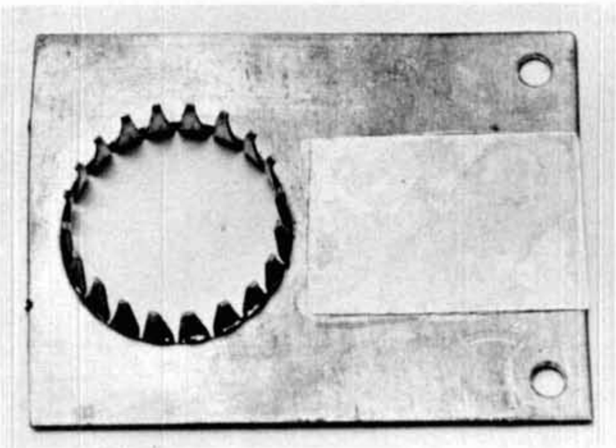


Plate line showing the plate collet (finger stock) and sheet mica insulator, right.

voltage power supplies to provide the plate voltage and grid bias. Since the grid will typically draw more than 50 mA, *the bias supply must have a low impedance and be capable of maintaining constant output voltage under fluctuating load conditions.* Apply filament voltage and cooling air, and after adequate filament warmup, gradually apply B+ while watching the plate current. Plate current should rise gradually, indicating normal tube conduction. Increase grid bias to reduce plate current until a plate voltage of about 300 volts can be applied with sufficient grid bias to limit plate current to about 20 mA.

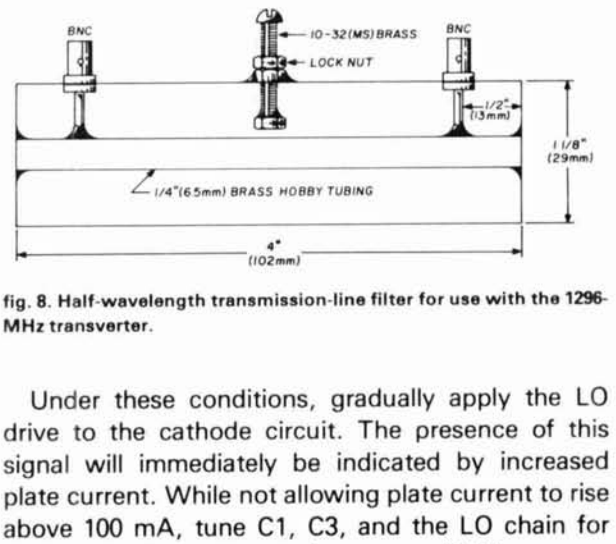


fig. 8. Half-wavelength transmission-line filter for use with the 1296-MHz transverter.

Under these conditions, gradually apply the LO drive to the cathode circuit. The presence of this signal will immediately be indicated by increased plate current. While not allowing plate current to rise above 100 mA, tune C1, C3, and the LO chain for maximum plate current. Increase grid bias as required to limit plate current to a safe value.

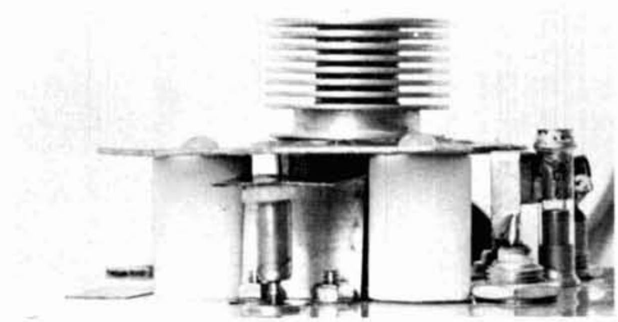
At this point, depending on your LO frequency, you *might* be able to resonate the plate circuit to the LO frequency by adjusting C4. Thus the unit operates as an amplifier to verify the behavior of the plate and output circuitry. If your LO is above 1296 MHz, tune toward minimum capacitance; if your LO is below 1296, tune toward maximum capacitance. Adjust C4, and the position of L5 and C5 for maximum indicated output power at the LO frequency.

Once cathode-line tuneup at the LO frequency has been accomplished and optimized, increase grid bias to reduce plate current to near cutoff — about 10 to 20 mA. Leave this grid bias at this value; the objective of this procedure is to bias the tube at or near cutoff (class AB or B) with plate voltage and LO drive applied but no i-f signal present. Now gradually apply a carrier at the i-f input and tune L2 for maximum plate current. Again, the presence of the i-f signal should immediately be indicated by increased plate current.

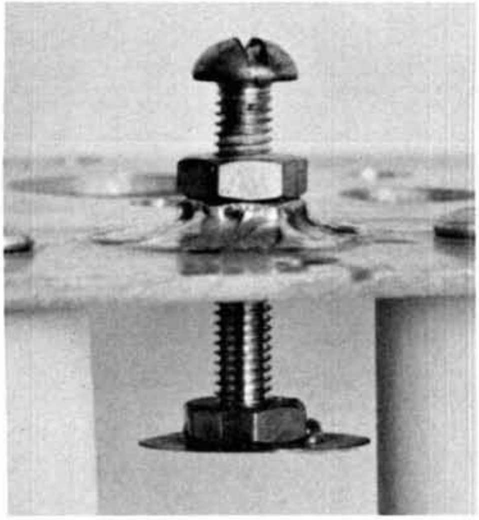
If all's well at this point, the LO and i-f signals are present in the tube's plate line; all that remains is to tune the plate line to the desired mixer product with C4. This process can be greatly simplified by placing a low loss filter tuned to 1296 MHz at the output of the transverter. An example filter of simple construction is shown in **fig. 8**.[6]

The filter can be pre-tuned to 1296 MHz by placing it in series with your 1296-MHz receiving system and tuning for maximum received signal from a nearby 1296-MHz transmitter, the third-harmonic of a 432-MHz transmitter, the 9th harmonic from a 144-MHz transmitter, or any other convenient rf signal source. This assures that any transverter output observed with the filter present will be on the desired signal frequency, and that power measurements will represent true power at the desired signal frequency, not the sum of the power contained in all the transverter's mixer products!

It is wise to leave the bandpass filter in the system to assure clean output. If, as is usually the case, your LO is *lower* than the desired mixer product on 1296 MHz, simply back out C4, (tuning toward minimum capacitance or higher frequency) while watching the output of the transverter for a peak. Once this peak is found, tune all screws for maximum output power. It may require a fair amount of time and patience to get the feel for the effects of each adjustment, and may even require repeating the entire procedure with a



The plate tuning capacitor, C4, is mounted under the plate line between the two Teflon pillars.



Construction of the home-made plate tuning capacitor, C4.

number of available tubes — some fly and some don't!

Once tuneup under reduced power conditions has been accomplished, full B+ may be applied; values from 500 to 1000 volts have been used successfully, but with lots of forced air cooling! The tube must then be re-biased for an idling plate current of 10 to 20 mA *with* LO but *no* i-f signal applied. Then increase CW i-f drive power up to the point of saturation (transverter output no longer increases with increasing amounts of i-f drive) and tune all adjustments for maximum output power. In the ssb mode, talk the transverter output up to an average of about half the maximum CW power output. Typical stage operating parameters are shown in **table 2**. The low efficiency is typical for this type of mixer circuit.

The half-wave, bandpass filter can also be used to transform a 50-ohm transmitter output impedance to 75 ohms to match the impedance of inexpensive,

table 2. Typical operating parameters for the 1296-MHz transverter. Output power was measured at the output of the bandpass filter with a calibrated Bird wattmeter and 50E slug.

Local-oscillator power	3 watts carrier
I-f power	4 watts CW
Plate voltage	+800 Vdc
Plate current	150 mA
Rf power input	120 watts
Grid bias	−35 Vdc
Grid current	50 mA
Rf power output	17 watts
Efficiency	14%

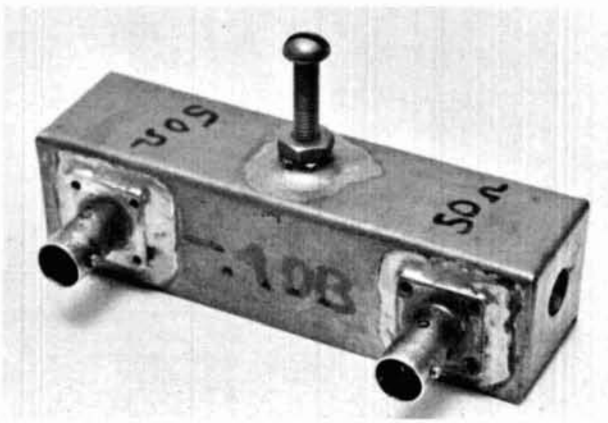
low-loss CATV coax. For 75 ohms simply space the tap 5/8 inch (16mm) from the line end.

conclusion

It is the intent of this article to describe a relatively simple and inexpensive method for the average amateur to get on 1296 MHz with enough power to make himself heard. Hopefully the ease with which this system can be put together will encourage more activity on the band and finally put to rest the myth that "you can't get there from here on 1296!" I am waiting now to hear from some hard-working and dedicated uhf buffs in Hawaii who would be interested in destroying the current 1296-MHz terrestrial record once and for all! How about it out there, any takers?

acknowledgements

Sincerest thanks are in order to the many people who contributed to this effort: notably to Bill Jungwirth, WA6NRV, for the beautiful construction of the first working prototype model of the transverter; to Tom Staller, WB6QHF, for the contribution of a commercial varactor tripler; and to the



Half-wavelength transmission-line filter for 1296 MHz. Loss of the filter shown here is 0.4 dB. Dimensions are shown in fig. 8.

West Coast's "Father of 1296," Bill Troetschel, K6UQH, who provided much good advice and encouragement amongst many good ribbings! Photo credits go to Alan Monie. Special thanks go to my wife, Jean, for typing the manuscript.

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ham radio

improved CW transceiver for 40 and 80

Update of a circuit previously presented in *ham radio* — a complete break-in and electronic-keyer circuit have been added to make a truly effective station for traffic and contest work

On Field Day 1968 I introduced a homebrew 10 - 75 watt CW transceiver for 80 and 40 meters. It performed so well during the contest that Rich Klinman, W3RJ, immediately adopted it, improved it, and used it as a home station. Since then it has been in almost constant use and has survived many a rigorous adventure, finally winding up as my home station.

An article describing the transceiver[1], noted that the lack of break-in operation was one of its more annoying shortcomings. That deficiency has now been corrected, and an ultra simple, built-in electronic keyer has been added to make the unit a truly complete CW station.

This article describes the break-in system and keyer, both of which have some novel features. The objective is to stress the applicability of each to other homebrew or commercial equipment. It's not necessary to duplicate the original rig to use the ideas presented here. A casual inspection of the original article reveals numerous areas in which present technology could be used to simplify the hardware, but we're concerned here only with the functional improvements that have been made.

Many forms of T/R switching are passed off as "break-in;" perhaps the most common being those that switch the transmitter on when the key is depressed and transfer control to the receiver after a delay of an appreciable fraction of a second. These systems drop out upon a deliberate pause but not during normal sending, whereas a really useful system permits interruption by the receiving operator at any time. The circuit described here is of the latter type, in which full recovery occurs even between dots at normal sending speeds.

To describe the break-in system, it's necessary to review a few details of the rig, which is a simple affair using a semiconductor, direct-conversion receiver with phase-shift sideband cancellation for single-signal reception. A 160-meter vfo with plug-in frequency multipliers for 80 and 40 meters provides excitation for both the receiver mixers and the vacuum-tube transmitter. The transmitter uses a 12AU6 driver and a 1625 final-amplifier tube. Both transmitter stages are cathode keyed through a transistor circuit that also provides sidetone and receiver audio blanking. Adjustable offset between transmit and receive frequencies is provided.

In the original design, the T/R function was performed by a dpdt toggle switch. One pole of the switch transferred the antenna from receiver to transmitter; the other pole supplied dc voltage to the keying circuits and also selected one of two reed relays, which controlled the receive/transmit frequency offset. The receiving reed relay connected a front-panel incremental-tuning capacitor across the vfo tank, while the transmit relay substituted a fixed capacitance equivalent to the midrange value of the incremental control.

All functions performed by the toggle switch must be duplicated at high speed by the key to achieve full break-in. Furthermore, a certain sequence of events

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is necessary. The required sequence for a single dot or dash is:

1. Blank receiver audio.
2. Shift vfo frequency.
3. Transfer antenna to transmitter.
4. Turn on transmitter and sidetone.
5. Turn off transmitter and sidetone.
6. Transfer antenna to receiver.
7. Reshift vfo frequency.
8. Restore receiver audio.

It is of course absolutely essential that antenna transfer and frequency shift be completed before activating the transmitter to avoid receiver damage, keying-waveform distortion, or the introduuction of a chirp. For the same reasons, the transmitter output must be allowed to decay fully before the reverse sequence is initiated. The order of sequences **1, 2, 3** and **6, 7, 8** is somewhat less critical; it's only necessary that audio blanking be initiated early enough and maintained long enough to block thumps due to front-end transients or received-signal chirps due to the vfo shift. In practice, simultaneous occurrence of the events yields satisfactory results if the transitions are fast.

The most crucial aspect of break-in design lies in the antenna coupling method. Low-loss switching of the transmitter output without relays is difficult, so most break-in designs leave the transmitter permanently connected to the antenna and provide means for disconnecting or limiting the rf voltage to the receiver during transmit. A classical approach is that of a cathode follower with a large grid-leak resistance coupled through a small capacitor directly to the coax antenna feed line. Grid current flow in the follower develops a negative bias that limits rf voltage to the receiver.

A serious disadvantage of feed line coupled receiver pickoff is the phenomenon of suck-out. The transmitter tank, usually a pi network, presents a low impedance to the feed line, thereby attenuating received signals. Such attenuation may be tolerable in receivers with a large gain and noise-figure margin but not in simple designs with limited sensitivity.

One solution to the suck-out problem involves coupling the receiver to the plate side of the final tank, so that conditions that provide proper transmitter loading also maximize the received signal. The cathode follower may be so coupled if a capacitive divider is used to reduce the rf voltage applied to the grid to manageable proportions. An additional advantage of the method is that the transmitter tank provides additional selectivity in the receiver front end, thereby improving rejection of out-of-band signals, which might produce intermodulation effects.

For the CW transceiver, it was desired to use an all-semiconductor T/R circuit in the interest of conserving filament power. It was also essential that no loss of gain be incurred and that the circuit introduce no intermodulation problems. These objectives were met using the hot-side coupling approach in conjunction with a shunt-transistor clamping circuit to protect the emitter-follower receiver pickoff.

T/R switch. The T/R circuit is shown in **fig. 1.** A 2N709 emitter follower, biased by network R1, R2, drives the low-impedance coaxial-line input to the receiver. Rf from the final tank is coupled to the 2N709 base through a capacitor constructed from a 4-inch (102mm) length of RG-58/U coaxial cable (capacitance approximately 10 pF). During receive, the 2N2222 shunt clamping transistor is inactivated by grounding the control line.

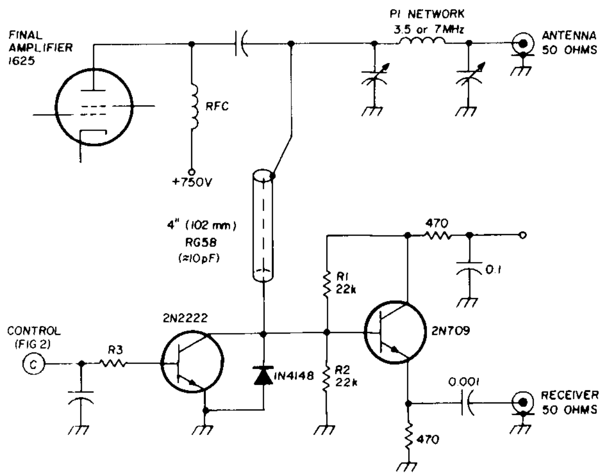


fig. 1. Solid-state T/R switch. In the original circuit (reference 1) transmit-receive toggle switch. In the improved version shown here true break in is obtained using a 2N2222 and 2N709.

For transmitting, +12 volts is applied to the control line. The resultant base current through R3 turns on the 2N2222, decreasing the 2N709 base voltage to near ground and cutting off receiver coupling. When the transmitter tank rf voltage goes positive, the 2N2222 conducts more heavily while remaining saturated so that insufficient rf voltage appears at the 2N709 base, which would cause energy transfer to the receiver. On negative rf swings, the 2N2222 conducts to some degree, but the transistor gain under these conditions is poor and assistance is provided by diode CR1, which clamps the negative excursion at about one volt.

The rf current that must be handled by the 2N2222 and the clamp diode is a function of tank rf voltage, frequency, and the value of coupling capacitance. With 750 Vdc on the 1625 plate, the rf voltage approximates 750 V peak, and the resultant peak current through 10 pF is 165 mA at 3.5 MHz and 330 mA at 7 MHz. The 2N2222 base current of about 22 mA is sufficient to maintain saturation with collector currents of these magnitudes.

An interesting aspect of the circuit is that the shunt transistor and clamp diode need not have spectacular rf response; in fact, a long storage time in the transistor aids to maintain saturation and reduces base drive requirements. During receive, both transistor and diode are cut off and are totally out of the act; therefore they contribute no loading or distortion on high-amplitude signals. The 2N709 follower is degenerative and has good linearity.

It is essential that the control voltage *always* be applied during transmit; a few microseconds at 750

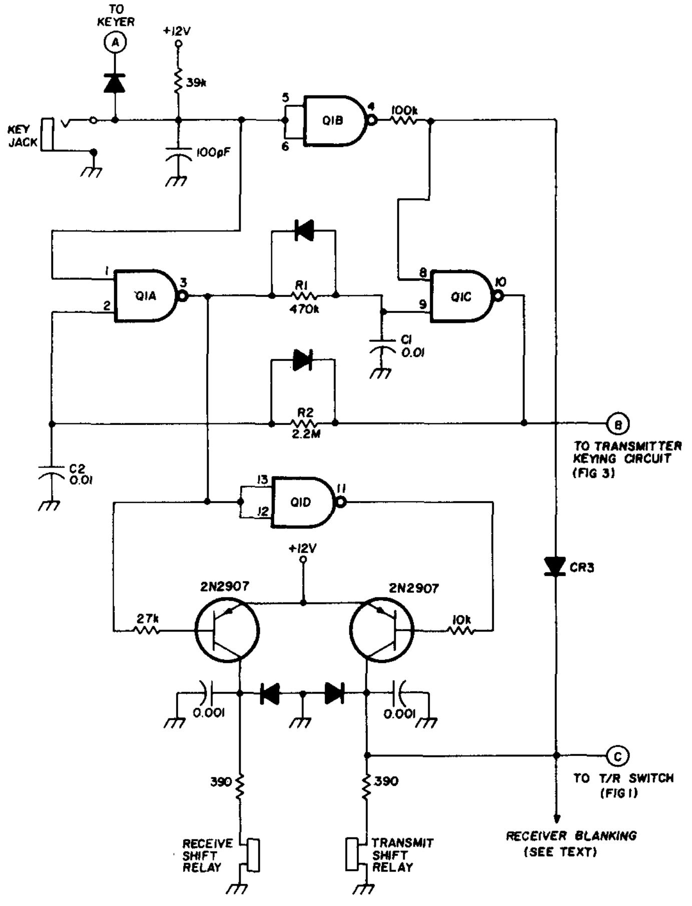


fig. 2. Sequencer circuit for CW transceiver break-in operation. Connect Q1-14 to +V and Q1-7 to ground. Q1 is a CD4011AE. All diodes are 1N4148. All resistors are in ohms ±10% ¼ watt unless noted otherwise. All capacitors in μF 50 WV unless noted otherwise.

volts will wipe out the whole semiconductor complement! There is a slight difference in optimum tuning points for transmit and receive because the 10-pF coupling capacitor appears as part of the tank capacitance during transmit. The effect is small and may be ignored if tuning is set for optimum transmitter loading. Sensitivity with the T/R circuit is considerably better than that achieved with the receiver coupled directly to the feed line (transmitter disconnected), and hash from upband shortwave broadcast monsters is to some degree attenuated.

Sequencer. The sequencing portion of the break-in system employs an RCA CD4011AE quad cmos NAND gate in a somewhat unorthodox delay circuit (**fig. 2**). The transmitter keying signal, taken from the output of gate Q1C, is held in the *high* (transmitter off) state until both inputs of Q1C go high. One input, supplied through inverter Q1B, is high when the key is closed; the second input goes high after a delay determined by R1C1. Thus, transmitter turn-on is delayed, but turn-off occurs immedately upon opening the key. On the other hand, the blanking, T/R, and shift functions are controlled by gate Q1A which goes high immediately upon key closure but is held high after key release by voltage fed back through the delay network R2C2. The delay networks include shunt diodes for fast restoration of quiescent conditions. Diode CR3, a redundant safety feature, prevents transmitter turn-on in the event that T/R control voltage is lost.

The rf portion of the original reed relay frequency-shift circuit was left intact. The miniature reed relays (which operate in less than a millisecond) are driven by 2N2907 buffers, with the fourth cmos gate used as an inverter for transmit shift. Receiver audio blanking (not shown) consists of a 2N3565 npn transistor shunted across the volume control.

Keying circuit. The transmitter cathode keying circuit (**fig. 3**) requires comment. In the original design, the 12AU6 and 1625 were keyed through separate transistors, with the cathodes unreferenced to any fixed voltage under key-open conditions. Despite the fact that no backwave was present with this circuit, severe interference with receiver operation resulted when the T/R switch was installed. It was necessary to key both stages from a common transistor (40327) and to bias the cathodes to approximately +100 volts under key-open conditions. Positive bias is obtained from network R3, R4, and R5 which also supplies the driver screen; the excitation-control pot, which was always set at maximum, was eliminated. Finally, the electrolytic keying-lag capacitors, which were of insufficient voltage rating, were replaced by a 1 μF, 400-volt paper capacitor.

Construction. Layout and construction of the break-in system is noncritical except for the T/R portion. This circuit must be located near the final tank but should be shielded from direct coupling to it. An ideal spot is on the grid side of the final-amplifier partition, with the coaxial coupling capacitor feeding through or looped around the partition. The T/R transistor circuitry should be compact.

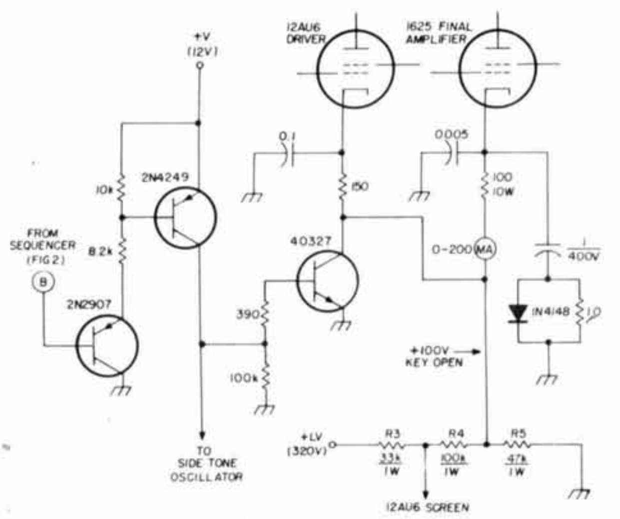


fig. 3. Transmitter keying circuit for transceiver break in operation. All resistors are in ohms ±10% unless noted otherwise. All capacitors are in μF 200 WV unless noted otherwise.

Performance. Operation is a CW man's delight. Aside from its harmonic content the sidetone signal sounds like any other on the band, and the merest tap on the receiving operator's key brings a screeching halt for "fills." Messages no longer need to be preceded by the frowned-upon "no QSK here," and DX is no longer called unwittingly from beneath somebody else's kilowatt. Contest operation is a breeze.

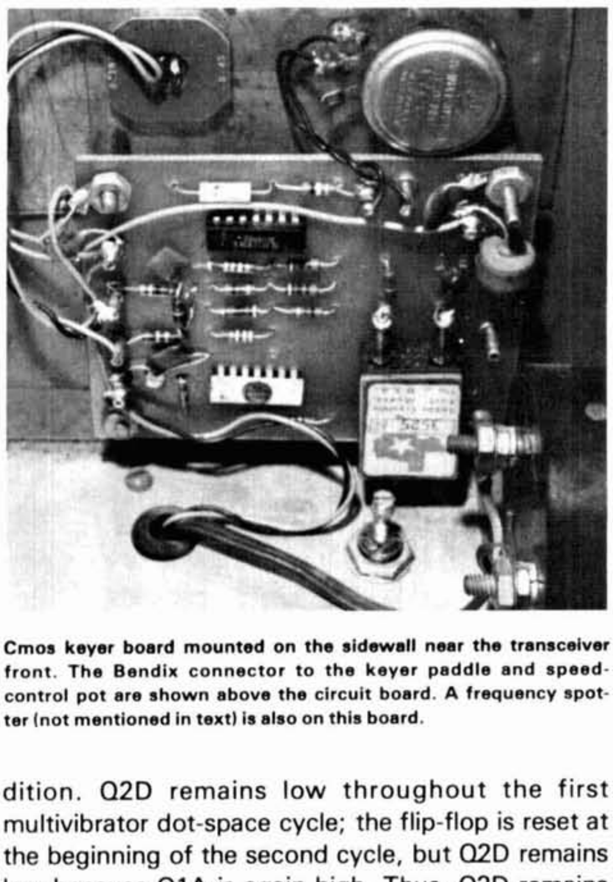
Various improvements will no doubt occur to the reader. For example, in a new design it would be logical to use a varactor diode for incremental tuning and frequency shift; this modification could be made with minor changes in the break-in circuit. Limitations of the system will also be apparent in an all-band transmitter; the clamp current might be excessive at the higher frequencies unless the coupling capacitance is reduced, resulting in loss of sensitivity at the lower frequencies. A compromise value for the capacitance should be possible.

cmos keyer

The cmos keyer (**fig. 4**) uses two low-cost CD4001AE quad 2-input NOR gates. Two of the gates (Q1A and Q1B) are used as the time-base multivibrator, and the remaining gates (Q1C and Q1D) form a dash flip-flop. Three of the remaining gates are used to synthesize a three-input NOR gate as required for dash control; the last of the eight gates controls the time-base multivibrator and also provides keyer output. Keyer output is positive (at the supply voltage) for **space** and at ground for **mark**, as required by the transceiver. Inversion or level-shifting circuitry could be added easily to adapt the circuit to other forms of keying; or a transistor driver could be added for relay keying.

Operation of the circuit is as follows: under quiescent conditions, Q1A is low, Q1B is high, Q1C is low, Q1D is high, Q2C is low, and Q2D is high. If the dot input is momentarily grounded, Q1A goes high, thereby causing Q1B and Q2D to go low, initiating a dot. If the key is released, the action stops after one dot because Q2D goes high again and locks Q1A in the low state; but if the key remains closed, the multivibrator runs free and produces a string of dots.

If the dash input is grounded, a different sequence takes place. One side of flip-flop (Q1C) receives a positive pulse of width determined by R1C1 each time a character is initiated. The effect of this pulse is to keep Q1C in the low state so that it has no effect on the operation of Q2D. However, the dash input feeds both the multivibrator and Q2A; grounding of the input to Q2A allows a positive pulse (of length determined by R2C2) to be fed Q1D at the same time Q1C is being reset. Since time constant R2C2 is longer than R1C1, Q1D conducts and the flip-flop changes state, thereby clamping Q2D in the low condition. Q2D remains low throughout the first multivibrator dot-space cycle; the flip-flop is reset at the beginning of the second cycle, but Q2D remains low because Q1A is again high. Thus, Q2D remains low for a period equal to dot-space-dot, as required for proper dot-dash ratio.



Cmos keyer board mounted on the sidewall near the transceiver front. The Bendix connector to the keyer paddle and speed-control pot are shown above the circuit board. A frequency spotter (not mentioned in text) is also on this board.

The connection from Q1C to Q2C prevents spurious setting of the flip-flop on the second

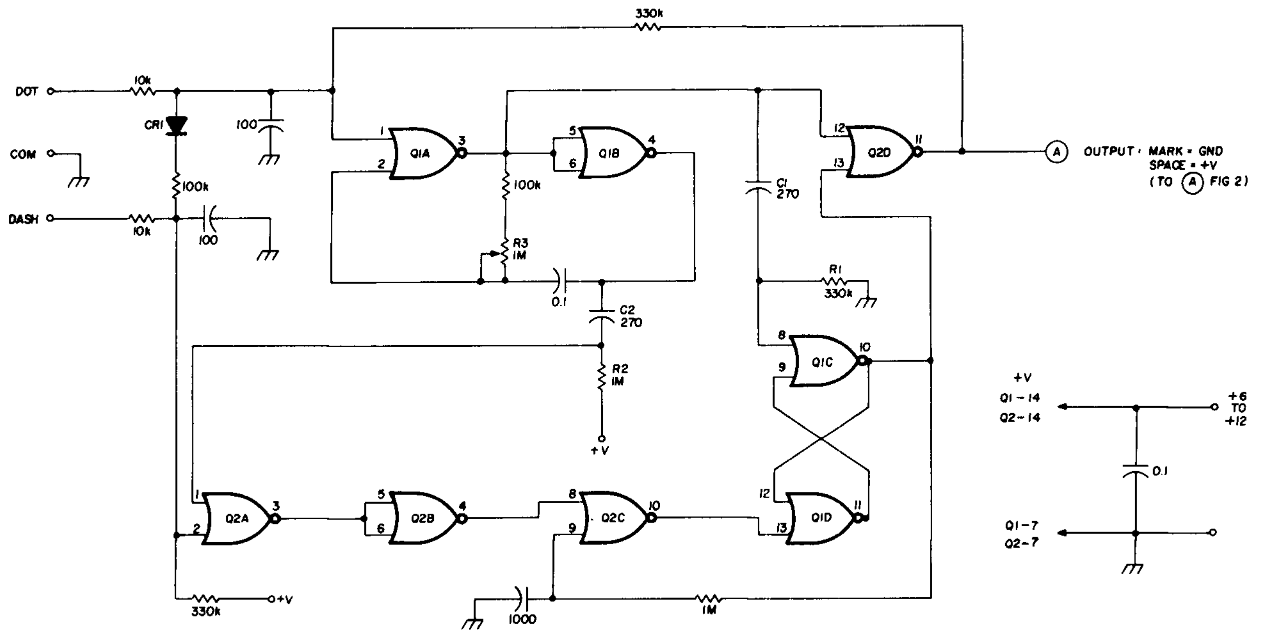


fig. 4. An ultra simple cmos keyer. The circuit uses two low-cost CD4001AE quad 2-input NOR gates. A triple 3-input NOR gate (CD4025AE) may be substituted for Q2, which allows an output for other functions.

multivibrator cycle. The 100k resistor in series with gate diode CR1 guarantees dash flip-flop setting at a higher voltage across the key than that required to start the multivibrator, which eliminates a tendency toward production of dots from the dash side; this occurred occasionally with dirty contacts.

Keying weight is determined by the multivibrator duty cycle. This, in turn, is related to the cmos input threshold voltages, which should be half the supply voltage for the desired 50% on time. Since cmos thresholds do vary (from about 33% to 67% of supply), some experimentation may be necessary to determine the best pair of gates for the multivibrator.* Duty cycle can be measured by connecting a high-impedance multimeter or vtvm to the keyer output; the meter will read half the supply voltage on high-speed properly balanced dots.

Construction. The keyer was constructed on a piece of glass epoxy circuit board without copper. Holes were drilled for component leads (including the IC pins). The components were inserted and leads crimped over. Point-to-point solder connections were made with small bus wire covered with Teflon sleeving.

This is admittedly a cheap-and-dirty construction but it has survived all kinds of banging around without failure. The circuit board and speed control, R3, are mounted on an inside wall of the transceiver. The key paddle was connected through a 3-wire cable and connector. In the transmitter keying circuit (**fig. 2**), a diode prevents the keyer from interfering with straight-key operation.

Performance. The keyer performs beautifully despite its simplicity. It draws no measurable current in the standby mode and only a fraction of a milliamp when running full tilt. The supply voltage is not critical; any voltage from less than 6 to 15 volts can be used. The upper limit is imposed by maximum cmos ratings. The speed, which is adjustable from less than 10 to more than 70 wpm, is nearly independent of the supply voltage. With the key line RC filtering shown, there's no problem due to rf pickup.

The reader with a selection of cmos devices may wish to substitute a triple 3-input NOR gate, type CD4025AE, at Q2 in **fig. 4**. In this case, one of the three gates replaces Q2A, Q2B, and Q2C; the second (with two of its inputs tied together) replaces Q2D, and the third remains free for output inversion or other desired functions. However, if cmos is to be purchased specifically for the keyer, it's better to stay with the two CD4001AEs to provide a wider choice of multivibrator gates.

acknowledgement

I'd like to express my thanks to Mark Pintavalle, who did the photography.

reference

1. Clifford Bader, W3NNL, and Richard Klinman, K3OIO, "CW Transceiver for 40 and 80 Meters," *ham radio*, July, 1969, pages 14-25.

ham radio

* It is also possible to modify the multivibrator circuit to permit adjustment of the duty cycle. See *RCA Application Note ICAN-6267*.

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high-resolution spectrum analyzer for single sideband

All you need is an oscilloscope and the simple circuits described here to display quantitative information about your ssb transmitter

Most amateurs keep abreast of recent developments in ssb transmitters and transceivers and, if you're like I am, you wonder how your equipment stacks up against the current crop. Such information, aside from boosting or busting your ego, can help you decide when equipment performance has improved enough to justify purchasing or building a new rig.

One of the most prominently reported performance parameters in equipment-review articles is intermodulation distortion (IMD). Often the IMD performance of an ssb exciter or linear amplifier is displayed as a photo taken from the CRT of an ssb spectrum analyzer. It occurred to me more than once that it would be desirable to be able to produce this kind of photo in my own station. The big hangup has been the lack of a high-resolution ssb spectrum analyzer.

This article describes the technique and hardware to allow the presentation of ssb transmitter IMD products on your oscilloscope. The hardware will cost less than $15 if you purchase everything new, and assembly should take no longer than one or two evenings. No permanent modifications to any of your other equipment will be required.

Also, for anyone not desiring to display the IMD data on a spectrum analyzer, a simpler (no cost) technique is described that will yield the IMD data to be recorded manually.

Most amateurs are familiar with the terms spectrum analyzer and panoramic adapter (Panadapter). Many have such units which have been obtained through military surplus or purchased new. Most of this equipment falls into the panoramic adapter category as defined in a *QST* article on the Heath SB-620 Scanalyzer.[1] Reference 1 differentiates between the two types of instruments, describes the basic theory of operation, and points out the most important performance limitations. That material will not be repeated here, but you're encouraged to consult reference 1 before proceeding further. For those interested in the history of spectrum analysis, more detail concerning analyzer design, or various analyzer applications, references 2, 3, 4, and 5 are recommended.

Since most of the units in the hands of amateurs are low-resolution devices (broad i-f bandwidths), they are not suitable for analysis of narrowband signals such as a two-tone test signal transmitted by an amateur ssb transmitter. Such an analysis requires a spectrum analyzer with a maximum resolution appreciably better than the spacing between the two tones. Laboratory-grade instruments costing thousands of dollars and designed for this purpose of course do the job handily. At present, the only alternative to the amateur is to buy the Heath SB-620 which, from its specifications, should provide the measurement of IMD product levels. The previous Heath Ham-scan (Model HO-13) instrument does not have sufficient resolution to perform these measurements.

The technique and hardware described below will enable the amateur not having a Heath SB-620 or access to a laboratory ssb spectrum analyzer to make scope presentations of transmitter IMD performance much like those seen in the equipment reviews of *QST*.

theory of operation

The block diagram of a spectrum analyzer is shown in **fig. 1**. To undertake construction of this whole system would be a major project. However, if you consider this block diagram in terms of equipment already present in many amateur stations, the problem becomes manageable. The blocks in the upper dotted box are part of most amateur receivers. The term narrowband i-f is of course relative, and the requirements of this particular application are discussed later in more detail. For the present, let's consider that narrowband means the narrowest selectivity position provided by the typical high-quality receiver or transceiver.

The blocks in the lower dotted box of **fig. 1** are part of all oscilloscopes. Some sweep generators are triggered; some are recurrent. Some vertical amplifiers are ac-

By Jeff Walker, W3JW, 4513 Mountain Road, Pasadena, Maryland 21122

coupled only; others are dc-coupled. The scope sawtooth may or may not be brought out to the front panel. All these factors will affect the ease with which a particular oscilloscope can be adapted to this application; however, the basics are all there.

The two blocks in **fig. 1** not included in either your receiver or oscilloscope are the swept oscillator and the detector. You may argue that your receiver has a detector and indeed it does. As we shall see, though, it's output is not easily accessible in a form that would be useful for spectrum-analyzer work.

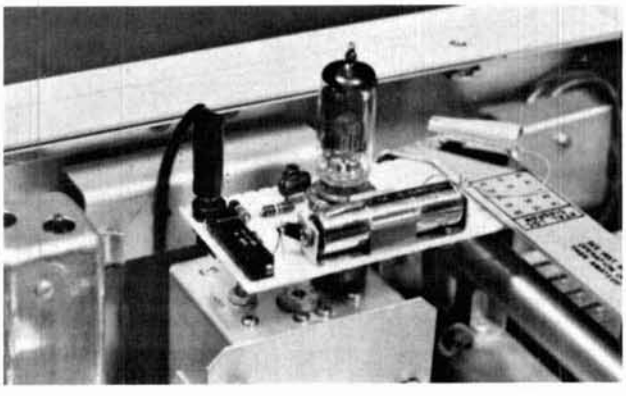
The receiver does have one or more local oscillators. If a way could be found to easily sweep this oscillator or oscillators one of our unaccounted-for blocks would be accommodated. Likewise, a detector can be a simple diode and filter network if we can determine the appropriate place to connect it.

Swept oscillator. Since a typical ssb two-tone test will employ tones from 1 to 2 kHz apart, and most applications will require the observation of only third- and fifth-order IMD products, the maximum sweep-frequency range required is 10 kHz. If you're interested in high-order products or wider tone spacing, the sweep frequency range will have to be increased.

Most modern amateur receivers are double-conversion superhets with a crystal-controlled first local oscillator (LO) and a variable-frequency second LO. Typical of this arrangement is the Collins 75S-3. The variable frequency oscillator (vfo) in this receiver is permeability tuned with a vacuum tube as the active device. It tunes from 2.7 to 2.5 kHz to give 200-kHz-wide bands. I decided it would be easier to vary the vfo frequency the required 10 kHz rather than that of the crystal-controlled first LO.

The vfo in the 75S-3 is a Colpitts oscillator; the basic circuit is shown in **fig. 2**. Collins biases diode CR1 on or off to achieve the proper vfo offset frequency when switching sidebands. I decided to use this same point (cathode of the oscillator tube) to apply my variable reactance, thus achieving swept-frequency operation. I elected to use a voltage-variable capacitance diode (varactor) as the variable-reactance source. The varactor tuning diode displays a varying junction capacitance as the reverse bias across a P-N junction is changed. (For more details on varactor diodes, see references 6-9.)

The varactor tuning diode is placed in the circuit of



Top view of the Collins 75S-3 with sweeper module installed.

fig. 2 so that the variable capacitance is from the tube cathode to ground. Bias voltage is obtained from a fixed battery source. The oscilloscope sawtooth output gives a repetitive excursion over a precisely set range of voltages that can be selected to yield the desired sweep frequency range.

Detector. As discussed previously, the receiver has a built-in detector; it puts out an audio tone near 1 kHz

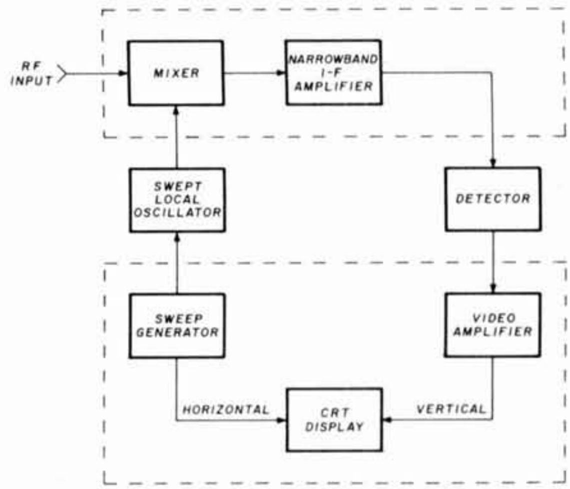


fig. 1. Basic block diagram of a spectrum analyzer.

whenever a signal is in the CW filter passband. This signal could be directly displayed on your oscilloscope, but the display would not be what is normally seen on a spectrum analyzer screen. You would see something like the display represented in **fig. 3A** as opposed to the more normal presentation of **fig. 3B**. The waveform filling the pips is the unfiltered 1-kHz audio tone.

For this application, I added an external diode detector (driven by the 1-kHz audio tone), a filter to remove any 1-kHz signal in the output, and a logarithmic shaping circuit to provide the desired dynamic range on the analyzer screen. The block diagram is shown in **fig. 4**. The hardware is built into the black plastic cover of a standard 1/4-inch (6.5mm) phone plug.

system requirements and limitations

The successful implementation of this system has several prerequisites. First, you must have a receiver or transceiver with adequate selectivity. The selectivity required is typically a 6-dB bandwidth no greater than 1/4 to 1/5 of the frequency to be resolved. That is, for 1-kHz tone spacing, the 6-dB i-f bandwidth should be 200-250 Hz or less. If you use a receiver with a 400 to 500-Hz, 6-dB bandwidth, don't plan to resolve tones much less than 2-kHz apart. Using this rule of thumb, no trouble should occur in resolving IMD products down at least 40 dB from a single tone.

Although I've not experimented with receivers having their narrow selectivity at the audio-output frequency, it should be possible to use them for this application providing care is used to keep signal levels low enough to prevent saturation of any stage preceding the selectivity.

The second prerequisite is that you must disable the agc system in your receiver. If the agc system isn't disabled, low-level signals adjacent to high-level signals on the display will be artificially depressed. If your receiver doesn't have a front-panel switch to turn off the agc, you can determine from the schematic where to break the loop.

Finally, the oscilloscope should have dc-coupled vertical amplifiers with a sensitivity of at least 100 mV/cm. The sawtooth voltage used to sweep the electron beam horizontally across the CRT should be brought out to a terminal on the front panel for use in sweeping the receiver vfo.

Possible system limitations include the lack of an rf gain control on the receiver; this control is desirable to limit the signal level to a point where all stages in the receiver operate linearly. (This function can be performed by a variable attenuator preceding the receiver.) Also, when measuring the IMD of a transmitter located close to your receiver, you must watch for chassis-to-chassis leakage above acceptable limits. This will be a problem particularly with transceivers if you get past the initial problem of operating the receiver and transmitter portions simultaneously. I strongly recommend that in all tests conducted with this technique, provision be made to place the test transmitter at some distance from the analyzer receiver, with the transmitter output attenuated and brought to the receiver through coaxial cable.

circuit details

The circuit is built in two separate modules: the vfo sweeper and the output signal processor. The sweeper circuit is shown in **fig. 5**. The MV-1404 varactor tuning diode is in series with capacitor C1. This series combination provides capacitance from the oscillator-tube cathode to ground, and thus a variable frequency. The varactor diode is biased by the 1.5-volt battery through R2, R3, and R4. The scope sawtooth (typically 100-150 volts) is reduced by the voltage divider composed of R1 in series with R3 and R4. The attenuated sawtooth is then applied to the diode through R2.

R3 adjustment determines the portion of the

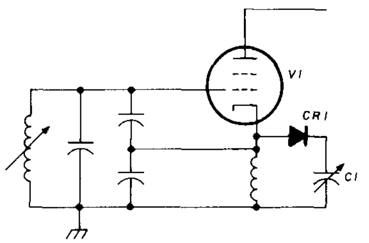


fig. 2. Basic vfo circuit of the Collins 75S-3.

capacitance-versus-voltage (C-V) characteristic of the diode over which it operates. In general, it is desirable to adjust the bias and the sawtooth amplitude so that the most linear portion of the C-V curve, which will allow the required 10-kHz sweep range, is being used.

All component values may be varied, as those listed are not the only ones that will provide proper sweep operation. If you use another varactor tuning diode, you'll have to design the circuit values as described above, using the C-V characteristics of the device you choose. For the economy-minded, the least-expensive tuning diode can probably be obtained using the technique described in reference 7.

The output signal processor schematic is shown in **fig. 6**. Diode CR1 rectifies the 1-kHz audio output. The

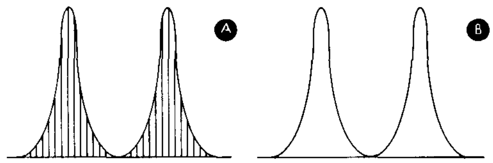


fig. 3. Spectrum-analyzer display when using a receiver detector, A; at the right is the same display on a more conventional spectrum analyzer.

signal is filtered by R1, C1. This low-frequency signal is then processed by the logarithmic shaping circuit consisting of R2, R3, CR2, and CR3. Operation of this shaper circuit depends on the fact that the junction voltage developed across the diodes has a characteristic, proportional over a large dynamic range, to the logarithm of the current applied. The 56k-series resistor allows the diodes to operate in a current-driven mode when being driven by the detector output voltage.

construction

The sweeper module is built on a piece of Vector board with holes on 0.1-inch (2.5mm) centers. The size of the board depends primarily on the size of the bias battery and its holder. Since only six small passive components are mounted on the board in addition to the battery, there's no reason for a large board if you use a battery smaller than the size AA cell shown in the photo. Those with facilities to produce PC boards should be able to lay out and etch this one in short order.

For installation in the Collins 75S-3, I epoxied the board to the top lip of a Pomona Electronics model TVS-7 test socket adapter. You'll have to punch a 5/8-inch (16mm) hole in the board to do this. The oscillator tube can now be removed from the receiver, inserted into the sweeper socket, and the whole assembly can then be plugged into the orignal tube socket. When adapting this scheme to other receivers, take care to select board and battery size that will fit into available space. For receivers that use a solid-state oscillator, it may be desirable to build the sweeper module on the smallest possible piece of board and install it permanently in or near the receiver vfo. No special board layout problems should be encountered, but keep the leads to C1 and CR1 as short as possible. The layout in the photograph is quite satisfactory. All wiring was point-to-point. Components were mounted on one side of the board with wiring on the reverse side.

The output signal processor was built into the handle of a standard 1/4-inch (6.5mm) phone plug. Use the large plastic handle type, not slim-line or right-angle types. The resistors are 1/4-watt and C1 is a miniature electrolytic. Pigtail leads work quite well with a typical oscilloscope test probe. You may want to affix a

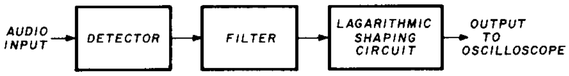


fig. 4. Block diagram of the output-signal processing system.

permanent output cable with a connector compatible with your scope; any length cable can be used.

test and operation

The only alignment required is for the sweeper module sawtooth-level adjustment. If you use any of the Collins S-Line gear and the components specified, the approximate value of R3 (**fig. 5**) is 3200 ohms. If you're using other equipment with a different vfo frequency, you'll have to use other values for R3, R4, and possibly C1 and the tuning diode. Just follow the general rule of using the most linear portion of the diode C-V characteristic that will allow the required 10-kHz sweep range.

When making your final adjustments to R3, display one of the receiver calibrator markers on the analyzer screen. You should be able to place the marker on the left or rightmost graticule line and, (assuming a 10-cm graticule) by tuning the receiver 10-kHz in the proper direction, move the marker to the other side of the graticule ±1 small division (20mm). A more demanding test of horizontal linearity and sweep range adjustment can be obtained if you have a counter or 1-kHz dial markings on the receiver vfo dial (such as on the 75S-3). With the marker positioned at any one of the cm marks across the graticule, you should move the marker horizontally on the display 1 cm for each 1-kHz rotation of the dial. Likewise, the rotation of the dial through 10 kHz should move the marker across the entire 10-cm graticule ±1 small division. Adjust R3 and, if necessary, the bias voltage to achieve this.

No adjustment is made of the signal processor output amplitude. This is a function of the receiver rf and af gain control settings, rf input signal level, and oscilloscope vertical amplifier sensitivity. These controls are set at the beginning of a series of measurements.

A typical spectrum analyzer test setup is shown in **fig. 7**. A two-tone oscillator is required to drive the audio input of the test transmitter. Several good articles on the subject have been published, including those by Hank Olsen[10] and Ray Colvin.[11]

Tune the transmitter for desired test power level and terminate the transmitter into a high-power attenuator or a dummy load with an rf coupler to sample the rf for the spectrum analyzer. Locate the test transmitter away from the spectrum-analyzer receiver to prevent unwanted coupling. The power attenuator output or sampled rf should be fed to the receiver through a low-power step attenuator.

Remove the receiver vfo tube, insert it into the sweeper module, and install the sweeper in the vfo tube socket. Connect a lead from your oscilloscope sawtooth output to the sweeper input terminal. Be sure to ground the sweeper to the receiver chassis ground. Insert the output signal processor into the phone jack and connect its leads to the oscilloscope vertical input. Be sure the receiver and oscilloscope have a common chassis ground.

You are now ready to display the two-tone test signal with its associated IMD products. Adjust the vertical amplifier sensitivity to approximately 100 mV/cm. Adjust the scope controls for the clearest trace possible. Key the transmitter and adjust the receiver tuned frequency near the transmitter output frequency. You'll find that the receiver calibration will be low by 10 to 15 kHz because of the minimum capacitance added to the vfo circuit by the sweeper module. It may be easiest to tune the receiver with the audio output coupled to your speaker.

After you hear the two-tone signal being swept, you can then plug the output signal processor into the phone jack. Adjust the receiver rf gain control and the step attenuator for the equivalent of an S9 to S9+20 dB signal. Be sure to turn the receiver agc control *off*! Adjust receiver rf gain, af gain, and oscilloscope vertical sensitivity to get a 40-dB dynamic range on the CRT. This can be accomplished by setting the rf and af gain controls to bring the two-tone test signal just barely out of the noise with the step attenuator increased 40 dB beyond a reference setting. Then, after returning the attenuator to the reference setting, adjust the af gain and scope vertical sensitivity for a full screen display. Repeat this procedure several times, and you should have the controls set for a 40-dB dynamic range.

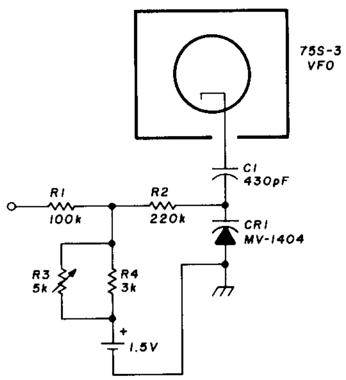


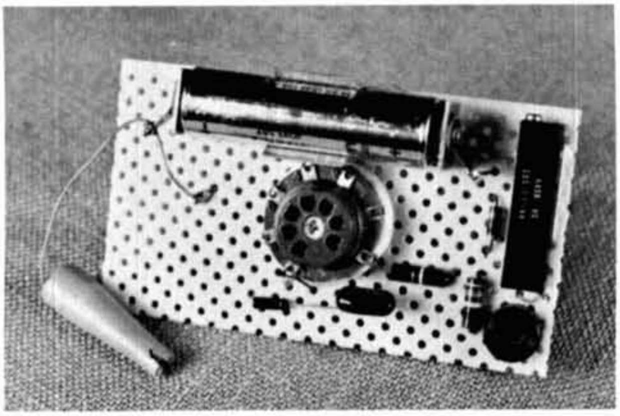
fig. 5. Sweeper module schematic. All resistors are ¼ watt.

Defining the maximum signal as 0 dB, you'll find that the linear display distance for the first 10-dB increment down is not as great as for the second 10-dB increment, and so on down through the fourth 10-dB increment. This is normal operation and is the price paid for using a relatively simple log shaping circuit. Using a circuit that would give a 10-dB/cm display over the entire 40-dB dynamic range would increase complexity and cost by several times and was not considered worthwhile.

If your step attenuator has 1-dB increments, you can use it with the 0-dB reference point at the top of the display to calibrate any point to the nearest dB. For most tests, 10-dB calibration marks on the CRT will be adequate. Examples of this type of calibration are shown in the section on results.

spectrum analysis without hardware

Earlier I stated that a no-cost method of obtaining the IMD performance of a transmitter would be given. The test setup for this method is shown in **fig. 8**. The same setup as before is used. In this case, the sweep action is accomplished manually by tuning the receiver vfo. The amplitude display is the S-meter. It is not necessary to disable the agc for this method.



Close up view of the sweeper module showing the small number of parts required. The unit is plugged into the VFO tube socket.

After the test transmitter is keyed with a two-tone input signal, tune to either of the two tones using the most selective i-f bandwidth available. If you have 100- to 250-Hz CW selectivity, you can resolve tones spaced 1 kHz apart to at least 40 dB. If you have 400- to 500-Hz selectivity, you can resolve tones spaced 2 kHz apart. If your receiver has 1-kHz dial calibration, this test is a lot easier to perform. For instance, if you use 1-kHz tone spacing, the two desired tones will be 1 kHz apart; and third-, fifth- and higher-order IMD products will be spaced at 1-kHz increments above or below the highest and lowest of the two desired tones respectively.

Set the step attenuator and receiver rf gain control so that the S-meter reads about S9+20 dB (this may vary depending on your S-meter characteristics). Just be sure you're not hitting the receiver with so much signal that it is saturating.

Now tune to the desired IMD product and use the step attenuator and rf gain control to set the level of this IMD product to a reference point on the S-meter, such as S-2 or S-3. Keep in mind that you can't increase the signal input to the receiver through the step attenuator to any great extent or the *desired* tones will saturate some portion of the receiver or desensitize it.

Now, retune to the closest desired tone. Increase the step attenuator setting until the desired tone is reduced to the reference level previously set for the IMD product (S-2 or S-3). The increase in attenuator setting is the number of dB that particular IMD product is below a single tone. To obtain the number of dB below PEP output, add 6 dB. Each IMD product (third, fifth, seventh, etc., on each side of the desired tones) must be measured using the above procedure.

This method requires a lot of manual effort but yields the same basic data as does the spectrum analyzer. The advantage of this method is that the only equipment required is a step attenuator and a receiver with good CW selectivity.

results

The ssb spectrum analyzer, built around the vfo sweeper and output signal processor modules, a Collins 75S-3 receiver with 250-Hz crystal filter, and a Hewlett-Packard model 170A oscilloscope have been in use at my station for about a year. Results have been gratifying. The 250-Hz, 6-dB bandwidth of the 75S-3 CW position allows me to resolve two signals 1 kHz apart to better than 40 dB on the spectrum analyzer screen.

Examples using a signal generator and the 32S-3 transmitter. Fig. 9A shows the spectrum of the SG-823/URM-144 two-tone test signal generator. This test set transmits two tones spaced a variable audio frequency apart. The generator is specified to have all IMD products down greater than 60 dB. This particular display is set up for a dynamic range slightly greater than 40 dB, and no IMD products are visible. This is what the *perfect* SSB transmitter output would look like! In this figure, the two tones are about 2 kHz apart. Ten-dB calibration marks are seen on the vertical centerline and at the side of the picture. The horizontal scale factor is 1 kHz/cm, and the sweep rate is 100 ms/cm.

Fig. 9B is a multiple-exposure photo of the same two-tone signal generator output with the signal level into the spectrum analyzer reduced 10 dB for each successive exposure. Note that the 0 dB reference is on the second horizontal graticule line from the top. The first 10-dB increment is not the same linear distance down on the screen as the succeeding 10-dB increments. A grease pencil was used to mark each 10-dB calibration point on the CRT face; these points were then transferred to the side of the picture.

Fig. 9C shows the spectrum of a Collins 32S-3 transmitter operating at maximum PEP output (117 watts). The frequency is 4 MHz and tone spacing is 1 kHz. The third-order IMD products on either side of the desired two tones are 31 dB down from a single tone (37 dB below PEP, which was determined using the step attenuator). The fifth-order products are barely visible 1

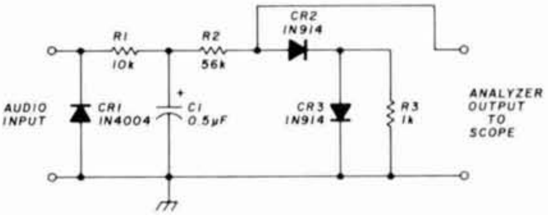


fig. 6. Output signal-processor schematic. All resistors are ¼ watt.

kHz on either side of the third-order products; they are just beginning to emerge from the noise. The horizontal scale is 1 kHz/cm and the sweep rate is 100 ms/cm.

Fig. 9D is the time-domain equivalent of the 32S-3 output pictured in the frequency domain of **fig. 9C**. This is the classic two-tone-test scope picture that amateurs strive to obtain from their ssb transmitters. The picture tells you that the transmitter is performing properly; but the previous picture, in the frequency domain, tells you much more quantitatively and will alert you more quickly that something is deteriorating IMD performance. This picture has a horizontal scale factor of 1 ms/cm. The frequencies of the two test tones are 1 kHz and 2 kHz. Note the 10-dB calibration marks on the vertical centerline used for the spectrum-analyzer display.

Example using an exciter-amplifier combination. Fig. 10A is the spectrum of a homebrew linear amplifier driven by the 32S-3 and operating well within its 500-watt PEP output ratings. Tone spacing is 1 kHz. Only the third-order IMD products are visible, which are 35 dB down (verified using the step attenuator). As before, the horizontal scale is 1 kHz/cm and sweep rate is 100 ms/cm.

Fig. 10B is a time-domain photo taken under the same transmitter conditions as for **fig. 10A**. As in **figs. 9C** and **9D**, the frequency-domain photo gives much more quantitative information. The horizontal scale for **fig. 10B** is 1 ms/cm and the vertical scale is 200 mV/cm.

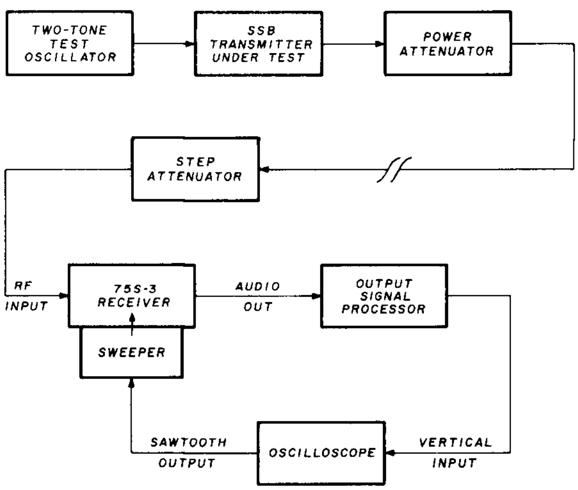


fig. 7. Typical test setup for measuring intermodulation distortion of an ssb transmitter.

Fig. 10C is of the same linear amplifier-exciter combination used for **fig. 10A**. In this case, however, the equipment was purposely misadjusted. Note that the third-order IMD products are now only 19 dB down. The fifth-order products have increased and are 33 dB below a single tone. The seventh-order products are just barely visible at the noise level. No final amplifier grid current was flowing under the operating conditions but 15 mA (positive) more screen current than normal is flowing. Whenever you see a two-tone test spectrum

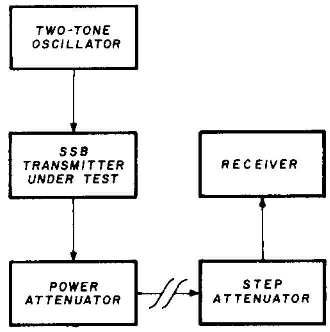


fig. 8. Test setup for a no-cost intermodulation distortion measurement method.

proportioned like this, you can be sure something is defective or out of adjustment. The tone spacing used here was 1 kHz, horizontal scale is 1 kHz/cm, and the sweep rate 100 ms/cm.

Fig. 10D is a picture taken under the same conditions as those of **fig. 10C**. The difference is that the display has been enlarged to fill the picture horizontally. The effective sweep speed in **fig. 10D** is 50 ms/cm; however, in this mode of operation the oscilloscope horizontal amplifier gain was increased by a factor of two – thus no loss of resolution occurred. The horizontal scale is 0.5 kHz/cm, and the vertical-scale 10 dB marks are as before. If I had left the sweep magnification at *X1* (normal) and simply changed the sweep speed control to 50 ms/cm, the display would have occupied the same position as in **fig. 10C** but resolution would have been lost (i.e., the pips would have overlapped). Note the ripple on the pips. This is a result of a small portion of the 1-kHz receiver detector audio tone passing through the lowpass filter in the output signal processor. I prefer the appearance of the display in **fig. 10C; 10D** looks a little busy plus the fact that with this expanded horizontal scale you can't see IMD products beyond the fifth-order.

Examples with equipment misadjusted. Fig. 11A is of the same transmitter as **figs. 10A** and **C**. In this case, the amplifier was misadjusted and the alc loop disabled. Approximately 25 mA (positive) more screen current than normal was flowing and 0.2 mA of grid current was flowing. The results are degradation of the third-order products to 16 dB below a single tone. The fifth-order products are about the same as those in **fig. 10C**, but the seventh- and ninth-order products are approaching the −40 dB level. The horizontal scale factor is 1-kHz/cm; sweep rate is 100 ms/cm.

Fig. 11B shows the time-domain display corresponding to the conditions of **fig. 11A**. It is obvious from this photo that something is wrong with transmitter adjustment. I believe you will agree, however, that there is not as much difference between **figs. 9D** or **10B** and this one as there is between **9C** or **10A** and **11A**! The horizontal scale factor is 1 ms/cm and the vertical scale is 200 mV/cm. The two tones are 1 and 2 kHz.

Example using a phasing-type exciter. Fig. 11C is the

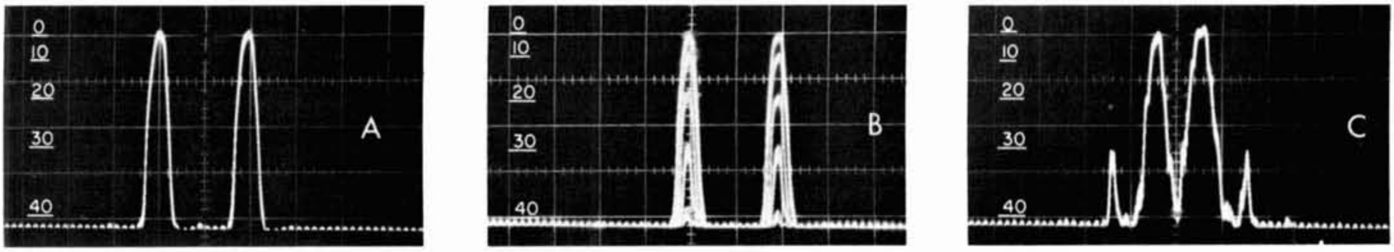


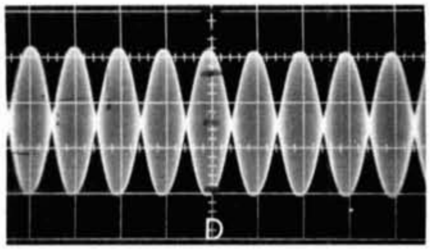
fig. 9. Examples using a two-tone test signal generator and the Collins 32S-3 transmitter.

A. Spectrum of an SG-823/URM-144 two-tone signal generator. Tones are about 2 kHz apart. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.

B. Multiple-exposure photo of the SG-823/URM-144 two-tone output with signal level into the spectrum analyzer reduced 10 dB for each exposure. Note relationship between each 10-dB increment on vertical scale, which is nonlinear. Horizontal scale factors are as in photo A.

C. Spectrum of a Collins 32S-3 transmitter operating at maximum PEP output (117 watts). Fifth-order products are barely visible 1 kHz on either side of the third-order products. Horizontal scale factors are as in photo A.

D. Time-domain equivalent of the 32S-3 output shown in C. This is the classic two-tone-test scope picture of amateur ssb transmitters. Proper transmitter operation is depicted, but the picture in C gives quantitative information on intermodulation distortion. The two-tone-test frequencies are 1 kHz and 2 kHz. Horizontal scale: 1 ms/cm.



spectrum of a phasing-type ssb exciter. The suppressed-carrier frequency is at the vertical centerline. The two largest signals to the right of the centerline are the desired lower sideband tones. The two pips located 1cm on either side of these two tones are the desired sideband third-order IMD products (26 dB below a single tone). Note that the left (high side) IMD product is at the carrier frequency. The carrier suppression for this rig was checked and found to be greater than 40 dB. So, this pip is the IMD product and not the carrier.

The fifth- and seventh-order IMD products (low side) are visible on the right. Both are about 37 dB below a single tone. At the first- and second-centimeter lines left of the vertical centerline are the two tones of the suppressed or undesired upper sideband. Note that the upper sideband is only suppressed 17 dB with respect to the lower sideband. The third-order IMD product to the right of these undesired pips is buried under the left-side third-order product from the desired sideband. The small pip at the left side of the display could be the high-side third-order product for the two tones of the undesired sideband, or it could be the ninth-order product (high side) of the desired sideband. As you can see, a lot of information is contained in this photo. The horizontal scale factor is 1 kHz/cm and the sweep rate is 100 ms/cm.

concluding comments

In comparing the time- and frequency-domain pictures here, it's clear that the information contained in the frequency-domain pictures is much greater. Also, when testing or adjusting an ssb transmitter or exciter using the spectrum analyzer, more noticeable changes occur in the display as you change transmitter operating conditions. This is particularly true as you approach optimum transmitter adjustment. That last little improvement is hard to detect on a time-domain picture.

Some points to ponder in using this equipment: The

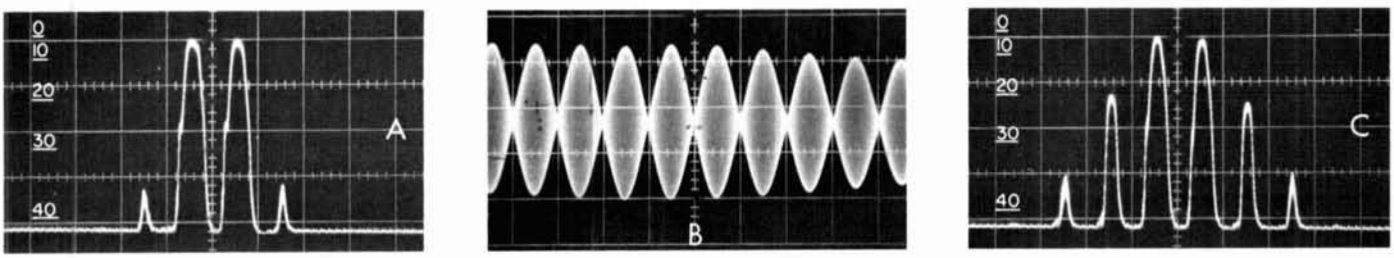


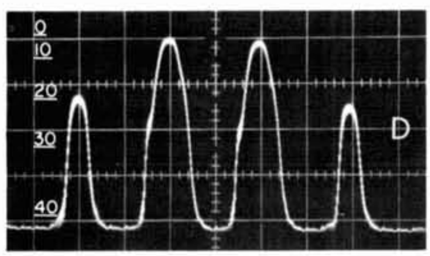
fig. 10. Example using the 32S-3 transmitter and a homebrew linear.

A. Spectrum of a homebrew linear amplifier driven by a 32S-3 operating within its output rating. Tone spacing is 1 kHz. Only third-order intermodulation distortion products are visible, which are 35 dB down. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.

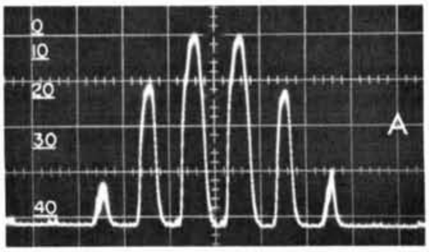
B. Time-domain spectrum of the setup in photo A. Horizontal scale: 1ms/cm; vertical scale: 200 mV/cm.

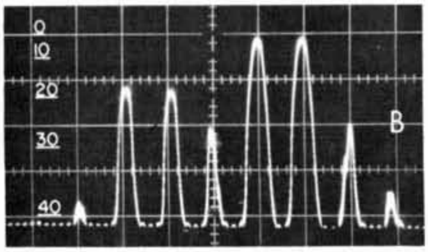
C. Same equipment setup as for photo A, except that equipment was purposely misadjusted. Note third-order intermodulation distortion products are 19 dB down; fifth-order products have increased and are 33 dB below a single tone. Seventh-order products are barely visible in the noise. Tone spacing was 1 kHz. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.

D. Spectrum picture taken under the same conditions as those of C, except that display has been enlarged to occupy entire horizontal area. Scope horizontal amplifier gain was increased by a factor of two to increase resolution. Horizontal scale: 0.5 kHz/cm.



sweeper bias battery life should equal shelf-life of the cell. Most of the pictures were made using a scope sweep speed of 100 ms/cm. This is a relatively slow speed and causes noticeable flicker on a normal oscilloscope CRT. If you attempt to operate at higher sweep rates you'll lose resolution. With 200-Hz receiver selectivity and 2-kHz tone spacing, you may be able to sweep at 50 ms/cm or possibly 20 ms/cm. If you attempt to photograph the analyzer display, the greatest success will be had using single-sweep operation of the scope while holding the camera shutter open (you must use a hood or dark room).





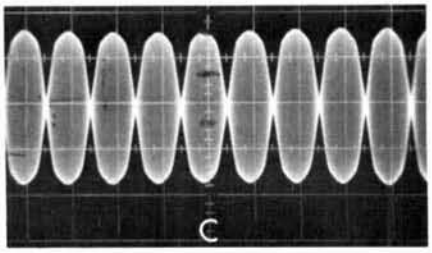


fig. 11. Examples with equipment misadjusted.

A. Spectrum of the same transmitter setup as in figs. 10A and 10C, except that the amplifier was purposely misadjusted and the ALC loop disabled. Third-order products are about the same as in fig. 10C, but seventh- and ninth-order products approach −40 dB. Horizontal scale: 1 kHz/cm; sweep rate: 10 ms/cm.

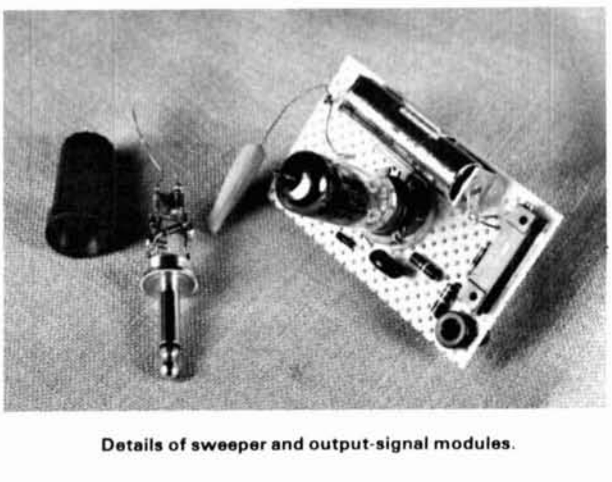
B. Time-domain display corresponding to the conditions of those in fig. 11A. Horizontal scale: 1 ms/cm; vertical scale: 200 mV/cm.

C. Spectrum of a phasing-type ssb exciter. Upper sideband is suppressed only 17 dB with respect to lower sideband. Fifth- and seventh-order intermodulation distortion products are 37 dB below a single tone. Horizontal scale: 1 kHz/cm; sweep rate: 100 ms/cm.

The display on my spectrum analyzer sweeps from left to right, but the highest input frequency is on the left also (increasing frequency from right to left). This condition is brought on by the mixing scheme of the 75S-3. It's quite easy to operate with experience.

The mixing scheme in some receivers will result in more-normal increasing frequency from left to right.

The linearity of this sweeper circuit is quite good considering its simplicity. Referring to **figs. 10D** and **11C**, you can see that the pips are 2 and 1 cm apart, respectivly, ±1 small division. This fact will help you properly identify IMD products by counting centimeter spacing, providing you use tones spaced exactly 1 or 2 kHz apart.



Details of sweeper and output-signal modules.

I mentioned earlier that the minimum capacitance of the sweeper module would upset the dial calibration of your receiver. In my case the calibration was offset 12 kHz low. If your measurements require an accurate knowledge of the carrier (suppressed carrier) frequency, I recommend the use of some form of digital dial readout. If you use a readout such as in the Heath SB-650 or the frequency measuring system I described,[12] your readout will automatically account for the sweeper offset. If you use a readout that is preset and adds or subtracts the vfo frequency, you'll have to recalibrate the preset.

I hope this inexpensive method of high-resolution ssb spectrum analysis will encourage more amateurs to reap the advantages of frequency-domain analysis. Once you start using a spectrum analyzer, whether for ssb IMD distortion analysis, harmonic distortion analysis, or in search of oscillations and spurious emissions, you'll find that you'll be quite dependent on the instrument.

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rat-race balanced mixer for 1296 MHz

Details for an easily built mixer for the 1296-MHz amateur band which can be used in both receiving and transmitting applications

The heart of most successful uhf ssb transceivers is a bilateral mixer, either singly or doubly-balanced. In previous articles[1,2] I have published designs for several such mixers. All have been built successfully by a number of readers, but each exhibited design characteristics which restricted universal reproducibility. I have thus received numerous requests, primarily from amateurs in Europe and Asia, to develop a mixer which could be built from readily available materials, without specialized dielectrics, advanced metalworking techniques, or expensive commercial microcircuits.

Actually, such a mixer was published some time ago by Paul Wade, WA2ZZF.[3,4] His mixer used a microstripline quadrature hybrid, and can be easily etched on a double-sided, glass-epoxy printed-circuit board at minimal cost. As a receive converter, the unit yields an outstanding noise figure because of its inherently low conversion loss.

Unfortunately, the Wade mixer falls a bit short of the mark in transmit service, due in part to its limited dynamic range, moderate isolation, and difficulties in obtaining a good wideband impedance match. The rat-race mixer described here overcomes some of these limitations and preserves the reproducibility of Wade's design.

mixer anatomy

Basically, any passive mixer assembly can be functionally divided into three segments, as shown in **fig. 1**. The coupler serves to apply components of the rf and local-oscillator (LO) signals to the mixer diodes in the correct phase relationship, and may consist of transmission-line delay networks, resistive or reactive power dividers, balun transformers, coaxial or waveguide directional couplers, hybrid couplers, or some combination. The coupler may be built with lumped constants, coaxially, with toroidal transformers, in stripline, or in microstrip.

The nonlinear network in which sum and difference frequencies are generated is represented by the diode array of **fig. 1**. Unbalanced mixers (which afford no isolation between the LO and rf ports) use a single diode, while most single-balanced mixers use a matched diode pair. Double-balanced mixers typically include four diodes in either a ring or bridge arrangement, while special-purpose mixers which offer image cancellation or extra-wide dynamic range may contain an array of eight or more diodes.

Diode selection is based upon noise figure, conversion efficiency, and the signal amplitude requirements of the system. Point-contact cartridge diodes, such as the 1N21 series, were the accepted standard in microwave mixers for many years. These devices are still used in high-power radar applications where diode burnout immunity and ease of replacement are major considerations.

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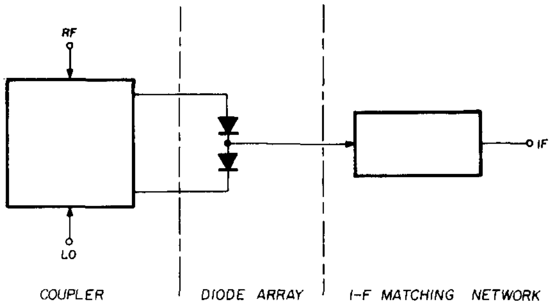


fig. 1. Basic anatomy of a balanced mixer. The coupler may take several forms as shown in fig. 2. The i-f matching network is designed to match the impedance of the diodes to 50 ohms.

The current favorite, so far as the uhf experimenter is concerned, is the Schottky barrier or *hot-carrier* diode. These diodes are available in low-cost glass packages from such microwave industry leaders as Aertech, Alpha, Hewlett-Packard, Microwave Associates, Parametric Industries and others, and combine low conversion loss with excellent immunity to rf burnout. An additional advantage, so far as balanced mixers are concerned, is that the manufacturing variations for hot-carrier diodes are minimal. This yields, for any two diodes of the same part number, rf characteristics which are closely matched, thereby eliminating the need for selecting matched pairs or quads of diodes to achieve good mixer balance.

Where the ultimate in conversion efficiency is required (especially at the higher microwave frequencies), tunnel diodes excel. However, their high cost and relative susceptibility to rf overload have limited their acceptance by radio amateurs.

Regardless of the diodes which are selected for the mixer's nonlinear network, it is unlikely that they will provide a good impedance match to 50 ohms at the i-f port. Bear in mind that diode impedance varies with diode current, which in turn is a function of local-oscillator injection level, applied dc bias, and the diodes' dc return path. The complex impedance of the i-f port, under the desired operating conditions, should be transformed to 50 ohms. To minimize intermodulation responses, this transformation should be effective at both the desired i-f frequency *and* the undesired mixing product (usually the *sum* frequency). This will serve to eliminate reflections at any frequency from re-entering the nonlinear network through the mismatched i-f termination. The effects of image frequency termination are discussed in detail in reference 5, as well as section 2.10 of reference 6.

The required impedance transformation at the i-f port can be accomplished by applying conventional techniques to the design of L, T, or pi networks. These networks also serve to suppress any components of the rf or LO signals which may be present at the i-f port because of mixer imbalance.

hybrid selection

The generalized balanced mixed circuit in **fig. 1** allows considerable latitude in the selection of coupling elements for applying the LO and rf signals to the mixer diodes. Hybrid couplers are available for dividing the input signals into two equal components, which are either in phase, 180° out of phase, or in phase quadrature.[7] Although mixers have been designed around virtually every imaginable coupler, the arrangements considered for this design were two versions of the 90° branch hybrid or quadrature power divider, and two types of 180° ring

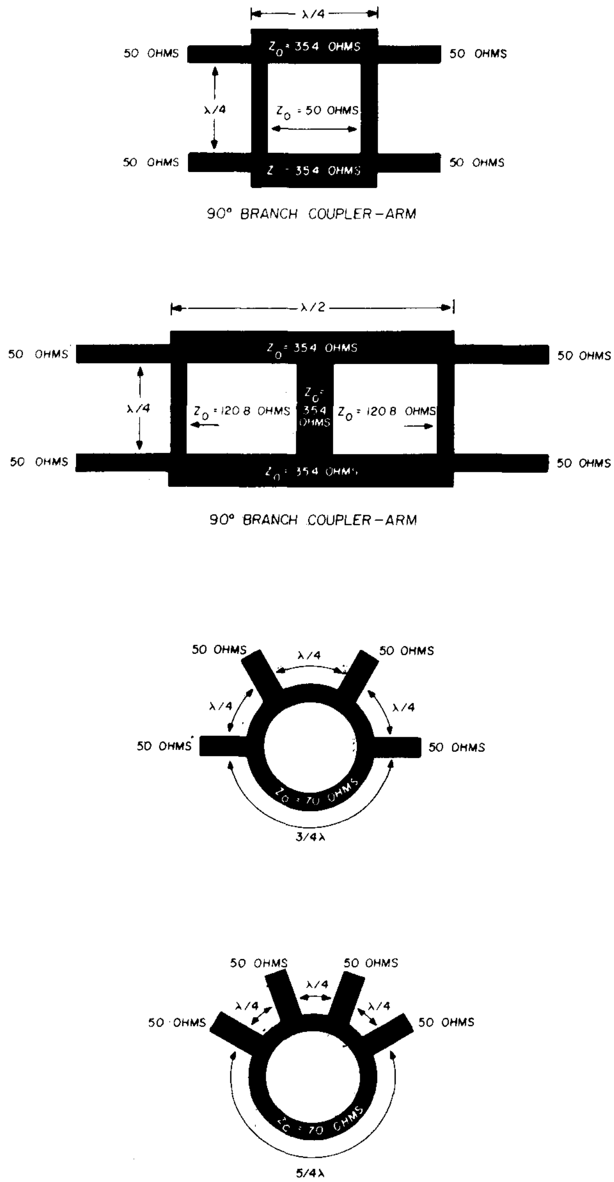


fig. 2. Four types of microwave hybrid couplers which are used in single-balanced diode mixers. Performance of the various types is compared in table 1.

hybrid or *ratrace* (see **fig. 2**). Each of these hybrids can be easily built with etched microstripline, and each offers certain performance advantages which should be considered.

Assuming an rf operating frequency of 1296 MHz, a 144-MHz i-f would suggest a LO frequency of 1152 MHz. Since the mixer hybrid must pass both rf and LO components, it is reasonable to specify 1224 MHz, the mean of these two frequencies, as the design center frequency for the hybrid. The ripple passband of the hybrid must, of course, include both the rf and the LO frequencies; thus the design must permit an operating bandwidth of 144 MHz, or 12%. Physical construction of the hybrid in microstripline, however, could introduce several sources of error as to actual center frequency. Because of the effect of variations in substrate dielectric constant, or that of cumulative dimensional tolerances, conservative design philosophy suggests characterizing the hybrid over a somewhat greater bandwidth. For this study a 20% operating bandwidth was assumed.

Table 1 compares various operating characteristics of the couplers shown in **fig. 2** over this 20% bandwidth. Information regarding vswr, isolation, and imbalance was determined from published design charts;[8] the relative microstripline losses at the center frequency were derived empirically.

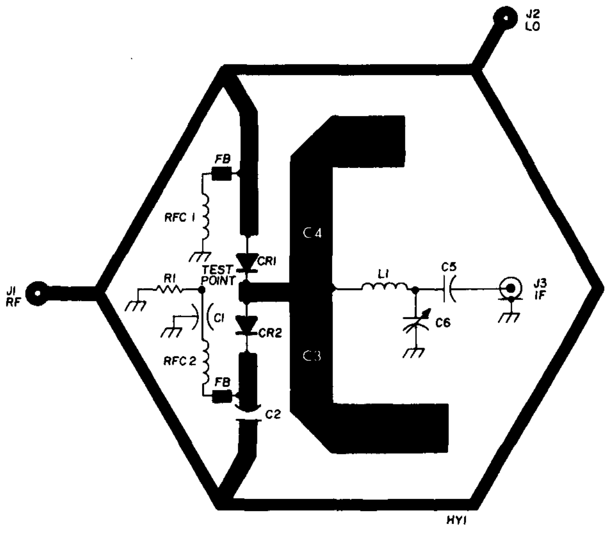
Since all of the listed hybrid characteristics contribute in varying degrees to the overall performance of the resulting mixer, a somewhat subjective tradeoff process is necessary to select the optimum mixer hybrid. I began by excluding the popular 90° branch coupler because of its high vswr and relatively poor isolation. Similarly, I eliminated the extended ring hybrid on the basis of its high stripline losses and poor amplitude balance. In selecting between the two remaining candidates, I opted for minimizing amplitude imbalance and stripline losses, trading off

table 1. Performance comparison of mixer hybrids over a 20% bandwidth.

hybrid type	vswr	amplitude imbalance	isolation	relative insertion loss
90° branch – 2 arm	1.45	0.7 dB	14.0 dB	1.0
90° branch – 3 arm	1.12	0.5 dB	25.3 dB	1.7
180° ring – 3/2λ	1.14	0.4 dB	23.0 dB	1.5
180° ring – extended	1.40	0.9 dB	23.0 dB	2.0

the resulting slight degradation in vswr and isolation. The form factor of the final mixer was a further consideration, although I must admit that a toss of a coin may have been just as scientific!

It is interesting to note that other uhf experimenters have also selected the 3/2 wavelength, 180° ring hybrid for mixer service.[9,10,11] All have achieved respectable mixer performance.



C1	1000 pF feedthrough capacitor
C2	100 pF chip capacitor (ATC 100B or equivalent)
C3	25-ohm open-circuit microstripline, 1 quarter-wavelength long at 1152 MHz (see fig. 4)
C4	25-ohm open-circuited microstripline, 1 quarter-wavelength long at 1296 MHz (see fig. 4)
C5	0.001 μF ceramic capacitor
C6	10-40 pF cermic trimmer capacitor
J1, J2	SMA coaxial receptacle (E.F. Johnson 142-0298-001)
J3	BNC coaxial receptacle (UG-1094/U)
L1	4 turns no. 20 (0.8mm), air wound, 3/8" (9.5mm) diameter, 1/2" (13mm) long
R1	10 ohm, ¼ watt carbon composition resistor
RFC1, RFC2	miniature 0.33 μH molded rf choke with ferrite bead slipped over hot end
Z1	70-ohm rat-race hybrid at 1224 MHz (see fig. 4)
CR1, CR2	hot-carrier diodes (Hewlett-Packard HP 5082-2817)

fig. 3. Schematic diagram of the rat-race balanced mixer for 1296 MHz. Full-size printed-circuit layout for the etched circuit board is shown in fig. 4.

The decision to use microstripline construction on fiberglass-epoxy PC board was due primarily to the success which Wade achieved with this medium, and partly due to my frustration in trying to duplicate the results of others. Rat-race mixers of slab-type[9] or coaxial[10] construction demand greater patience and mechanical ability than I possess, and the slight, almost immeasurable performance advantage of etching microstripline couplers on expensive Teflon substrates[11] didn't seem worth the effort.

i-f selection

Factors governing the selection of a conversion system's intermediate frequency include the band width capabilities of hybrid junctions, image rejection limitations in practical bandpass filters, the availability of i-f equipment or components, and spurious responses (related to possible harmonic relationships between signal, LO, and i-f). Unfortunately, the first of these considerations seems to suggest selection of a low i-f, the second generally favors a high i-f, and the third all too often dictates selection of an i-f which violates the fourth!

I have built various conversion systems for 1296 MHz, using amateur 28, 50, 144, and 432 MHz bands as a receive and transmit i-f. My experiences tend to favor the 2-meter i-f as a good compromise between mixer bandwidth and image rejection. The recent proliferation of 2-meter ssb transceivers further supports this choice. However, the actual operating frequency within the band must not be selected arbitrarily. For example, placing a 1296.0-MHz signal at a 144.0-MHz i-f would require a 1152.0-MHz LO. More than one experimenter (myself included) has started with a 48.0-MHz crystal oscillator, multiplied by 24 to 1152 MHz, and ended up with a strong interfering signal at 144 MHz on the receiving dial.

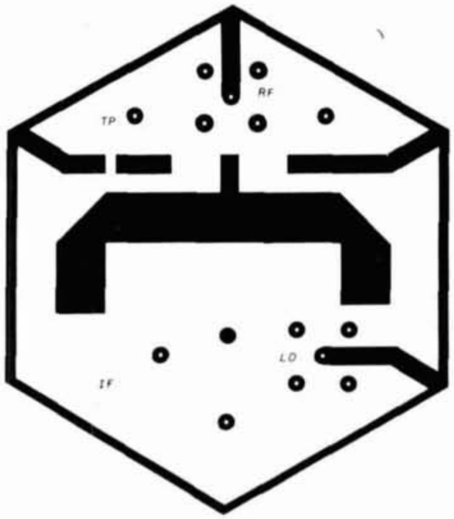


fig. 4. Full-size layout for the rat-race mixer circuit board. Drilling instructions are shown in fig. 5.

The spurious signal, of course, is the third harmonic of the crystal oscillator. If you try to circumvent this difficulty by starting at 96.0 MHz and multiply by 12, it's possible to end up with transmitted spurious signals from the ninth harmonic of the injected i-f signal.

Another constraint on the i-f selection is the undesirability of having a popular operating frequency in the 1296-MHz band fall at (or worse, just past) the end of the i-f tuning range. For best results, 1296.0 MHz should fall in the middle of the tuning dial. In my present system, this dictates an i-f of 145.1 MHz; a 1150.9-MHz LO derived from a 95.9083-MHz crystal keeps me free from birdies.

Although most of my activity on 1296 MHz is on ssb, the beauty of a linear translation scheme is that it's compatible with virtually any mode. This mixer will accommodate fm, a-m, video, or whatever. Depending on the mode, tradition, rather than technology, may govern the selection of operating and i-f frequencies. A group planning narrow-band fm simplex, for example, might choose to place 1290.52 MHz on 146.52 MHz. Nonetheless, the above considerations deserve attention, especially when selecting the LO multiplying scheme.

The mixer design presented here will actually accommodate an i-f between about 70 and 150 MHz, with little noticeable performance variation. Obviously, great care should be exercised when selecting the actual i-f and oscillator frequencies.

construction

Fig. 3 shows a schematic diagram of the ring hybrid or rat-race mixer for the 1296-MHz amateur band. Note that the microstripline hybrid, which is normally built as a circular trace with a circumference 3/2 wavelengths long, is shown as a hexagon with quarter-wavelength sides. I used this same simplification in an earlier mixer,[12] as did Dick Bingham, WB6BDR.[11] Neither of us noticed any measurable degradation over the usual circular arrangement. Leroy May's successful rat-race mixer (reference 9) used a triangular layout, and this led me to believe that the layout itself is relatively unimportant. The 70-ohm characteristic impedance of the ring, however, is critical, and requires a 0.05 inch (1.3mm) microstripline width for G-10 circuit board.

Aside from the diode bias path (to be discussed

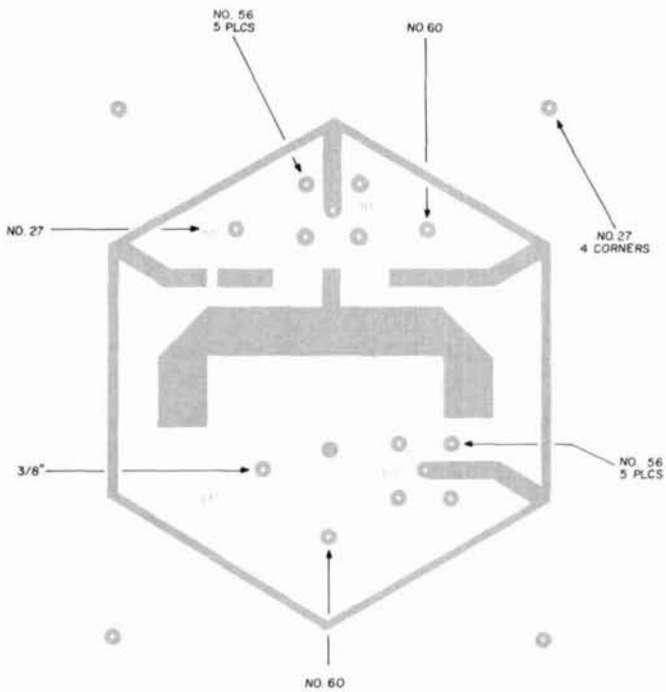


fig. 5. Drilling template for printed circuit board.

shortly), the bulk of this mixer design, including the i-f port matching network, is borrowed directly from Paul Wade's design[4] and will not be discussed here. I should point out, however, that the pi-network matching network at the i-f port does little to properly terminate the image component; the importance of this was discussed previously. For mixer applications which dictate maximum freedom from intermodulation distortion, the user should consider adapting the Bridge-T interstage isolator circuit.[14]

Fig. 4 is a full-scale layout for the rat-race mixer. The pattern should be etched on 1/16 inch (1.5mm) thick G-10 fiberglass-epoxy printed circuit laminate, double-clad with 1 ounce copper (1.4 mils or 36 microns thick). One side of the board is unetched and serves as a groundplane for the microstriplines on the etched side. Drill the board as shown in **fig. 5**.

Assemble the mixer as shown in **fig. 3** and the photographs. When mounting the SMA connectors at the LO and rf ports, be sure to countersink the groundplane side of the board at the connector center conductor to avoid a short to ground. Assembling the mixer is straightforward, except that the leads of the diodes must be bent carefully to prevent damage to the glass diode package.

Once assembled, only one mixer adjustment is required: matching to the i-f port. This can be done by injecting a signal into the rf port, connecting the i-f port to a 2-meter receiver, and tuning capacitor C6 for maximum signal level. Alternatively, with the rf port driving a power meter, inject a 3mW, 2-meter signal into the i-f port, and tune C6 for maximum output. In either case, the LO port must be driven with a clean 5 to 10 mW signal in the vicinity of 1152 MHz and the i-f port must be terminated in 50 ohms. The 50-ohm load can be provided by tuning the mixer with a fixed, 50-ohm pad inserted between the i-f port and the 2-meter transmitter or receiver.

Fig. 6 shows the isolation of this mixer, as measured on a Hewlett-Packard 8507A Automatic Network Analyzer. The isolation of the assembly is greater than 20 dB from 1050 MHz to over 1300 MHz, and at least 30 dB over the anticipated operating bandwidth; isolation over a 20% bandwidth is quite close to that predicted in **table 1**.

bias considerations

Wade operates his balanced mixers in the *starved-LO* mode.[3,4] That is, he applies a rather low level of LO injection (typically 1 mW) to the diodes, and supplements this low LO drive with dc bias. By varying the dc bias, the impedance of the diodes can be controlled, providing an improved match to the mixer hybrid. Obviously, it is far easier to generate the required dc bias than it is to increase LO injection by 10 dB.

Other advantages of DC mixer bias were reported by Pound as far back as 1948:[6] "Accompanying the reduced LO power requirement is a reduction in the reaction of the local-oscillator circuit on the signal circuit in the mixer. The over-all noise figure becomes less dependent upon the amount of incident local-oscillator power at the crystal, because the conversion loss does not increase so rapidly as the local-oscillator drive is decreased. Finally, the i-f conductance of the crystal is less dependent on the amount of local-oscillator drive."

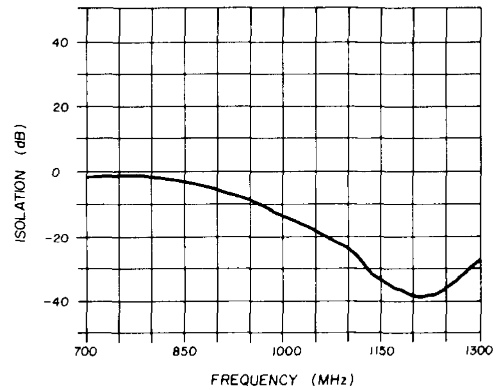


fig. 6. Isolation of the rat-race balanced mixer, as measured with a Hewlett-Packard 8507A Automatic Network Analyzer.

Starved-LO operation was attempted in one of my bilateral conversion systems, but with disastrous results. In transmit mode application of 1 mW of i-f injection resulted in a severely distorted rf output, measuring several dB below the power level which the characteristic conversion loss of the mixer would suggest. Further, a spectrum analyzer revealed numerous unwanted frequency components which were not previously a problem. It appeared that the dynamic range of a balanced mixer in the *starved-LO* mode was far below that of a similar mixer with the customary +3 dBm (2 mW) per diode injection level. This conclusion was confirmed by Harlan Howe:[8] "Naturally, intermodulation products and saturation levels will be the same as if the LO drive were unsupplemented by dc power." Thus *starved-LO* operation was abandoned for this mixer in favor of a full 4 to 10 mW of LO injection. A similar choice is indicated for any balanced mixer used in transmitter service, or any time large-signal handling capability is important.

It will be noted from the schematic in **fig. 3** that, although external dc bias is not used in this design, the diode's self-bias return is brought out through feedthrough capacitor C1. Since the LO signal is effectively rectified by the diodes, a dc voltage will appear across R1; the magnitude of the dc voltage is a function of applied LO injection level. Thus, the feedthrough capacitor is a convenient test point for measuring (and optimizing) LO injection, as indicated in **fig. 7**. The object is to minimize conversion loss, which is a direct function of LO injection (see **fig. 8**).

Like other microwave experimenters, in recent years I have sought to master the techniques of microstripline design and construction by developing individual circuits on etched circuit boards. Thus my

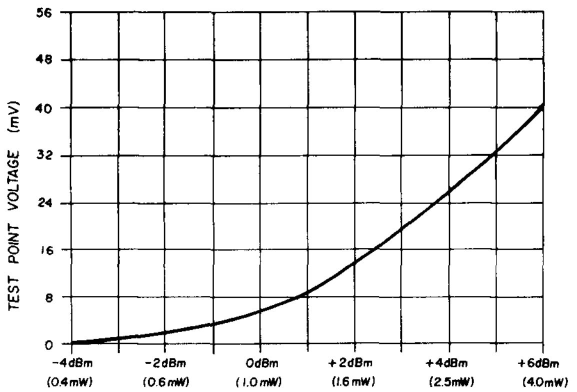


fig. 7. Since the diodes rectify the LO signal, a dc voltage measurement can be used to check LO injection level. Plotted here is the test-point voltage (mV) vs LO injection for the rat-race balanced mixer.

early 1296-MHz systems consisted of numerous modules, each containing a single stage, each matched to 50 ohms, and all inter-connected by coaxial cable. This approach at best is a crude utilization of microstripline technology. One of the main benefits to be derived from the etched-substrate approach is that numerous associated circuits may be built as an integral unit, thereby minimizing interconnections, reducing cost, and improving reliability.

I have made several attempts to integrate circuits onto common substrates. My process of system integration began with a two-stage preamplifier,[12] progressed to a combined balanced mixer, image filter, and LO filter,[2] and culminated in a complete transverter-on-a-board.[13]

Unfortunately, acceptance of these integrated assemblies by amateur microwave enthusiasts has not been overwhelming. It was brought to my attention by numerous readers that the tasks of tuneup and testing present major problems when multiple, interactive stages are involved, especially when the available test equipment is limited to a grid-dip oscillator and a number 47 lightbulb! Individual stages, on the other hand, may be readily tested into 50 ohms, and then placed into the desired system with few adjustments. Joe Reisert has been promoting such a modular approach for some time; his articles are highly recommended.[15,16,17]

*Not to be confused with RG-141/U, which is a ¼ inch (6.5mm) diameter flexible cable.

My 1296-MHz systems have recently gone through a process of de-integration (not to be confused with *dis*integration) with a return to small modules, each containing a single stage, all stages interconnected with 50-ohm coax.[18] This mixer is one such module. Unlike my previous mixer, the one presented here contains no rf or LO filters. Filters will of course be required in most applications but are easily added to the system as individual modules after they have been tuned and tested.[19]

A word about rf cabling and connectors is in order. Whenever possible I recommend the use of high quality microwave connectors such as type SMA. The JCM series of coaxial connectors from E.F. Johnson are low cost, SMA-compatible units exhibiting excellent rf properties through 4 GHz. For microstripline launchers I recommend E.F. Johnson part number 142-0298-001, and I use their 142-0161-001 connectors on all jumper cables. These connectors are priced in the $3.00 range and are available from electronics distributors in many parts of the United States.

Of all the considerations surrounding the selection of interconnecting coaxial cable, cost and availability are the major factors for amateur applications. Obviously, low loss and constant impedance must also be considered. Semi-rigid coax, such as Uniform Tubing UT-141,* perform exceptionally well, but are virtually unobtainable in many parts of the world. A good second choice is ¼ inch (6.5mm) diameter flexible coax, if the lengths are short. Desirable features are double shielding, Teflon dielectric, and a silver-plated, solid center conductor. One cable meeting these requirements is RG-142/U, but its cost per foot discourages many experimenters. Nonetheless, a moderate investment in jumper coax can spare you a world of grief. Other usable cables which cost considerably less include types RG-141/U, RG-223/U, and RG-55/U. As a last resort, RG-58/U may be used, but its length must be kept to an absolute

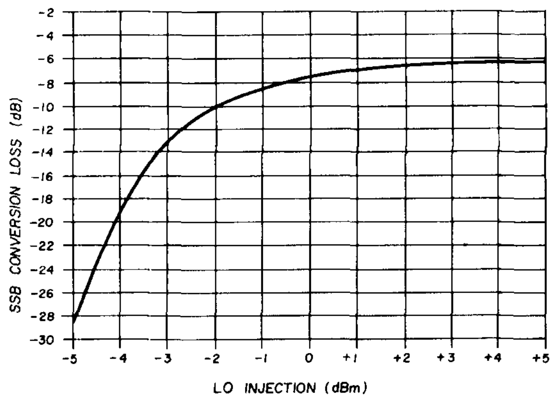


fig. 8. Conversion loss of the rat-race single-balanced mixer as a function of local-oscillator injection.

minimum. Just remember that most 1296 operators use RG-58/U to build calibrated attenuators! All of the flexible cable types listed here will accept the recommended SMA plug.

parts availability

The components required to build the rat-race mixer described in this article are available from various sources in the United States; the printed-circuit board may be etched from the full-size artwork shown in **fig. 4**. For those uhf experimenters who don't have the facilities to etch their own boards, commercially etched, drilled, and plated boards are available from Microcomm.*

*An etched, drilled, and plated circuit board for the rat-race mixer is available from Microcomm, 14908 Sandy Lane, San Jose, California 95124, for $6.50 post paid within the United States and Canada ($7.00 elsewhere). A wired and tested mixer is also available; Microcomm will *not* offer a complete kit of parts.

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More Details? CHECK—OFF Page 126

self-supporting coils

A set of inductance and *Q* graphs for airwound coils using screws and other threaded forms as mandrels

Self-supporting coils without forms are the cheapest, most convenient, high-*Q* inductors for inductance values under 0.4 microhenries. Unfortunately, the wire is difficult to wind on a smooth form and you have to account for wire diameter plus form diameter for accurate calculation. Presented here is a method for using conventional screw threads as forms along with measured inductance values.

Form and wire size, coded by the letters of **fig. 1**, are given below:

curve	wire size	winding form	
A	26 enameled or Soldereze	4-40	screw
B	22 enameled or Soldereze	6-32	screw
C	22 enameled or Soldereze	8-32	screw
D	22 enameled or Soldereze	10-32	screw
E	18 tinned or enameled	1/4-20	screw
F	18 tinned or enameled	5/16-18	bolt
G	18 tinned or enameled	3/8-16	lag bolt
H	14 bare copper	7-watt Christmas bulb base (3/8″ ID, 10 TPI)	
J	12 bare copper	Paint roller ferrule (5/8″ ID, 1 turn per 3/16″)	
K	12 bare copper	Standard 117 Vac lamp base	

The method is very simple: Take any convenient screw or bolt and wind the wire firmly on the threads. When finished, allow the wire to release its tension, then carefully remove the form. **Fig. 1** shows inductance vs the number of turns for ten different screw threads on easily obtained sizes.

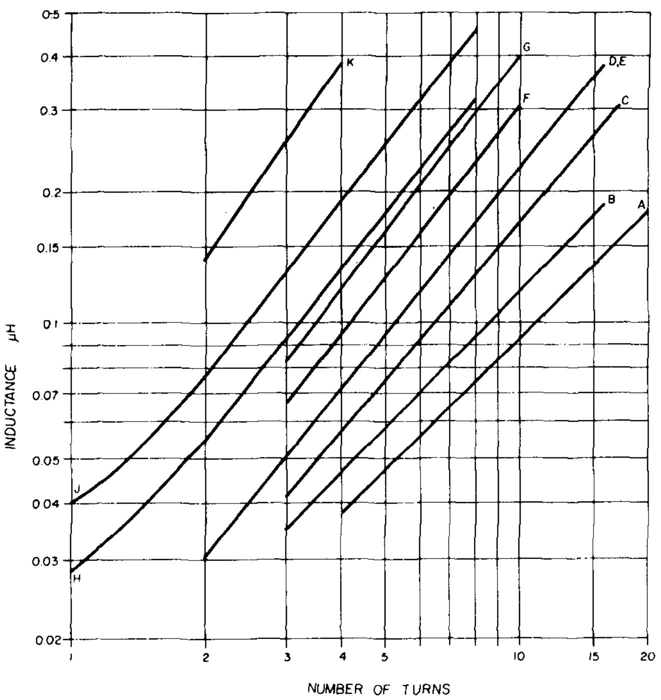

fig. 1. Inductance of airwound coils using threaded forms as mandrels. Extensive use of this chart has shown the inductance values to be reproducible to within about 5%. Inductor Q for each of the 10 forms is shown in figs. 2 through 11.

This data was originally collected by the author in 1969. It has been used since then in many commercial applications, and the values checked with other Q-meters. In all cases the inductance values were within 5% of the predicted value; *Q* was up to 20% greater, depending on the measuring instrument. **Editor.**

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

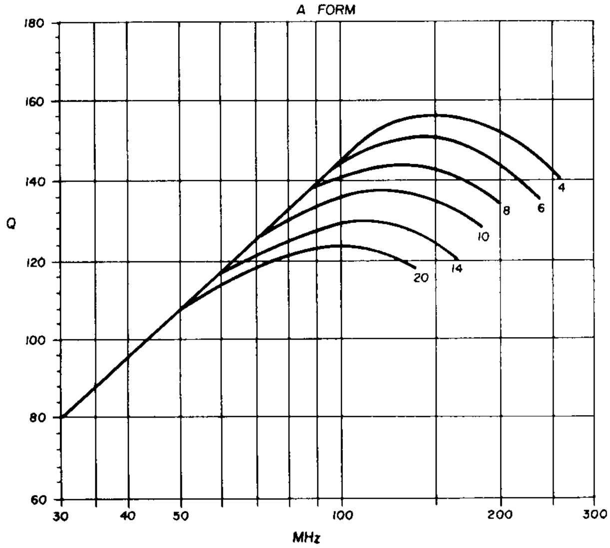

fig. 2. Q of inductors of no. 26 enameled wire wound on 4-40 screw form (curve A in fig. 1).

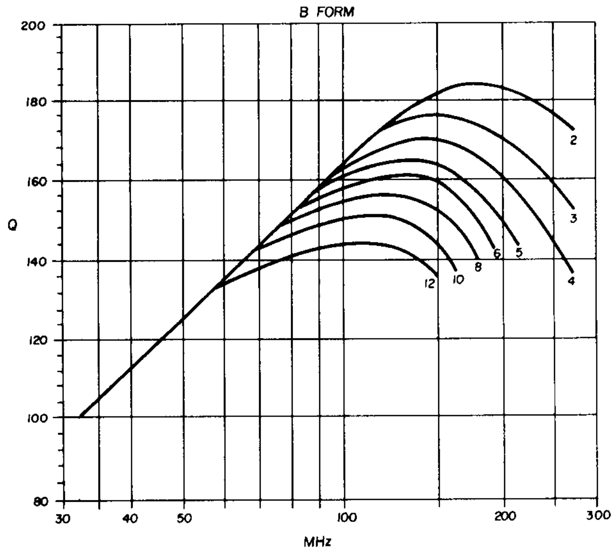

fig. 3. Q of inductors of no. 22 enameled wire wound on 6-32 screw form (curve B in fig. 1).

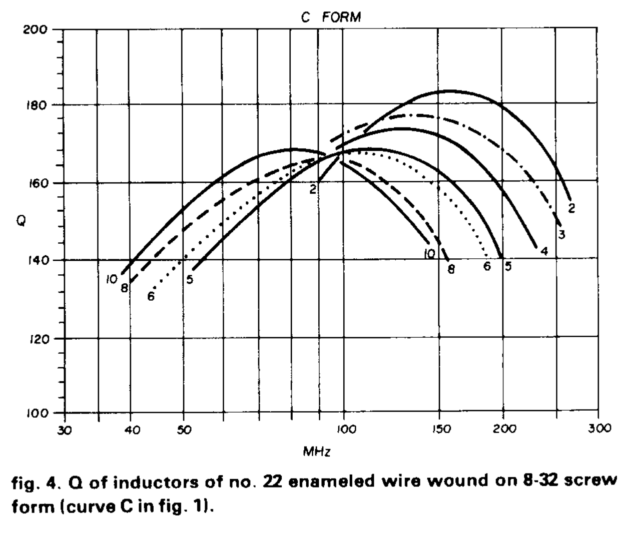

fig. 4. Q of inductors of no. 22 enameled wire wound on 8-32 screw form (curve C in fig. 1).

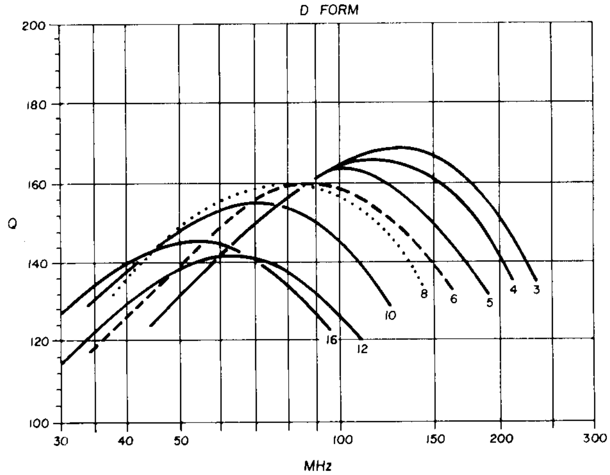

fig. 5. Q of inductors of no. 22 enameled wire wound on 10-32 screw form (curve D in fig. 1).

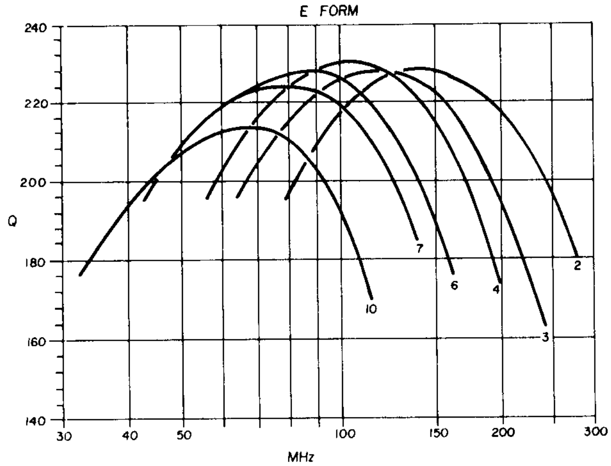

fig. 6. Q of inductors of no. 18 tinned or enameled wire wound on 1/4-20 screw form (curve E in fig. 1).

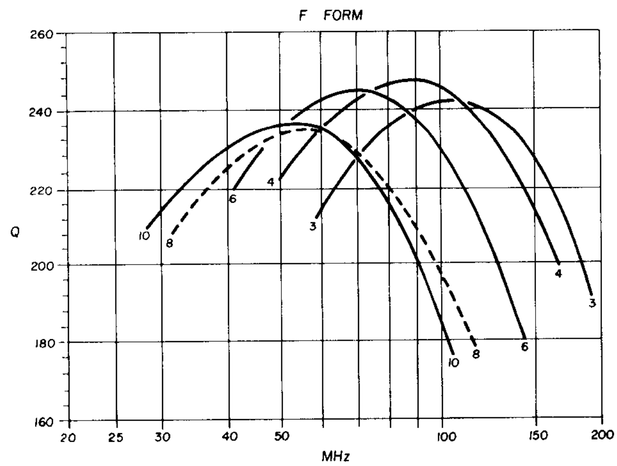

fig. 7. Q of inductors of no. 18 tinned or enameled wire wound on 5/16-18 bolt (curve F in fig. 1).

Wire size is fixed by the number of turns per inch on the screw form. This has been selected to allow a slight space between turns so that adjustment for higher inductance by squeezing may be done if

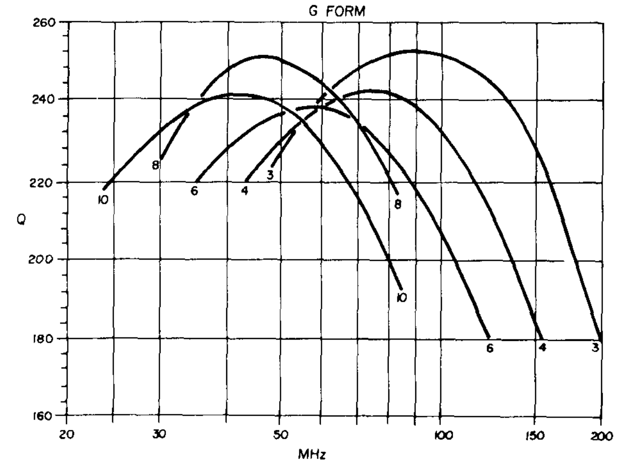

fig. 8. Q of inductors of no. 18 tinned or enameled wire wound on 3/8-16 lag bolt (curve G in fig. 1).

desired. It was also found that wire diameter had to to be restricted to just under full winding by the screw threads.

The no. 12 and 14 AWG wires are household power wires (*Romex*) with the insulation stripped off. If this material is not available, check for wire scraps at industrial plants or at new construction sites; quite a bit is thrown away. The paint-roller ferrule can most often be found on extender poles sold for that purpose and are quite uniform in dimension.

The *Q* curves are shown in **figs. 2** to **11.** Values in between the indicated number of turns may be interpolated with reasonable accuracy.

All data was obtained by construction and

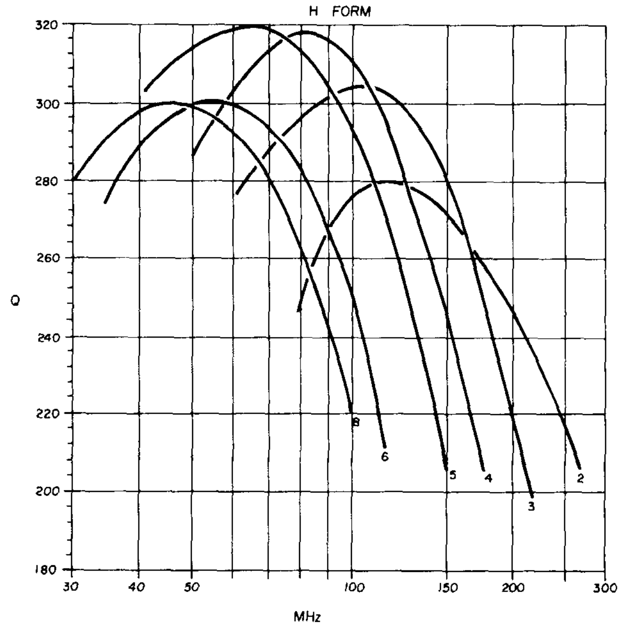

fig. 9. Q of inductors of no. 14 bare copper wire wound on 7-watt Christmas bulb base (curve H in fig. 1).

measurement with a Boonton 190A Q-meter. This limited the measurements to the 20 to 240 MHz instrument range. No compensation for lead length has been made and all coils have an assumed 3/4-inch connection spacing. Each coil was kept at least one diameter away from the top surface of the Q-Meter, consistent with shortest lead length.

Reproducibility of inductance should be within 5% and Q within 20%. This assumes standard coil wire tolerances and lead lengths given. Tolerance may

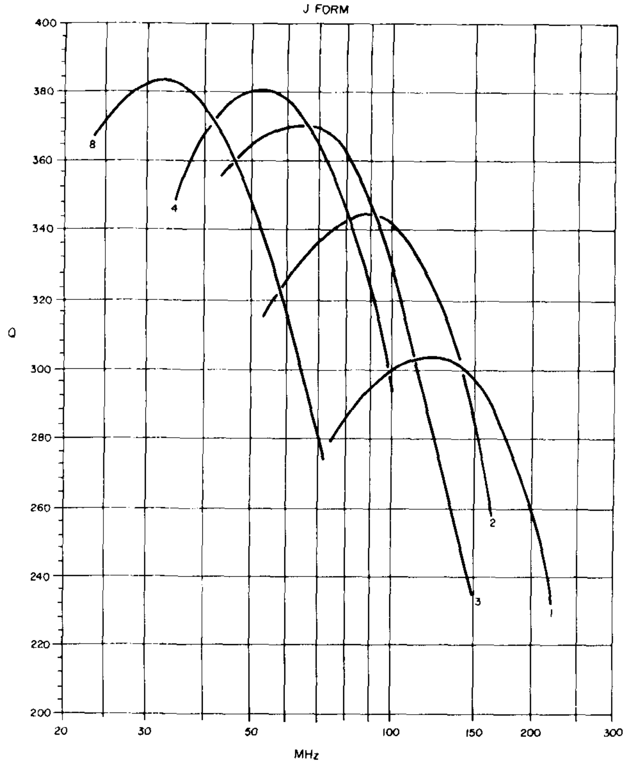

fig. 10. Q of inductors of no. 12 bare copper wire wound on paint roller ferrule (curve J in fig. 1).

drop to 10% for inductors using stripped power wires, depending on the brand; sampling wire from different manufacturers showed 5% tolerances were possible.

Where long leads are necessary, inductance can be adjusted by the following formula for straight wires:

$$L = l[11.7 \log_{10} (0.12 \cdot l \cdot g^2)]\, nH$$

Where: l = *length of wire (inches)*

g = *wire gauge (AWG)*

The formula is an approximation but accurate to 5% for AWG wire sizes from 12 through 26.

All constructions are quite stable and will hold up under most conditions encountered in amateur use. Like all self-supporting inductors made of soft wire, they are flexible and should not be used in vfo tank circuits or other critical applications.

Bare copper may be coated with varnish to retard oxidation. A light application of spar varnish or polyurethane varnish will lower *Q* by only 5 to 8 per cent. **Note**: Do note use *Q Dope* since, like all lacquers and acrylics, it will transmit moisture and lift from the non-porous surfaces.

Bare copper holds up surprisingly well. To prove a point, I wound an 8-turn coil on a paint roller ferrule (form **J**) and buried it in an outdoor planter along with two new plants. The coil was compressed to

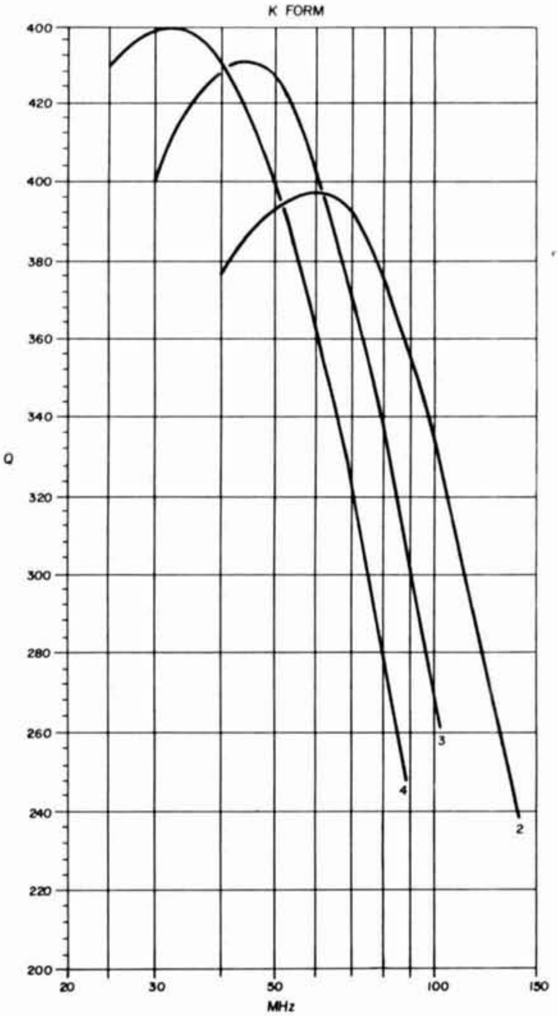

fig. 11. Q of inductors of no. 12 bare copper wire wound on 117 Vac household lamp base (curve K in fig. 1).

about 60% of finish length and readings were taken at 50 MHz prior to planting and four months later. Despite watering every other day, the untreated coil had only a 2.3% reduction in inductance and 16% drop in *Q*.

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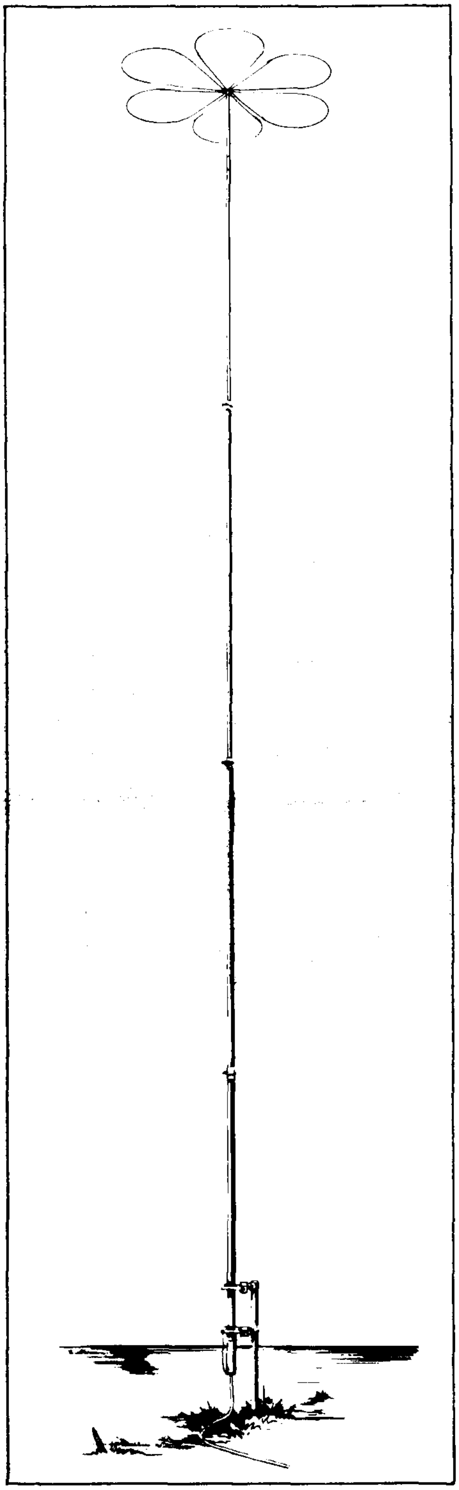

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gain control IC for audio signal processing

As a multi-purposed IC, the NE570 analog compandor fulfills many audio processing needs

Two ICs recently introduced by Signetics, the NE570 and NE571, permit the design of efficient and practical audio-signal control functions with a minimum overall parts count. These devices are primarily designed to act as compandors; the complementary processes of compression and expansion.[1,2] They are both dual-channel ICs and either portion can be used individually as a compandor. However, as will be seen in this article they are also well suited to a variety of other tasks useful to the amateur.

basic device operation

Each channel of the 570 and 571 consists of the functional components shown in **fig. 1A**. Packaged in a 16-pin DIP, the only items common to the two signal channels are the power supply, ground connections, and an internal 1.8 volt bias regulator.

The three principal components of each section are a ΔG cell, full-wave rectifier, and an output amplifier. The ΔG cell is used to control the gain over a range greater than 80 dB. The control voltage for this cell is generated by rectifying an input signal (RECT IN). The final output is then developed by the buffered output amplifier from the scaled signal current supplied by the ΔG cell. The 570 and 571 are identical electrically, but the 570 is selected for lower inherent distortion and a higher supply voltage range.

The ΔG cell, as shown in **fig. 1B**, consists of an op amp, A1, and transistor pairs Q1-Q2 and Q3-Q4. The input signal is first converted by R2 into a current that drives A1. The feedback for this op amp is via the transistor pair Q1-Q2. Therefore, the amount of current in this pair is the same as the current through R2. In addition to driving Q1-Q2, the op amp is also connected to Q3-Q4. Unlike Q1-Q2, this transistor pair does not have a constant-current source. By scaling the Q3-Q4 emitter current, their output is a linear product of the input signal from A1 and the scaled current. This circuit is a linearized transconductance multiplier[4-7] which cancels the inherent non-linearity and temperature sensitivity of the differential pairs, greatly enhancing the usefulness of this gain-control technique.

The rectifier portion consists of op amp, A2, class-B transistors Q5-Q6, a pnp current mirror Q7, and an npn current mirror Q9. When rectifying a signal at the RECT IN terminal, Q5 and Q6 produce pulses of current proportional to the positive and negative input signal swings. The output current of Q6 is used directly, while the Q5 current is mirrored by Q7. Thus, the drive to Q9 is a positive going, full-wave rectified pulsating dc. These pulses are filtered by an external smoothing capacitor attached to the C_{RECT} terminal.

The output stage is a simple inverting op amp similar in performance to a 741. Various options are possible by use of either R3, external input, or feedback resistors. The overall circuit gain is unity (ΔG IN to OUT), with R3 connected as a feedback resistor and 70 μA rectifier current into Q9.

In addition, the THD TRIM terminal allows a small offset to be introduced into the ΔG cell to null its distortion. The two input op amps (A1 and A2) are connected to the internal 1.8 volt regulator. Each op-amp input should be capacitively coupled while the input impedance is determined by R2 or R1, respectively. Circuit operation is very stable and immune to power-supply variations. A single supply voltage from +6 to +18 volts (571) or +6 to +24 volts (570) can be used, though the following applications will use a +15 volt supply.

basic compandor circuits

The 570 and 571 can be quite simply connected for their basic functions of expansion and compression, as illustrated in **fig. 2**. These circuits will not be dealt with in great detail because most amateurs will probably be more interested in some of the other uses. Also, compandor operation is covered in detail in other literature.[1-3]

The gain through the expandor shown in **fig. 2A** is $1.43\ V_{IN}$, where V_{IN} is the *average* input voltage.

By Walter G. Jung, 1946 Pleasantville Road, Forest Hill, Maryland 21050

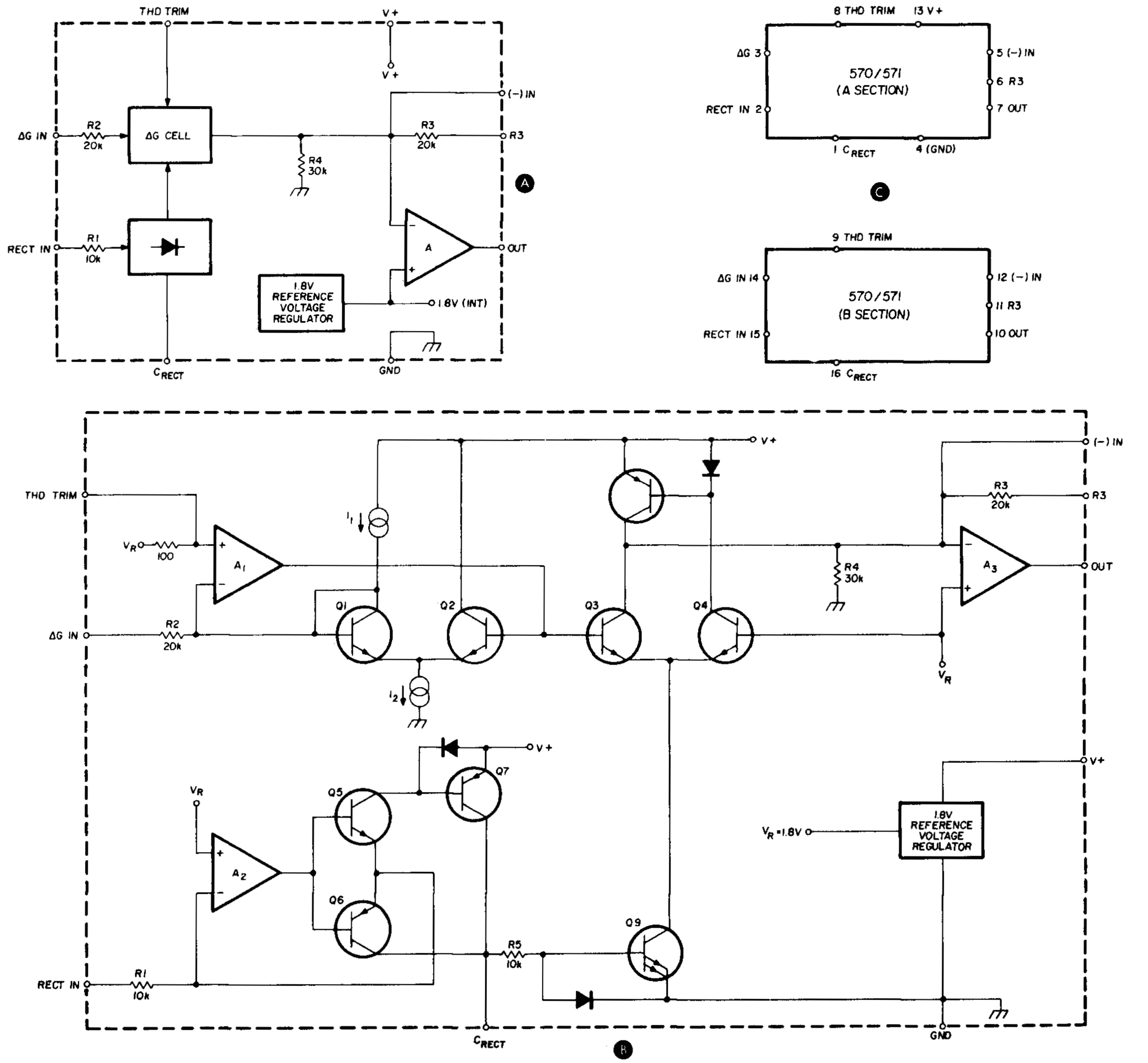


fig. 1. Functional diagram of the Signetics NE 570/571. A simplified schematic diagram of the device is shown in B. Constant current sources I_1 and I_2 feed transistor pair Q1-Q2.

The 570/571 circuit constants are set up such that unity gain occurs at an rms input level of 0.775 volts, or 0 dBm in 600-ohm systems. The C_{IN} and C_O are coupling capacitors, chosen for the desired low-frequency rolloff. C_{RECT} is selected for the desired time constant (10 ms) in conjunction with the internal 10-kilohm resistor (R5).

Resistors R_A, R_B, R_C and C_B are not essential to basic operation, but are desirable. R_B furnishes short-circuit protection for the output and capacitive load buffering, while R_A and R_C polarize C_{IN} and C_A. C_B is a power supply bypass, typically an aluminum electrolytic.

The compressor configuration in **fig. 2B** also has unity gain at 0.775 volt (rms) input, but, a complementary in/out characteristic. The main difference in this circuit is that the ΔG cell is connected as a feedback impedance via C_F, and the input is applied to R3 through C_{IN}. Bias for the output stage is set up by the RC-decoupling network, with the values shown appropriate for 15-volt power supply.

In general, the OUT terminal should be biased to one-half the supply voltage. Use of a 570 or 571 as a compandor is not limited to the gains shown, but may be extended to other ranges by use of additional components.

trimming techniques

Device performance can be enhanced by judicious trimming, as shown in **fig. 3**. Each technique is optional, and can be applied in any combination when the highest performance is desired. The most useful of the three methods is probably the THD trim, which minimizes the gain cell harmonic distortion. In this case, a small voltage (0 to 3 volts) is used to inject a current into the THD TRIM terminal through the 100k resistor. By biasing the rectifier terminal as shown, the inherent current flow in the rectifier is compensated for and permits better low-level signal tracking. Typically, the gain-control signal to A1 should not be reflected in the output. The control feedthrough trimmer will minimize that signal during periods of low input voltage.

applications

An interesting and versatile group of circuits, the *gated or switched-mode amplifier, can be built from* the 570/571. With the device controlled by external logic applied to the RECT IN input, the *on* gain is normally set to any value and the *off* attenuation can be in excess of 80 dB. Use of the 570 or 571 is advantageous in that all portions of the function can be performed entirely within the IC. Further, the on/off transition times can be set to a value determined by *the time constant from* C_{RECT}.

Fig. 4A is a logic controlled amplifier configured for a HIGH input to be on, and LOW off. When the control input is HIGH, CR1 is off and the current

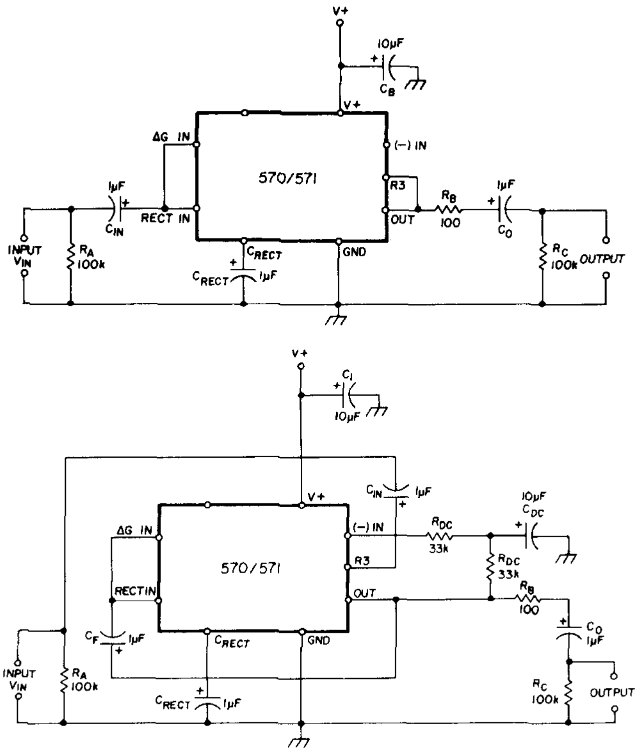


fig. 2. Schematic diagram of the devices connected as an expander, A, and a compressor, B. The voltage gain through the expander is $1.43 \cdot V_{IN}$, while for the compressor it is $\sqrt{0.7/V_{IN}}$. V_{IN} is the average input voltage.

developed by R_{GAIN} *flows into the rectifier input,* which turns on the ΔG cell allowing the signal to be amplified. R_{GAIN} can be selected for the desired *on* state gain, which is unity with a rectifier current of 70 μA. R1 and R3 also effect device gain, but R3 is selected basically for an optimum output bias of 7.5 Vdc. R1 can also be adjusted for gain, but as shown the value allows up to 3 volts rms input/output signal levels.

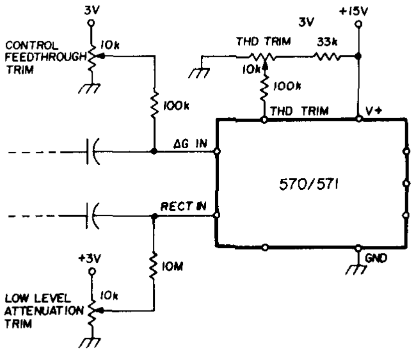


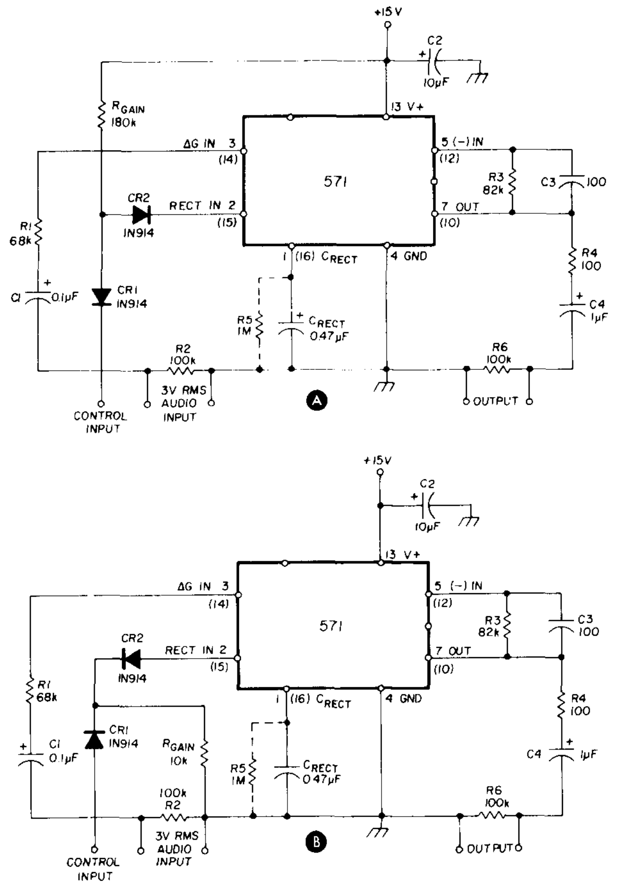
fig. 3. By applying the different trimming methods, distortion through the 570/571 can be reduced. Though each method is optional, they can be applied in any combination.

As can be seen in the control characteristics plotted in **fig. 4C**, the gain is unity (or its nominal value, if chosen otherwise) for control inputs greater than 3 volts. Switching is quite abrupt, with full attenuation being achieved at levels less than 1.5 volts. This narrow transition width and the nominal dc center of 1.8 volts allows direct control from CMOS, TTL, DTL, or other positive logic. The ultimate voltage of the HIGH state is non-critical, due to the 100-volt rating of CR1. Unfortunately, this circuit has one inherent weak point. Gain is sensitive to supply voltage due to the connection of R_{GAIN}. Thus, the supply voltage should be stable while choosing R_{GAIN} for 70 μA into the RECT IN terminal.

A companion circuit with complementary control characteristics is shown in **fig. 4B**. In this case, the gain is determined by the current developed through R_{GAIN} in conjunction with the internal voltage reference (1.8 V). With a low control input, the normal current will flow out through R_{GAIN}. When the control signal is high, CR1 is forward biased, interrupting the current flow. Therefore, the output will be attenuated since Q3-Q4 have been turned off.

Both circuits can be tailored for specific on-off transition times by selection of C_{RECT}. The time constant is simply $10k \cdot C_{RECT}$ (10 kilohms is the internal resistor). Thus, the audible switching effect can be smoothed, eliminating the transients produced by an asynchronous fast switch. The C_{RECT} value shown yields nominal times of 5 milliseconds.

Use of C_{RECT} in a switched amplifier of this type is



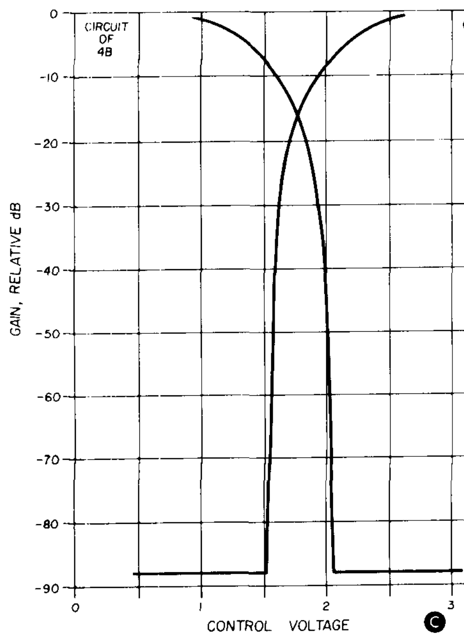


fig. 4. The logic controlled amplifiers can be configured to provide an output with either a HIGH or LOW input. This allows the amplifiers to be interfaced with many different logic types, TTL, CMOS, DTL, etc. The resistor R_{GAIN} should be selected to provide 70 μA at unity gain. Each amplifier has an on/off time determined by the time constant of C_{RECT} with internal 10k resistor.

optional and not absolutely necessary. However, to minimize noise pickup some capacitance will be found useful. Also, the ultimate *off* state attenuation will be limited to about 60 dB due to the internal bias current. This *feedthrough* error can be eliminated by connecting a 1 megohm resistor from C_{RECT} to ground to bleed away the error current. This allows attenuation of 80 dB or more.

Fig. 5 illustrates two sections of a 571 combined as a two-input multiplexer, for FSK or other uses. This circuit operation is similar to the others, but is biased and switched in a simpler manner. Gain of each *on* channel is unity, as determined by R_{GAIN}. The output of the B channel ΔG cell is summed with channel A by connecting the (−) IN terminals of the A and B sections. The respective channels are gated *off* by a low control logic input, which clamps the rectifier current, switching the ΔG cell off. For fsk or alternate channel use, the CONTROL A and CONTROL B signals should be complementary. Thus, the input is "instantaneously" switched between the **A** and **B** inputs.

Control signal suppression can be optimized with the CHOPPER NULL control, which trims the control signal component in the output. Suppression is better than 60 dB after trimming. Response time is quite fast, and is actually limited by the slew rate of the output op amp rather than the ΔG cell itself. This makes the switching interval a function of the signal's peak amplitude. For instance with a 4-volt peak amplitude signal the 0.5 V/μs slew rate will

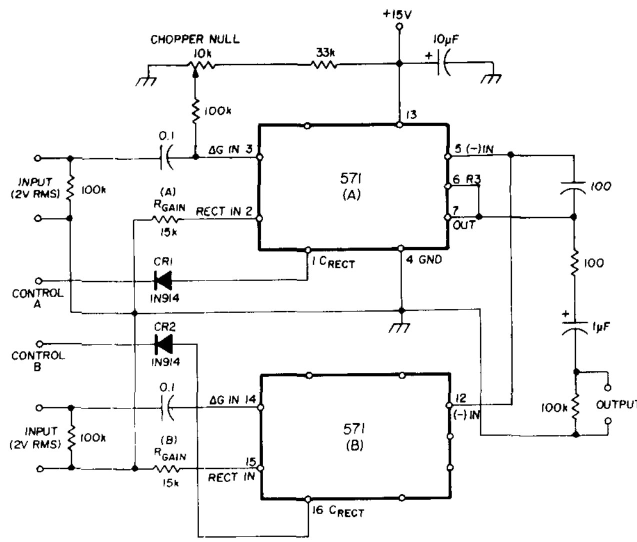


fig. 5. By providing complementary control signals, the FSK generator will switch between the two signal inputs. The outputs, when ON, are summed through the first operational amplifier.

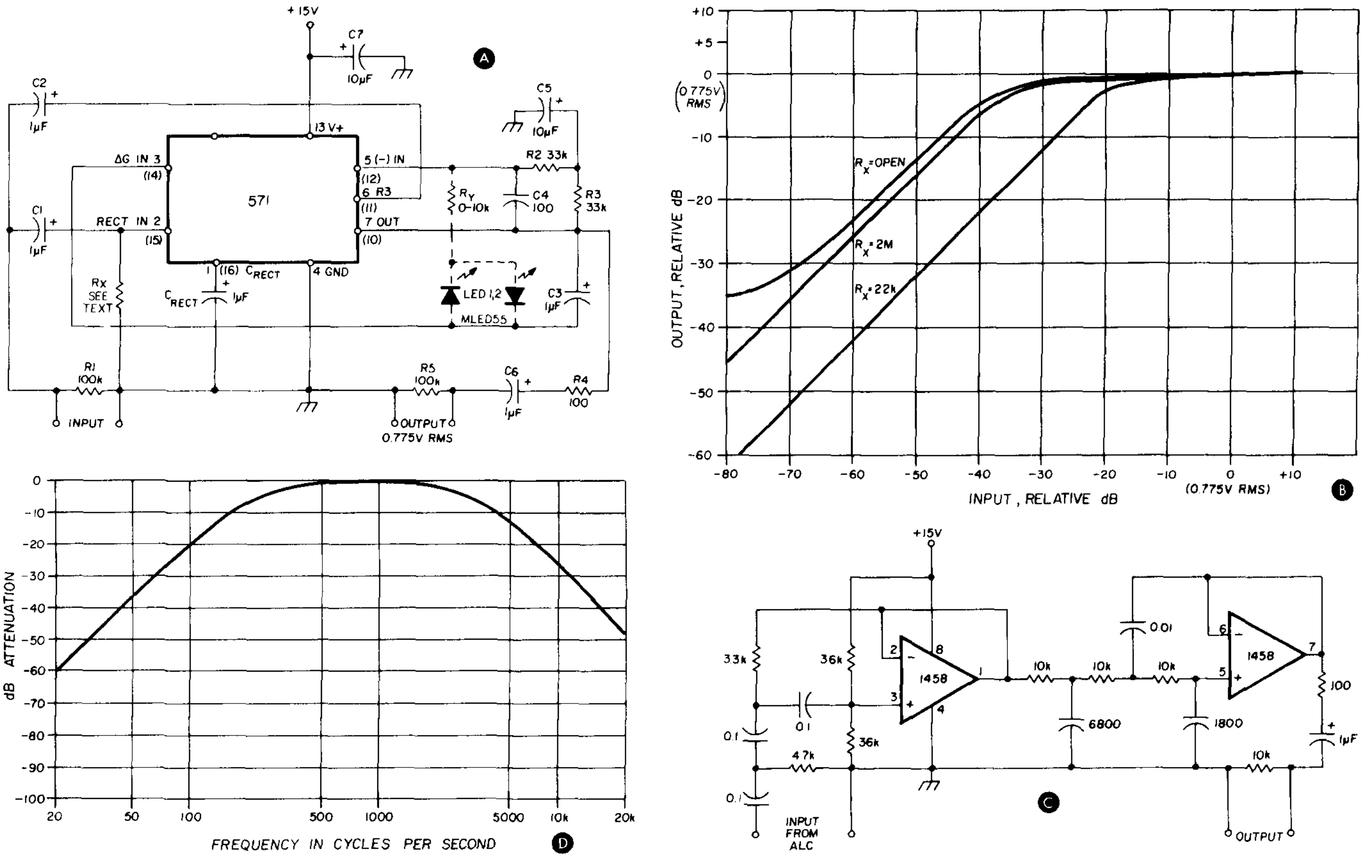


fig. 6. Schematic diagram of an NE570 connected as an automatic level control (A). The speech filter (C) is added to remove the undesired components created by the peak clipping. The attenuation curve (D), though not sharp, provides excellent rejection of the sharp pulse spikes generated by the clipping.

allow switching in 8 μs; lower amplitudes will be proportionally faster. Although the circuit is touted as a multiplexer, it can also be used as a summing switch, with both signals on at any given instant.

automatic level control

Automatic level control is a relatively common requirement in audio signal processing. A 571 automatic level control circuit can provide constant, high percentage modulation with varying input levels, yet without danger of overload if properly handled.[8] This circuit (**fig. 6**) is adapted from the 570/571 data sheet. There is one additional feature which may be useful, however: an optional resistor allows the threshold of level regulation to be varied. If R_X is left open, the circuit will have its widest range of gain control. As this resistor value is lowered, a larger input signal is required for full output. The general effect, for various R_X values, is shown in **fig. 6B**.

Since the 570/571 device operates on the principles of *average* level detection, it will easily saturate on large crest factor input levels such as speech, or large transient envelope changes. This condition is highly undesirable, but, fortunately its effect can be negated quite simply. Since the result of overshoots are peak-to-peak amplitudes in excess of the regulated output level, it follows that appropriate peak-level clipping can effectively control the overshoots. In this case, the rms output amplitude is 0.775 volt or 2.2 volts p-p. This particular level is conveniently clipped with a pair of reverse paralleled LEDs, which will limit to 3.2 volts p-p. The LEDs can be connected as shown in **fig. 6** with a series resistor R_Y, which is used to regulate the clipped amplitude.

This clipping technique, while not a requisite part of the automatic level control, greatly enhances its regulation with nonsinusoidal signals. Use of this circuit with speech inputs will necessitate the diodes, which will typically be clipping a good portion of the time. This, of course, adds audible distortion to the output. Therefore, a useful item with the ALC/clipper is a speech filter to remove the superfluous high and low frequencies. The filter will also greatly attenuate harmonic components generated by clipping.

The circuit in **fig. 6C** is a bandpass speech filter which uses a single IC. The circuit is simply a pair of cascaded Sallen and Key[10] highpass filters (3-pole Bessel type). The Bessel response is one of the poorest in terms of cutoff sharpness, but good from a pulse response standpoint. This feature is important to minimize amplitude overshoots which could occur with severely clipped inputs.

A common 1458 (dual 741) op amp is used with nearest 5 per cent component values for the filter elements. If low-power operation is desired, the 1458 can be replaced directly with a 358. If a 358 is used, 10 kilohm resistors should be added from each output terminal to common. With unity gain, the circuit can drive load impedances greater than 10 kilohms.

One very effective use for the 570 and 571 device is an amplitude-regulated RC sine wave oscillator. Typically, such circuits use a Wien bridge or other frequency-selective RC network, with some form of amplitude stabilization to maintain constant and correct loop gain, and also to guarantee output waveform purity. A 570 or 571 is nearly optimum for this type of circuit because it contains the required functions of amplifier, rectifier, and gain-control circuits.

Two types of sine wave oscillators are shown in **fig. 7**. The oscillator circuit (**fig. 7A**) based on the Wien network is formed by the combination of R1-C1 and R2-C2. This network is placed around the output amplifier of section A, which effectively makes it a bandpass amplifier resonant at

$$f = \frac{1}{2\pi RC}$$

With equal values of R and C, the input/output voltage ratio is 2 to 1.

To originate and sustain oscillations, the 571B section is used as an inverting amplifier with a nominal gain of 2. A slightly greater initial gain is established by the combination of R6, R7, and R8, which ensures startup. The B section ΔG cell is connected as a compressor, which regulates this stage's gain at the precise value required to maintain undistorted, stable amplitude oscillations.

There are two main steps taken to enhance flexibility of the circuit. A separate dc feedback path (R3, R4, C4) is used around the A stage, to remove value restrictions on R2 due to bias considerations. This allows R2 (R1) to range from 10k to 1 megohm without a major performance compromise. C1 and C2 have an even greater range, from 1 μF down to 100 pF. To minimize error due to strays, the lowest value should be used. With the values shown, the circuit is capable of reasonably low harmonic distortion. For example, 0.03 per cent distortion was measured at 1.6 kHz and THD (Total Harmonic Distortion) can generally be held below 0.1 per cent. This will vary according to the specific frequency, and the selected impedance of the Wien network. The low value of distortion is due to the light degree of ΔG cell regulation.

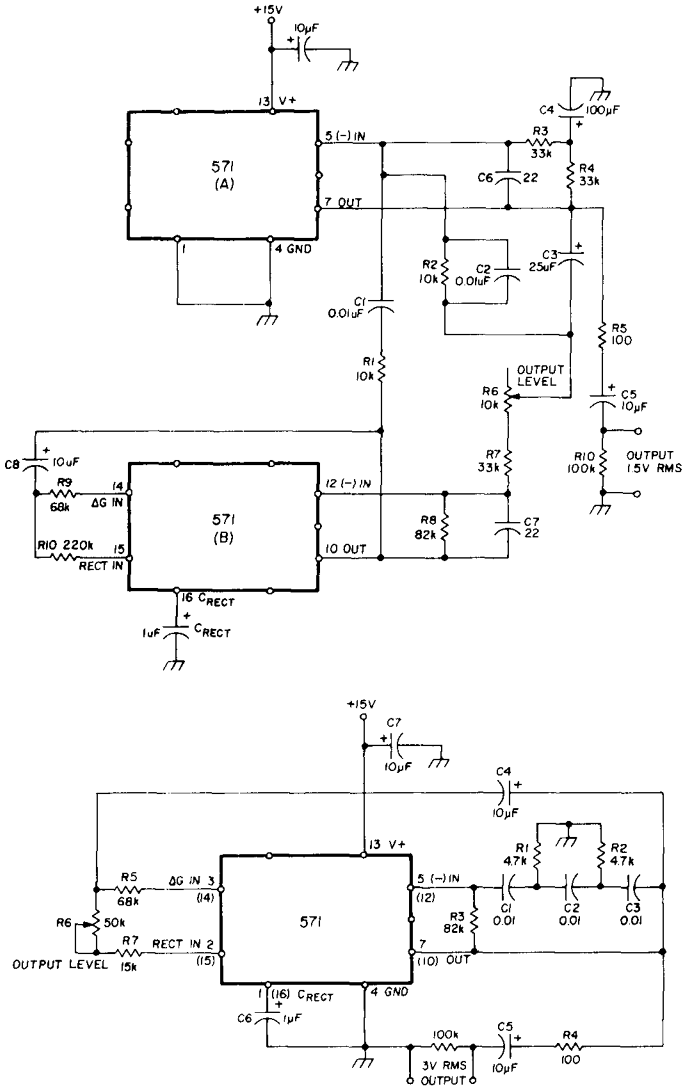


fig. 7. The NE570/571 can be connected as a sine-wave oscillator. The Wien bridge type oscillator is shown in A. For R = R1 = R2 and C = C1 = C2, the operating frequency is 1/2πRC. Resistor R should be limited between 10k and 1 megohm with C between 1000 pF and 1μF. The normal frequency range can be varied from 10 Hz to 10 kHz. The phase-shift oscillator should be used to generate discrete frequencies only. Depending upon the selection of parts, the output frequency will be 1/2πRC√3.

The circuit will operate as shown over the range from 10 Hz to 10 kHz. Below 10 Hz component size becomes impractical, and above 10 kHz slew limiting in the output amplifier causes distortion to rise. The circuit is useful as a fixed frequency oscillator, but can also be tuned if a matched dual pot is available for R1-R2. Output amplitude is set by R6, and is optimum at 1.5 volts rms output, from section A. If a higher output level is needed, section B output can also be used, at 3 volts.

The circuit of **fig. 7A** may be unduly complex for some uses, so an alternate and much simpler sinusoidal oscillator is shown in **fig. 7B**. This circuit is a form of phase-shift oscillator, similar to that described by Tobey, Graeme, and Huelsman.[9] A 571 is well suited for a phase-shift oscillator because it contains the necessary inverting amplifier to sustain oscilla tion. In the circuit shown, C1, C2, and C3 are the timing capacitors, while R1 and R2 are the resistors for the phase-shift network. R3 must be at least 12 times the R1-R2 value for adequate loop gain. AGC is provided by using the ΔG cell as a compressor.

This circuit is not suitable for tunable use. It should only be used as a spot frequency oscillator, by varying C1, C2 and C3. This is because R1 and R2 are related, by the design, to R3; in this specific case R3 cannot be variable because it is used to set the output dc bias point.

Although it uses a simple design, this circuit produces excellent results. At the frequency indicated, a laboratory test indicated a THD of 0.01 per cent at 3 volts output, which is remarkable in view of the circuit's simplicity. To take full advantage of this performance, an output buffer may be useful; for this you could simply use the remaining channel as a simple unity gain inverter.

conclusions

This discusssion has covered a few uses for a new and interesting chip. In the course of this article's preparation several other potential uses suggested themselves, such as phase comparators, phase-locked loops, voltage-tuned oscillators, and others. Unfortunately, space and time restrictions did not permit their complete examination.

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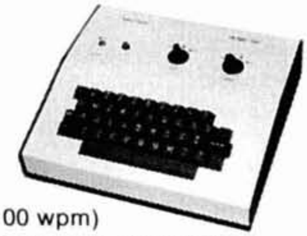

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high dynamic range two-meter converter

Circuit details for a 2-meter converter with a +15 dBM intercept point and 5 dB noise figure

This two-meter converter is an improved version of an earlier model[1] and is a result of research and development which I did around 1969 at AEG-Telefunken in Ulm, Germany. The German Ministry of Postal Affairs (equivalent to the U. S. FCC) requires that for all approved receivers for commercial and military application, a dynamic adjacent channel measurement must be performed. The result of the test depends upon the third-order intermodulation-distortion characteristic of the input stages.

A Radio Amateur group has used this particular converter circuitry which was later referred to as the "Martin front end."[2] This technique is also successfully used in the Rohde & Schwarz high-frequency Communication Receiver EK47, and was later adopted by Southcom and Atlas. The basic purpose of this circuit is to get the best performance out of a mixer by obtaining the lowest possible noise figure. To achieve this, the i-f output circuit has to be properly terminated over a large frequency range. This feature was previously described in *ham radio*.[2] During the last few months, the third-order intermodulation distortion has been evaluated in several magazines. However, a so-called *second-order* intermodulation distortion problem has received little mention.

A two-tone test is used for both second- and third-order IMD performance measurements. Second-order performance is checked at $f_1 \pm f_2$. Third-order is the performance at $f_1 \pm 2f_2$.[5] To reduce the effects of second-order IMD, it is necessary to use as much selectivity as possible and then compensate for the losses of these filters by using appropriate amplifiers. A suitable low-noise preamplifier with wide-band matching will be discussed later.

The two-meter converter shown in **fig. 1** combines these techniques. It was designed for extreme linearity and selectivity with the added goal of keeping the noise figure below 5 dB.

Many converters with noise figures below 3 dB described in the literature use neutralization. This method has two distinct disadvantages in the way it is currently done:

1. The neutralization can be made only over an extremely narrow frequency range, and

2. Since, in most cases, amateurs do not have ade-

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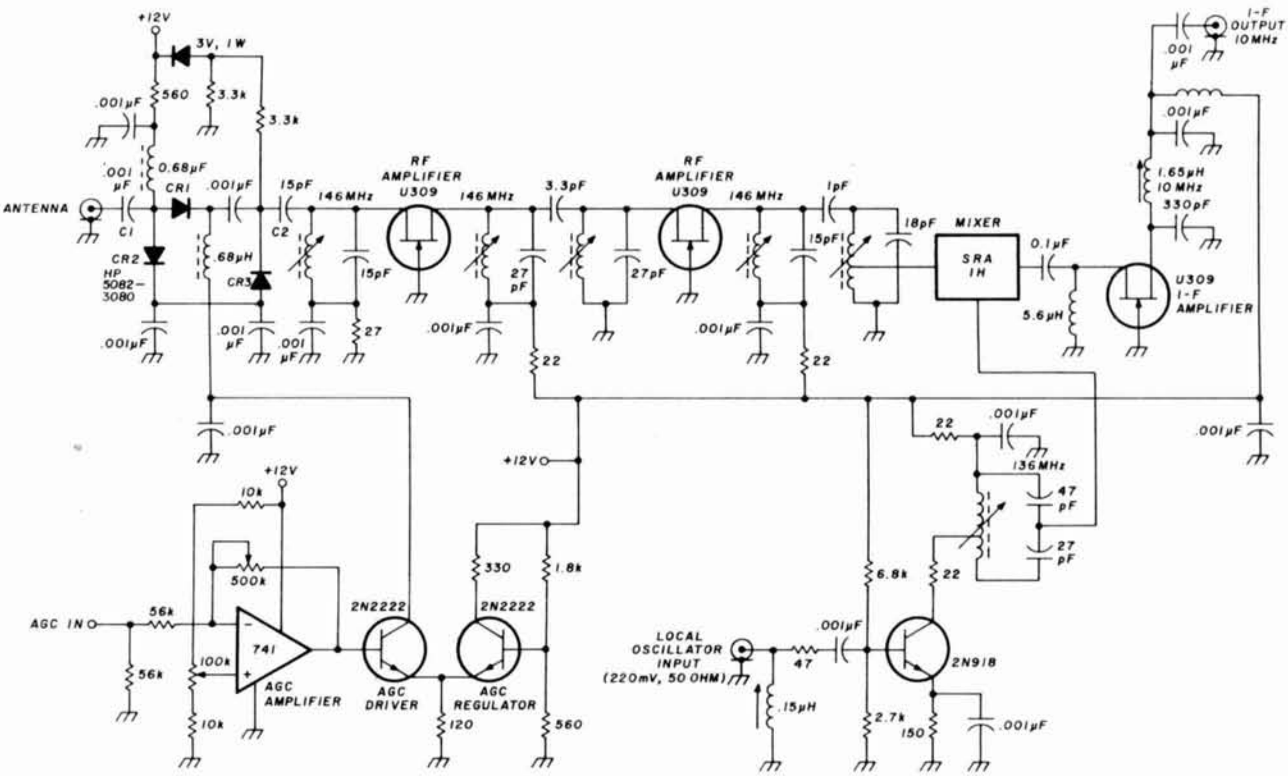
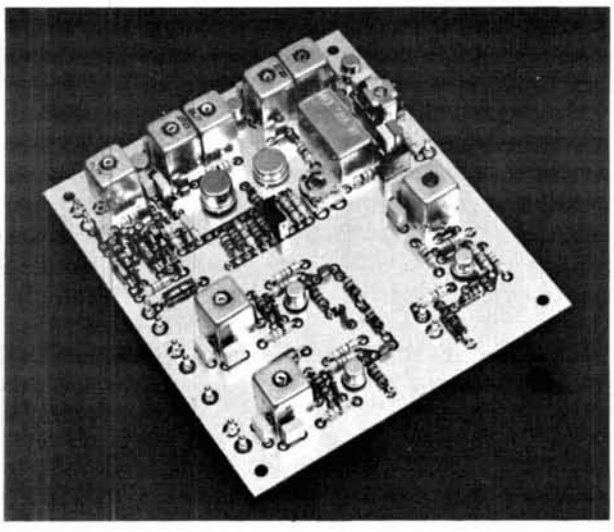


fig. 1. A two-meter converter with +15 dBm intercept point, 16 dB power gain, and less than 5 dB noise figure.

quate test instruments, an exact adjustment is seldom achieved, which degrades the two-tone IMD performance.

This converter has five tuned circuits at the input frequency which results in an overall bandwidth of barely 4 MHz. Proper tuning of this converter can best be achieved by the use of a suitable sweep generator.

These five tuned circuits provide an image suppression of 60 dB for an i-f of 10 MHz, or more than 80 dB for an i-f around 30 MHz. To simplify the circuit, U-309 fets are used which should have an $I_{DSS} = 20\ mA$. The third-order IMD of these transistors can be neglected as compared to the performance of the SRA-1H mixer. The overall gain between the antenna termination and the mixer input is about 10 dB. Therefore, the overall intercept point of the converter is +15 dBm with a noise figure of slightly less than 5 dB. Part of the fairly high noise figure is due to the 1-dB loss in the pin-diode attenuator.



Photograph of the converter board. Two additional output stages have been added.

This converter uses the ground-gate field-effect transistor circuit as described in reference 2. However, in the original version,[1] a special bipolar transistor in grounded-base configuration was used and provided a suitable wideband match for the mixer. This converter is shown in **fig. 2** as part of a transceiver that uses a frequency-locking system to stabilize the free-running oscillator in increments of 6.25 Hz (semi-synthesized).

unconditionally stable low-noise input stage

The low-noise preamplifier shown in **fig. 3** is based upon a circuit suggested by AEG-Telefunken in the 1950s and first published by me in English.[3,4] It was used to achieve extremely low-noise input stages with triodes while avoiding neutralizing circuits with

fig. 2. Photograph of a semi-synthesized 2-meter transceiver using the converter shown in fig. 1.

their inherent mass production problems. The same feedback arrangement not only avoids all instability problems but also improves the dynamic range. With practically all other neutralizing circuits a certain degree of performance reduction is observed.

The amplifier shown in **fig. 3** is a mixture between a grounded-gate and a grounded-source circuit. It is a bridge arrangement which neutralizes the feedback

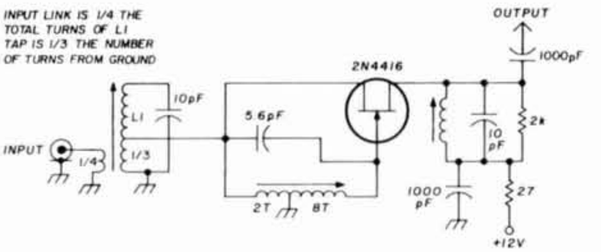


Fig. 3. Low-noise pre-amplifier with rf feedback to provide unconditional stability and low distortion.

capacitance between gate and drain. In addition, the input impedance (*I/S*) is transformed in parallel between the gate and ground and provides the necessary wideband characteristic. A noise figure of between 1 and 2 dB, using the inexpensive 2N4416, can be easily achieved and the gain is roughly 15 dB. The circuit is unconditionally stable. This is the only circuit known to me which combines optimum matching for best noise, lowest input swr, and best matching for high power gain.

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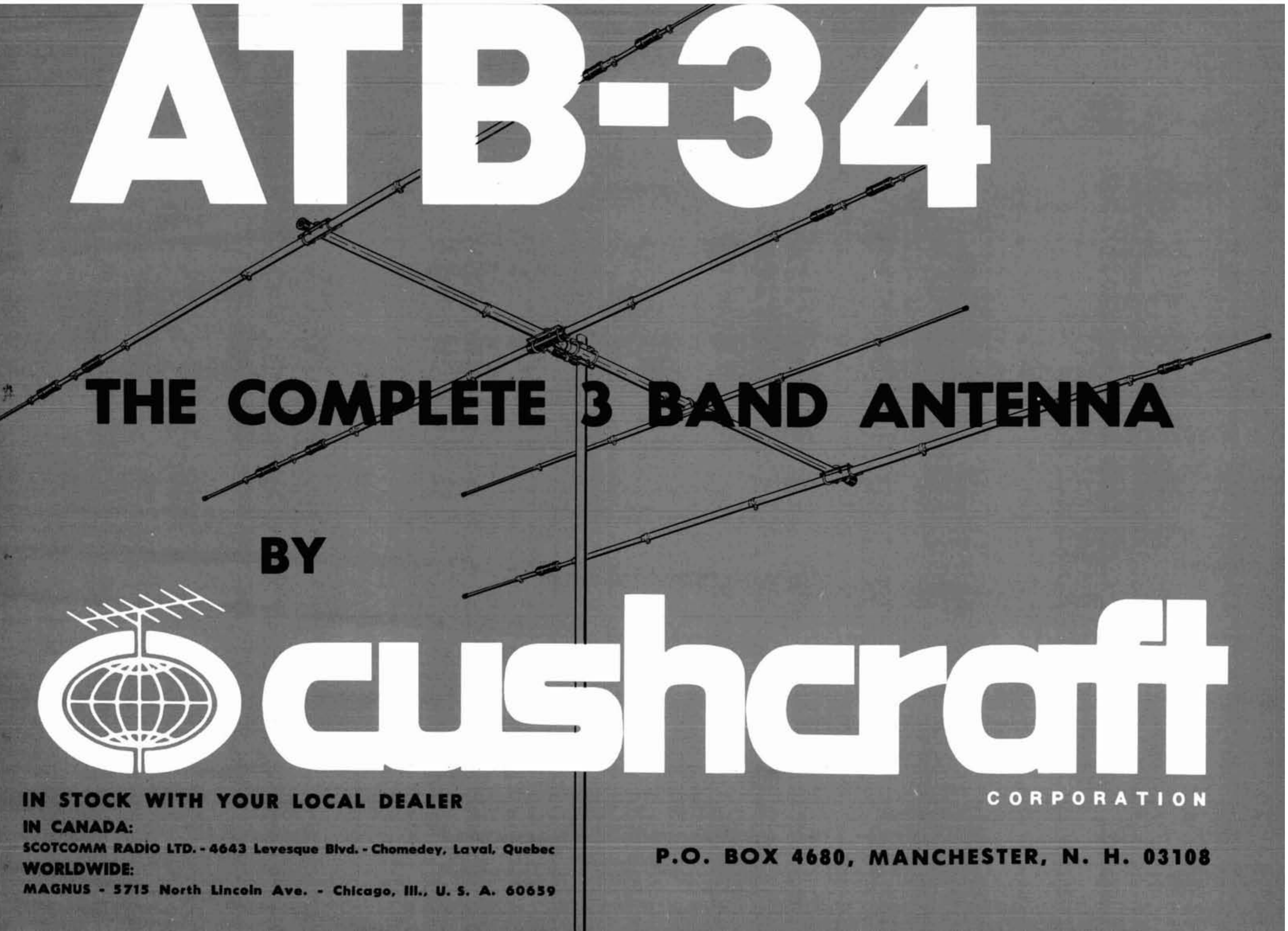
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More Details? CHECK—OFF Page 126

caption device for slow-scan television

An easy way to dress up your video with professional-appearing captions

I built a camera for slow-scan television (standard 625 line) and needed a method to rapidly change the material being televised. Because of space limitations, the camera had to be located fairly close to the subject. Using an *f*1.9, 16-mm focal-length lens resulted in a subject area of 4 x 4 inches (102 x 102mm) with the camera 9 inches (230mm) away, giving a 3¼-inch (83mm) square reproduction on my monitor, which uses a 5-inch (127.5mm) type 5FP7 tube.

On the receiving end the monitor-screen size is unimportant since, with the 128-line system used, the screen will be filled depending on amplitude-control settings.

description

The caption device is based on the use of plastic-strip magnets. The type I used are strip magnets for refrigerator doors; however, small bar magnets about 0.2-inch square by 1-inch long (5mm by 25.5mm) work very well.

A sheet of 16-gauge steel plate 9-inches (230mm) square was used as a backing plate to hold the caption board (described later). The backing plate was mounted on a frame so that it was square with the camera. The backing plate can be painted a matte black or covered with plastic self-adhesive material, which is obtainable at most hardware stores.

The next step is to make the caption boards. Two boards were made, one for letters and numbers and one for prepared pictures and captions. These are called the "variable caption board" and the "prepared caption board." Each is described separately.

variable caption board

For this board I obtained a piece of 0.1-inch-thick (2.6mm) card stock and cut it to 9.5 by 5.5 inches (242 by 140mm). The size depends on the lens you are using, but this size was for the *f*1.9, 16-mm lens mentioned previously.

I then cut the plastic magnet material into strips about 0.15 to 0.2 inch (3.8 to 5mm) long using a fretsaw. If you can obtain magnets near this size, so much the better. The objective is to obtain strips of magnetic material measuring 0.15 to 0.2-inch (3.8 to 5mm) square.

The card was marked as shown in **fig. 1**. Slots were cut using a sharp modeling knife and straightedge. The card was then put onto a flat surface and the magnets inserted into the slots so that the two original surfaces of the magnets were face down and face up, respectively. The rough-cut edges should mate with the walls of the cutout slots.

The magnets should now project from the card on the side facing you but will be flush with the face of the card on the other side. While the card is in this position, mix some two-part quick-set epoxy and place blobs around the magnets. When the epoxy is cured, the card can be covered on the flush side with a piece of plain white paper using impact adhesive.

We now have a basic board, plain white on one side with slightly projecting magnets on the other. The board will remain in position when placed against the steel supporting plate.

Alphabets. The next step is to prepare a couple of

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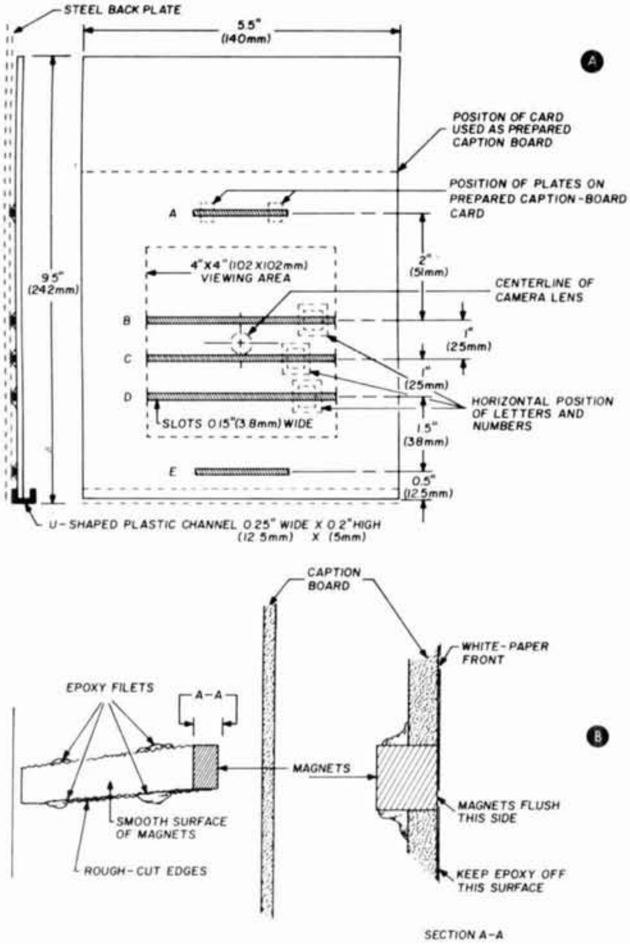

fig. 1. Caption-board construction, A, and details for mounting the magnets onto the board, B. Dimensions are for the variable caption board. For the prepared caption board only cutouts and magnets marked A and E are used.

alphabets and numbers to use with the variable caption board. Purchase a sheet of rub-on letters (available from stationery stores or commercial-artist suppliers). Choose a bold condensed type face about 3/4-inch (19mm) high.

Using a piece of clean white card stock, rub on two of each letter in the alphabet and all the numbers. Keep them within the lines as printed on the sheets and try to space them equally for a professional appearance. Burnish letters and numbers using the rub on sheet backing.

Next, carefully cut the characters into identical-height strips (except the letter **Q**, which must be slightly larger than, say, **A** or **B**). Use a square and cut each character from the strip into a rectangle as shown in **fig. 2**.

Obtain some thin steel washers about 3/8-inch (9.5mm) in diameter. Mark the exact center of the back of each letter card, then affix the washer in position using double-sided adhesive tape. You now will have alphabets and numerals that can be put onto the prepared magnet card in all sorts of combinations. If you cut the letter or number cards so they are square, they should mate evenly.

Applications. The magnet board can be made off-camera then placed against the steel supporting plate for on-the-air display. The result will be a neat and flexible method for originating captions. If you can't find the recommended steel washers, small pieces cut from tin-can tops can be used.

prepared caption board

This board is similar to the variable letter and number board but uses fewer magnets. It is prepared in the same way as the variable caption board (see details A and E, **fig. 1**). The two magnets shown hold the board against the steel backing plate and also hold the prepared material in place.

For this board, cut thin pieces of card stock to 5½ by 7 inches (140 by 180mm). Use a prepared stencil with a 4 by 4-inch (102 by 102mm) cutout. Draw lightly in pencil a 4 by 4-inch (102 by 102mm) outline on each sheet so that it's centered and in exactly the same position each time. (Retain the stencil for future use.)

Next, paste up your photos, rub-on letters, or test cards on each sheet within the 4 by 4-inch (102 by 102mm) outline. These sheets will usually contain standard material used on the air such as, "CQ de XXX," "My name is Joe," and so forth.

Returning now to the cardboard with the two magnets inserted, epoxied and covered with white paper, the next step is to make a small ledge along the bottom to hold the bottom of the prepared material. The top is retained by steel washers or tinplate affixed to the back of each sheet next to the magnet inserted in the cardboard. The bottom holding rail is made from a piece of plastic sliding door runner such as used in cupboards. It is held in place with impact adhesive.

The finished board can now be placed onto the steel

Rear of the variable caption board, left, and front of the prepared caption board, center. The rear of a prepared caption card is shown at right. The two steel wahsers hold the card to the magnet board. A prepared caption card is shown at the bottom.

Storage rack for the caption letters and numbers. It is made of cardboard with strips of magnet cemented to the surface for easy removel of letters. Note the pieces of tinplate on the backs of removed letters.

backing plate, which holds the board vertical and square. Prepared sheets can be quickly dropped into the bottom rail and pushed gently back. They will remain in position easily.

You can also use this system to televise unprepared material simply by placing your drawing or picture onto the steel back plate. Use one or two bar magnets to hold the material to the plate.

final remarks

The system has proved its worth in use. With the two

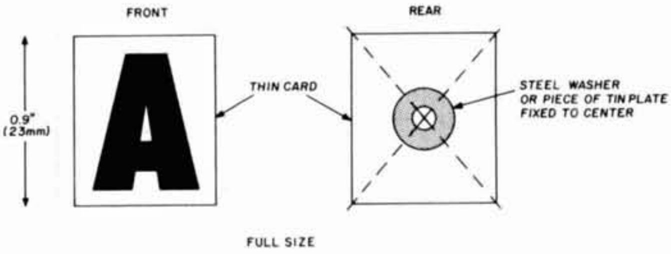

fig. 2. Construction of individual letters or numbers. The author used Letraset sheets (2683 in Europe; 113-72-CN in the USA).

magnet boards and steel back plate, material can be changed quickly and positioned in the same place during on-the-air contacts. Obviously the idea can be adapted to fit your own camera. The dimensions given are such that, with my camera on a flat table, the center of the magnet board 4 by 4-inch (102 by 102mm) viewing area lies in line with the center of the camera lens. A couple of stick-on dots indicate the horizontal position of the magnet board when placing it onto the steel back plate, and two pieces of tape are used to position the camera.

acknowledgement

My thanks to Bob Weston for the photographs.

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An effective noise blanking circuit developed by Kenwood that virtually eliminates ignition noise is built-in to the TS-520S.

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The VFO-520 remote VFO has been designed to match the styling of the TS-520S and provide maximum operating flexibility on the band selected on your TS-520S.

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The TS-520S is completely self-contained with a rugged AC power supply built-in. The addition of the DS-1A DC-DC converter (option) allows for mobile operation of the TS-520S.

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The TS-520S has 2 convenient RCA phono jacks on the rear panel for PHONE PATCH IN and PHONE PATCH OUT.

CW-520 — CW FILTER (OPTION)

The CW-520 500 Hz filter can be easily installed and will provide improved operation on CW.

AMPLIFIED TYPE AGC CIRCUIT

The AGC circuit has 3 positions (OFF, FAST, SLOW) to enable the TS-520S to be operated in the optimum condition at all times whether operating CW or SSB.

The TS-520S retains all of the features of the original TS-520 that made it tops in its class: RIT control • 8-pole crystal filter • Built-in 25 KHz calibrator • Front panel carrier level control • Semi-break-in CW with sidetone • VOX/PTT/MOX • TUNE position for low power tune up • Built-in speaker • Built-in Cooling Fan • Provisions for 4 fixed frequency channels • Heater switch.

Specifications

Amateur Bands: 160-10 meters plus WWV (receive only)
Modes: USB, LSB, CW
Antenna Impedance: 50-75 Ohms
Frequency Stability: Within ±1 kHz during one hour after one minute of warm-up, and within 100 Hz during any 30 minute period thereafter
Tubes & Semiconductors:

Tubes (S2001A x 2, 12BY7A)	3
Transistors	52
FETs	19
Diodes	101

Power Requirements: 120/220 V AC, 50/60 Hz, 13.8 V DC (with optional DS-1A)
Power Consumption: Transmit: 280 Watts Receive: 26 Watts (with heater off)
Dimension: 333(13⅛) W x 153 (6-0) H x 335(13-(13-3/16) D mm(inch)
Weight: 16.0 kg(35.2 lbs)

TRANSMITTER

RF Input Power: SSB: 200 Watts PEP CW: 160 Watts DC
Carrier Suppression: Better than -40 dB
Sideband Suppression: Better than -50 dB
Spurious Radiation: Better than -40 dB
Microphone Impedance: 50k Ohms
AF Response: 400 to 2,600 Hz

RECEIVER

Sensitivity: 0.25 uV for 10 dB (S+N)/N
Selectivity: SSB:2.4 kHz/-6 dB, 4.4 kHz/-60 dB
Selectivity: CW: 0.5 kHz/-6 dB, 1.5 kHz/-60 dB (with optional CW-520 filter)
Image Ratio: Better than 50 dB
IF Rejection: Better than 50 dB
AF Output Power: 1.0 Watt (8 Ohm load, with less than 10% distortion)
AF Output Impedance: 4 to 16 Ohms

DG-5

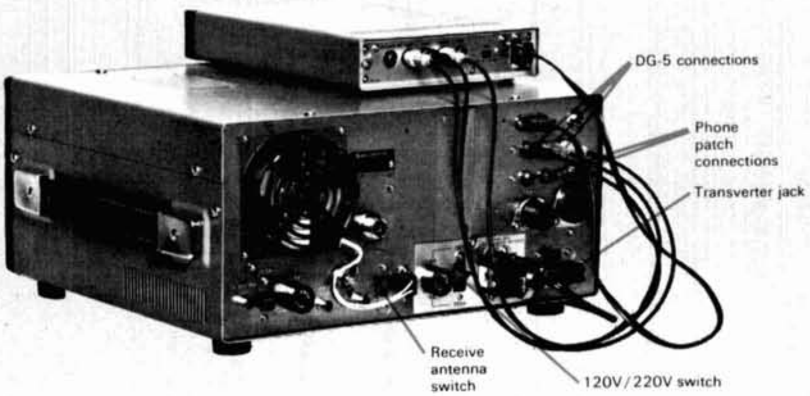
SPECIFICATIONS

Measuring Range: 100 Hz to 40 MHz
Input Impedance: 5 k Ohms
Gate Time: 0.1 Sec.
Input Sensitivity: 100 Hz to 40 MHz . . . 200 mV rms or over, 10 kHz to 10 MHz . . 50 mV or over
Measuring Accuracy: Internal time base accuracy ±0.1 count
Time Base: 10 MHz
Operating Temperature: -10° to 50° C/14° 122° F
Power Requirement: Supplied from TS-520S or 12 to 16 VDC (nominal 13.8 VDC)
Dimensions: 167(6-9/16) W x 43(1-11/16) H x 268(10-9/16) D mm(inch)
Weight: 1.3 kg(2.9 lbs)

DG-5
(optional)

The luxury of digital readout is available on the TS-520S by connecting the new DG-5 readout (option). More than just the average readout circuit, this counter mixes the carrier, VFO, and heterodyne frequencies to give you your exact frequency. This handsomely-styled accessory can be set almost anyplace in your shack for easy to read operation . . . or set it on the dashboard during mobile operation for safety and convenience. Six bold digits display your operating frequency while you transmit and receive. Complete with DH (display hold) switch for frequency memory and 2 position intensity selector. The DG-5 can also be used as a normal frequency counter up to 40 MHz at the touch of a switch. (Input cable provided.)

NOTE: TS-520 owners can use the DG-5 with a DK-520 adapter kit.



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More Details? CHECK—OFF Page 126

continuity bleeper

for circuit tracing

The four comparator amplifiers in the Motorola MC3302P IC are put to work in this circuit tester

When tracing circuits with an ordinary continuity tester, the procedure is slowed by the need to look away from the circuit to read the meter. The procedure is also risky because of the inability to distinguish between conductors and low-resistance components, such as inductors. Electromechanical buzzers are better, but they pass currents or induce voltages that damage transistor circuits.

These thoughts led to the design of the continuity tester described here. It delivers a continuous audio tone when the test terminals are connected by a resistance of less than about 1 ohm. The circuit under test never sees more than 3 volts or 300 mA, depending on its resistance. The tester ticks gently when switched on but open-circuited, as a reminder that the battery is on load, although the battery load is only 0.8 mA.

description

This complex pattern of operation is made easy by the Motorola low-power MC3302P IC, which contains four identical comparator amplifiers (**fig. 1**). These amplifiers are used in the tester as a measurement comparator, audio oscillator, loudspeaker amplifier, and economy ticker. The unusual differential input arrangement, **fig. 1**, allows the chip to operate down to zero common-mode voltage with a single supply of only 3 volts nominal. The current-sinking output transistor is particularly appropriate for the measurement comparator, U1A (**fig. 2**).

packaging

The MC3302P IC, together with battery and a 2-inch (51mm) diameter speaker, is enclosed within an aluminum box measuring 4¾ x 3¾ x 2 inches (121x96x51mm). The method of construction is shown in the photos.

circuit

Referring to **fig. 2**, U1A is a balance detector in a

By **R. C. Marshall, G3SBA**, 30 Ox Lane, Harpenden, Herts, England

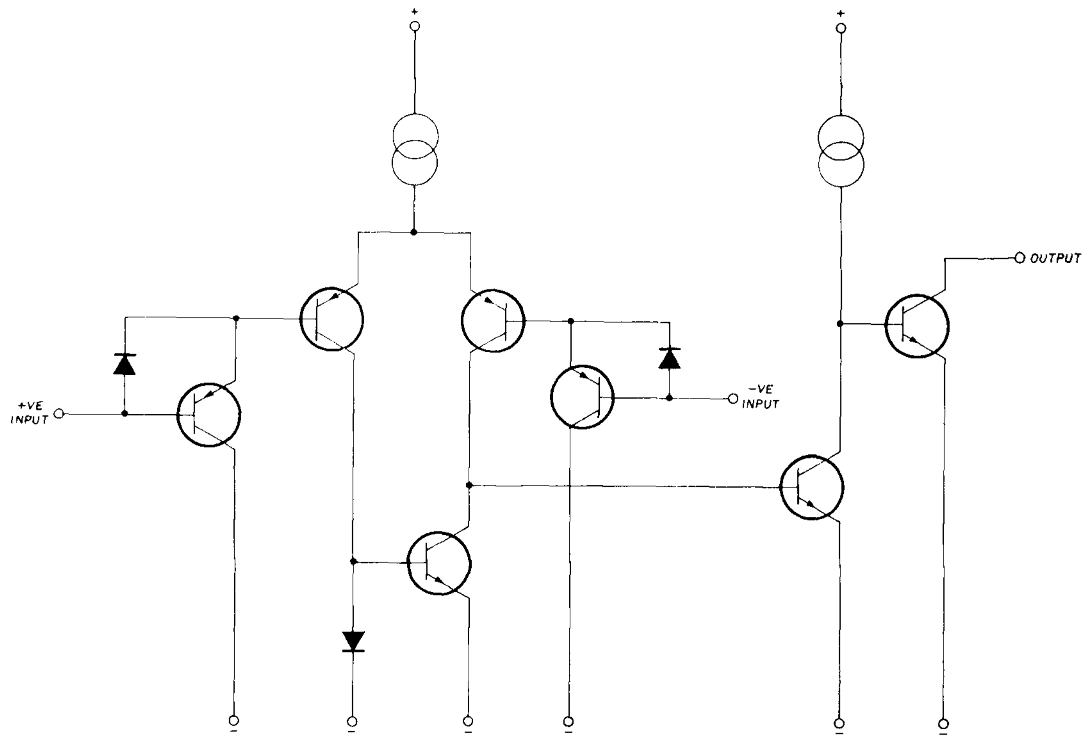


fig. 1. Schematic of one of four identical comparator amplifiers contained in the Motorola MC3302P IC, which is the heart of the continuity bleeper circuit.

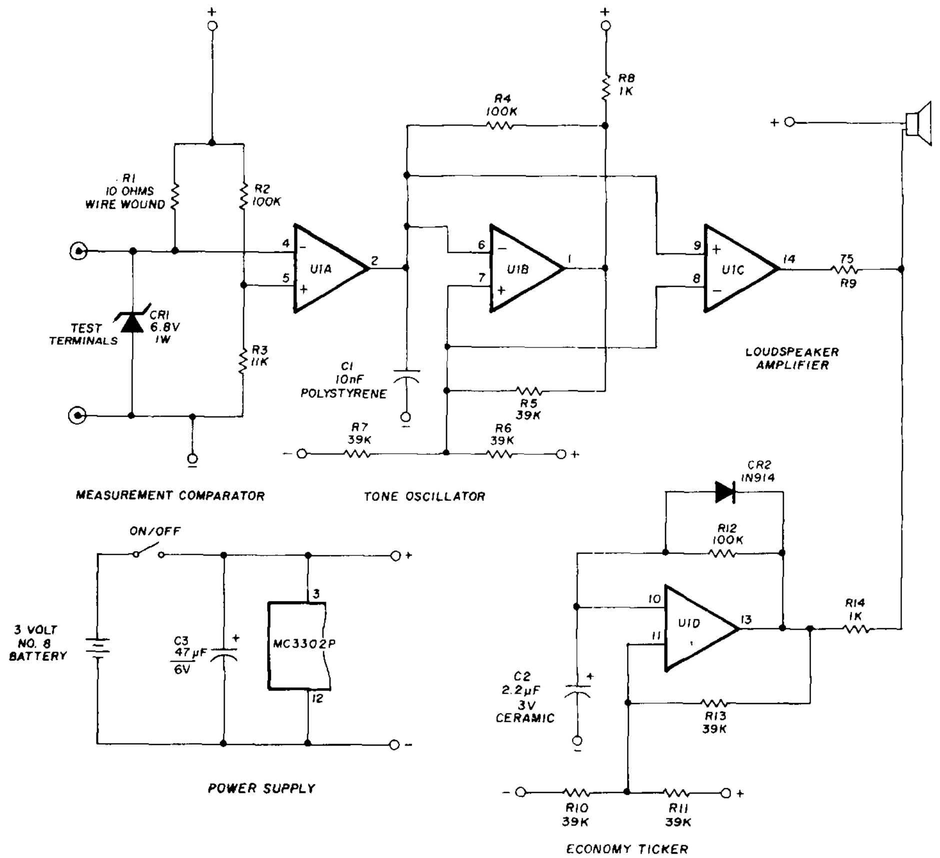
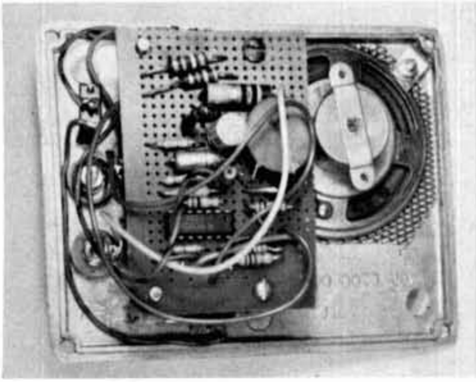


fig. 2. Continuity-bleeper schematic. U1A through U1D, all contained in the MC3302P, provide the four circuits for the tester. Pins 3 and 12 of the IC are used in the power supply. All resistors are ± 10% 1/8 watt, except R1.

bridge circuit. If the external resistance between the test terminals exceeds the product $R1 \times R3/R2$, the output short circuits C1 and disables tone oscillator U1B.

Measuring-circuit tolerance is determined by R1, R2, R3 in the measurement comparator, and by the offset of U1A. The threshold is between 0.7-1.6 ohms. If desired, R2 or R3 may be trimmed to set the threshold to, say, exactly 1 ohm or 0.5 ohm. Power zener CR1 protects the tester should it be connected to a "live" circuit.



Entire circuit including speaker and power supply is enclosed in a small aluminum box.

Comparator U1B is a 600-Hz oscillator. Resistors R5, R6, R7 provide positive feedback, so that comparator output snaps rapidly through a range nearly equal to the supply voltage. This output is returned to the other input through R4 and C1, whose voltage is an approximate sawtooth that causes the comparator to switch when the voltage increases to 2/3 and falls to 1/3 of the supply voltage, thus generating the square-wave output.

The input of speaker amplifier U1C is connected to ensure miminum supply current when U1A turns off the tone oscillator. Resistor R9 defines volume and current consumption when the tone is on.

The economy ticker, U1D, draws a 0.25 millisecond pulse of current through the speaker every half second. Its operation is similar to that of U1B, apart from the extra diode, CR2, which gives the short discharge time for C2 that corresponds to the "tick." If the tick isn't loud enough, resistor R14 may be reduced to 220 ohms (at the cost of increasing current consumption to 1.8 mA).

power supply

The battery must be able to deliver 0.3 ampere to the measuring circuit, so an inexpensive two-cell lantern battery is used in this application. Battery replacement will be rare, so the battery is secured between the supports of a Veroboard component panel. Connection is by insulated spring clips that bridge these supports.

With the availability of the MC3302P, the circuit described here has become a practicality. It can be a great time saver for both experimenter and professional in circuit testing.

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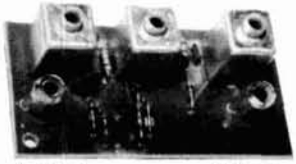

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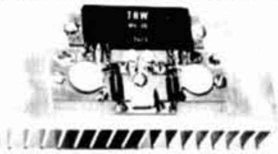

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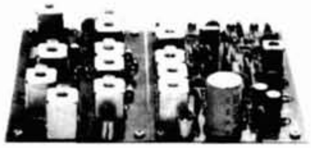

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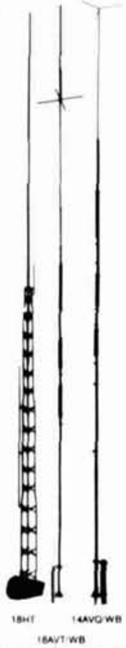

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repair bench

Joe Carr, K4IPV

simple antenna instrumentation

Regardless of the type of antenna you erect, it will rarely be absolutely correct despite the fact that instructions were followed carefully and textbook formulas were used. This is because such formulas are either based on an ideal situation or use certain assumptions that may or may not be valid in your particular case.

Fine tuning an antenna system is all but impossible without some sort of instrumentation to tell you what's going on. In fact, it's wise to have the instruments on hand that are described here, because they will tell you different things and will find their best use in different situations.

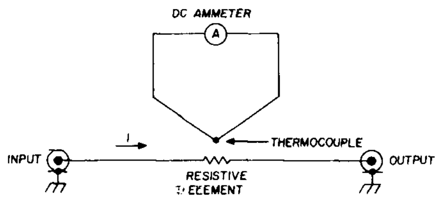


fig. 1. Internal circuit of an rf ammeter uses a thermocouple to sense heating in a resistance element.

rf ammeters

Radio-frequency currents can't be measured easily by regular ac ammeters but require a special type of instrument based on the thermocouple. The internal schematic of an rf ammeter is shown in **fig. 1**. The rf current, I, flows through a resistive element. Power lost in this resistance is converted to heat, which is applied to a thermocouple. The thermocouple consists of two strips of dissimilar metals joined to form a V. When this junction is heated, dc will flow through the meter. This direct current will be proportional to the rms value of the rf current. Rf ammeters are generally free of frequency effects up to 50 or 60 MHz. Some instruments, especially older types, must be used with care since they may be sensitive to the type of panel on which they're mounted. Some, for example, are marked "calibrated on a 1/8-inch (3mm) steel panel" and are only accurate when used with such a panel. The rf ammeter can be used to indicate current levels in a transmission line or as a means of calibrating a more convenient form of rf wattmeter. The power-indicating instrument is connected in series with the rf ammeter and is inserted between the transmitter and a non-inductive dummy load. Various power levels are applied

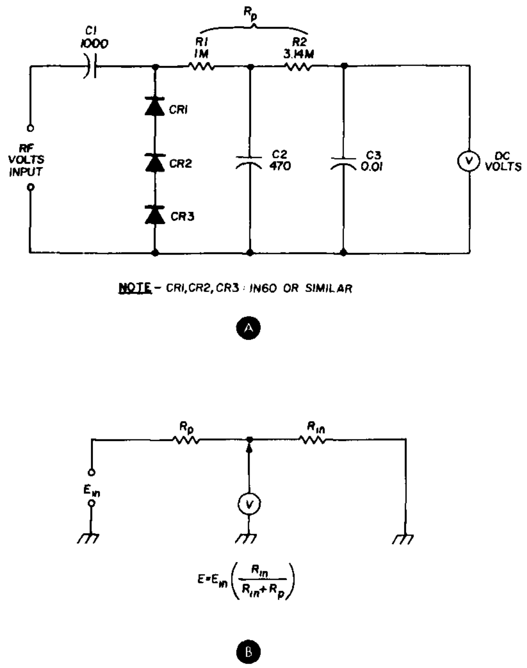


fig. 2. Rf voltmeter adapter for a vtvm. Sketch (A) shows a typical probe; (B) shows a voltage divider for indicating rms values.

By Joseph J. Carr, K4IPV, 5440 South 8th Road, Arlington, Virginia 22204

and the wattmeter is calibrated from the ammeter indications and the $P = I^2R$ relationship.

rf voltmeters

Rf voltmeters can be made using a dc voltmeter and a special probe or adapter. Such probes are called "demodulator" probes. This probe is shown in **fig. 2A**. Diodes CR1-CR3 are standard germanium signal-detector devices such as the popular 1N60. Use one diode for every 15 or so volts of peak rf voltage amplitude. Capacitors C2 and C3 are filters. This type of probe is a

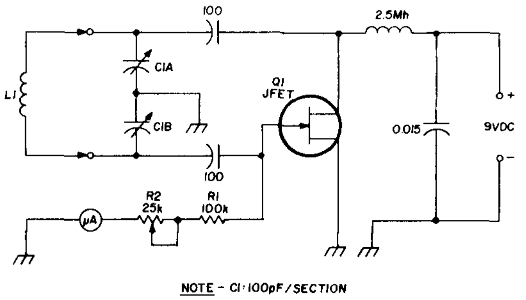


fig. 3. Circuit for a dip meter uses a jfet as the active element. This circuit is known as a gate dipper.

peak-reading type, but most of our interest will be in the rms values. These values are related by:

$$E_{rms} = \frac{E_{peak}}{1.414} \qquad (1)$$

Since a resistor is needed in the probe, and the voltmeter has an input impedance, we can construct a divider so that the voltmeter will read the rms value of rf sine waves. The voltage divider in **fig. 2B** follows the rule

$$E_{meter} = E_{in} \frac{R_{in}}{R_{in} + R_p} \qquad (2)$$

Where

R_{in} is the meter input resistance.

R_p is the probe resistance.

E_{in} is the input rf-signal peak voltage.

By combining the two equations we see that

$$\frac{R_{in} + R_p}{R_{in}} = 1.414 \qquad (3)$$

But R_{in} is usually equal to 10 megohms in most electronic voltmeters, so

$$\frac{10\ meg + R_p}{10\ meg} = 1.414 \qquad (4)$$

$R_p = (1.414)(10\ meg)$ or 14.14 megohms

Use precision 1% (or better) noninductive resistors. The resistors in **fig. 2A** are shown as two separate components because they were available.

dip meters

Originally known as grid-dip meters, these instruments are used to find resonant frequencies of tank circuits, or as a signal source, or to find (approximately) the values of capacitors and inductors. **Fig. 3** shows a typical dip-meter circuit; this one is a gate dipper using a junction fet. Inductor L1 is mounted outside the instru-

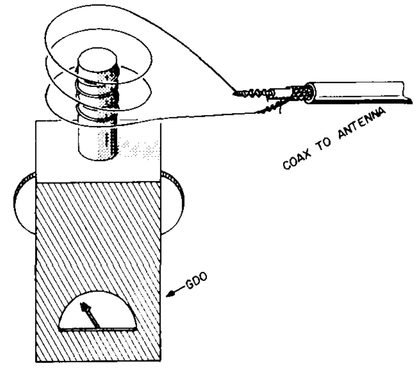


fig. 4. Method for coupling a dip meter to an antenna.

ment case. Energy from this coil is coupled to the coil in the resonant tank circuit being tested. When the L1-C1 tank of the dipper is tuned to the resonant frequency of the tank being tested, the amount of energy transferred will increase sharply which produces a slight dip in gate current. This property can be used to indicate the resonant frequency of the tank under test, which is read from the calibrated dial ganged to C1.

Dip meters usually are not too accurate as indicators of frequency, so a calibrated receiver should be used to sense the signal and read its frequency.

An antenna is basically a tuned resonant circuit, so it can be tested using the dip meter. **Fig. 4** shows a method by which the dipper is coupled to an antenna. The point at which the dip is noted is the antenna resonant frequency. A mistake often made when using dippers is

tuning too fast. The dip is slight, so one must tune very slowly.

Wheatstone bridge

Basic to many forms of instrument is the Wheatstone bridge of **fig. 5**. This circuit compares the outputs of two voltage dividers Z1/Z3 and Z2/Z4. Under these circumstances, where Z1/Z3 = Z2/Z4, the voltage between points A and B is zero. Under these conditions,

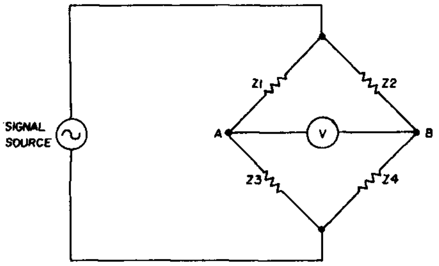


fig. 5. Wheatstone bridge, which is the basis of many test instruments.

$$Z4 = \frac{Z3Z2}{Z1} \quad (5)$$

In most cases impedance Z3 will be a calibrated standard resistance and Z4 will be the unknown. The values of Z1 and Z2 are not too critical, but they should not be too different from the other resistances in the circuit. This keeps the necessary amplitude of the signal source within reasonable bounds. In many cases $Z1 = Z2$ so that $Z3 = Z4$ at null. This makes calibration of Z3 easier. The meter at the center of the bridge may be either a voltmeter or a current meter. In fact, it is the current meter that's most frequently used as the null indicator.

the noise bridge

An adaptation of the Wheatstone bridge is shown in **fig. 6A**. This is the noise bridge and uses a diode noise generator and amplifier to drive two arms of the bridge. Transformer T1 is a toroid with trifilar winding. One coil is connected across the output of the noise generator amplifier, while the remaining two coils form arms of the bridge circuit. In another arm of the bridge is a calibrated 0-250 ohm potentiometer, R1. The last bridge arm is the antenna or other unknown.

The null detector in this version of the bridge circuit is a receiver. Although the coax cable to the receiver may be any convenient length, the cable to the antenna must be either extremely short relative to a half wavelength or it must be an integer multiple of half wavelength (1, 2, 3, . . .). This length is found from

$$L = \frac{492}{f_{MHz}} \cdot (V \cdot N) \quad (6)$$

where V is the cable velocity factor (0.8 for foam dielectric and 0.66 for regular coax) N is any integer.

This length is necessary because the impedance at the antenna feedpoint is reflected at half-wavelength intervals along the line. The noise bridge is connected to both the receiver and the antenna. The receiver is tuned to the desired frequency. Resistor R1 is then varied slowly until a null in noise level from the receiver is noted.

Fig. 6B shows an alternative form of noise bridge in which a 140-pF variable capacitor is connected in series with a 250-ohm potentiometer. A compensating 70-pF fixed capacitor is connected in series with the antenna. The variable capacitor is fitted with a calibrated dial that has a zero marked at the point where C1 = 70 pF (in other words, C1 = C2). Arbitrary calibration marks are scaled ± from this center zero. A null occurring when C1 is set to its zero point (70 pF) means that the antenna is resistive, which implies that the receiver is tuned to the antennas resonant frequency. If the null occurs either side of the zero point, then the antenna is reactive at the frequency indicated on the receiver.

The scale is marked X_C on one side of the zero point and X_L on the other side. If a null occurs on the X_C side, then the receiver is tuned on the high side of resonance. Alternatively, if the null occurs on the X_L side, then the receiver is tuned to a frequency that is too low. The antenna is resonant at the frequency indicated on the receiver dial when a null occurs at a point where the capacitor dial reads zero (C1 = 70 pF). Amateur noise bridges are made by Omega-T Systems and by Palomar Engineers.

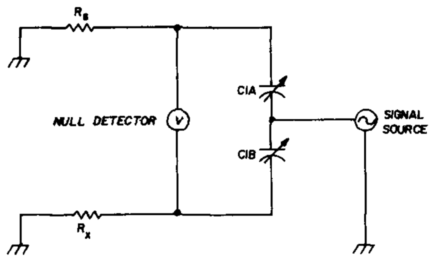


fig. 7. The antenna impedance bridge, another adaptation of the Wheatstone circuit.

antenna impedance meters

Another adaptation of the Wheatstone bridge is the antenna impedance bridge of **fig. 7**. Bridge arms consist

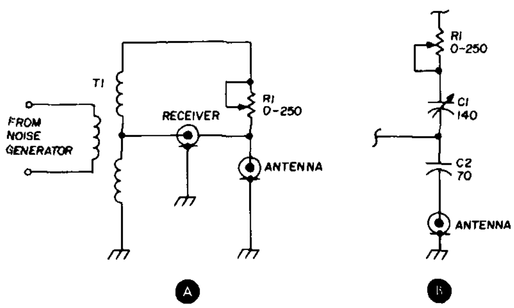


fig. 6. Applications of the Wheatstone bridge for antenna measurements. The circuit in (A) is the classic noise bridge; (B) is a version in which a capacitor is used to null reactance.

of C1A, R_s, C1B, and R_x. Capacitor C1A and C1B are respective halves of a differential capacitor. In this type of variable, capacitor C1A increases as C1B decreases. R_s is a standard resistance, 50 or 75 ohms in most cases. R_x is the resistive component of the antenna impedance.

This circuit will not measure reactance but it does allow you to tell whether there is a reactive component

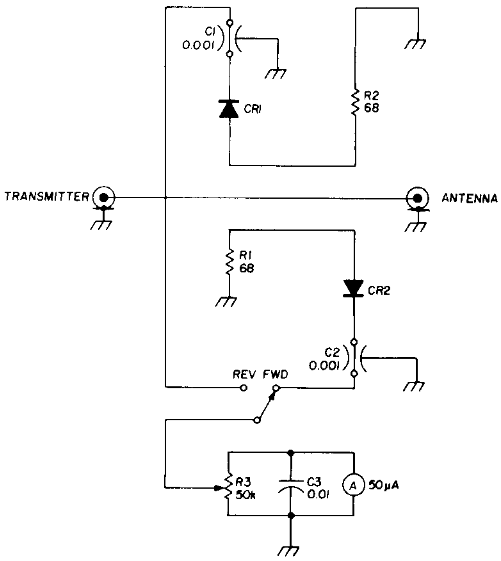


fig. 9. The Monimatch swr meter, popularized by ARRL publications. Heart of the circuit is the pickup sensor, which consists of two conductors, one for forward and backward swr indication (see text).

present. In a purely resistive situation the meter will null all the way to zero when the dial on C1 is set so that the bridge is balanced. If the meter does not null all the way to zero, then there is some reactance present. You can tell whether it is capacitive or inductive by varying the signal-source frequency until a frequency is found where the meter does null to zero. Antennas are capacitive below their resonant frequency and inductive above their resonant frequency. This information, then, tells you whether to lengthen or shorten the radiator to achieve resonance at the desired frequency.

Signal sources for the antenna impedance meter include ordinary signal generators, dip meters, or individual crystal oscillators. If you're very careful and use loose coupling, it's also possible to use a low-power transmitter as the signal source. I've used both a dip meter and those low-cost International Crystal OX oscillator kits in this application. I've also used the Leader LIM-870A antenna impedance meter (1.8 - 150 MHz). It is designed to mate with their dip meter, back-to-back, so that a complete signal source-bridge assembly may be formed.

swr bridges

There are several ways to measure antenna swr. One is to measure the voltage components of the forward and reflected power at some point in the transmission line. This relationship is given by:

$$swr = \frac{V_f - V_r}{V_f + V_r} \qquad (7)$$

where:

V_f is the forward voltage
V_r is the reflected voltage

Voltage varies along the line, and it's possible to reduce the *apparent* swr by trimming the coaxial cable. However, this results in an erroneous reading that may give you a false sense of security.

Fig. 8 shows a resistive swr bridge. It is a Wheatstone bridge in which R1, R2, R3 and the antenna impedance form the respective arms. Diodes CR1 and CR2 should be a matched pair but should be satisfactory if you match their forward and reverse resistances with an ohmmeter. Resistor R5 is used to trim out differences in R6 and R7. Normally, the microammeter is set to full scale with S1 in the forward position. Resistor R4 is adjusted to make M1 indicate full scale (100 μA). The switch is then turned to the reflected position and the meter needle drops to a current proportional to the swr. In most swr meters the dial will be calibrated in swr units.

Another type of swr indicator popularized in certain ARRL publications, is the Monimatch of **fig. 9**. The heart of this instrument is the pickup sensor. It consists of two conductors, one for each direction, arranged in parallel and in close proximity to the coaxial cable center conductor. In some versions enameled wires are slipped underneath the shield conductor of a short

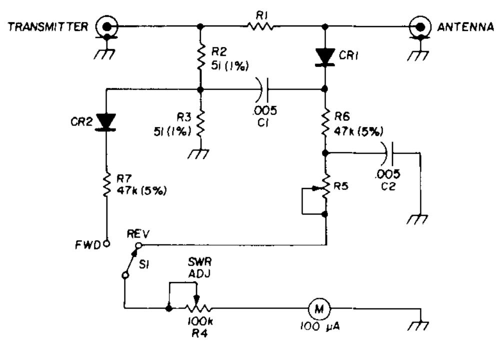


fig. 8. Resistive vswr bridge. R1 is 51 or 75 ohms, depending on coax-cable impedance.

length of coaxial cable. In others, the conductors are etched onto a piece of printed-circuit board inside a shield box. This class of instrument is different from other swr meters in that it becomes more sensitive as frequency increases.*

rf wattmeters

Figs. 10 and **11** show two popular forms of rf power meter. Of course, you can compute the swr of an antenna system by comparing forward and reflected power levels. The circuit of **fig. 10A** is known as the Micro-

*For still another swr instrument, see "Using the swr Indicator," *repair bench, ham radio*, January, 1977, page 66. **Editor**

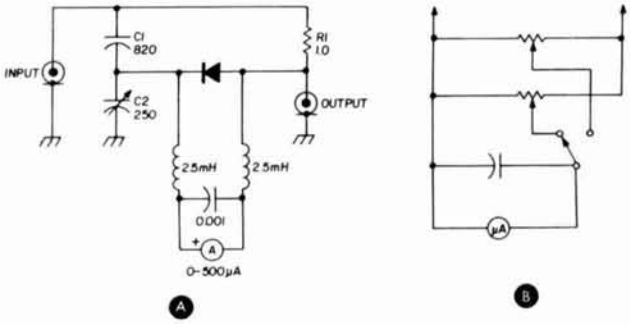


fig. 10. The Monimatch circuit for an rf wattmeter, (A). Various power ranges can be selected by the circuit in (B).

match, which was available commercially for years. It is also based on the ubiquitous Wheatstone bridge.

The bridge arms consist of C1, C2, R1 and the impedance of the antenna. R1 is only 1 ohm, so very little power is lost due to insertion of the instrument into the line. Various versions of the Micromatch have one, two, or three power ranges (10, 100, or 1000 watts full scale). These are selected as in **fig. 10B**.

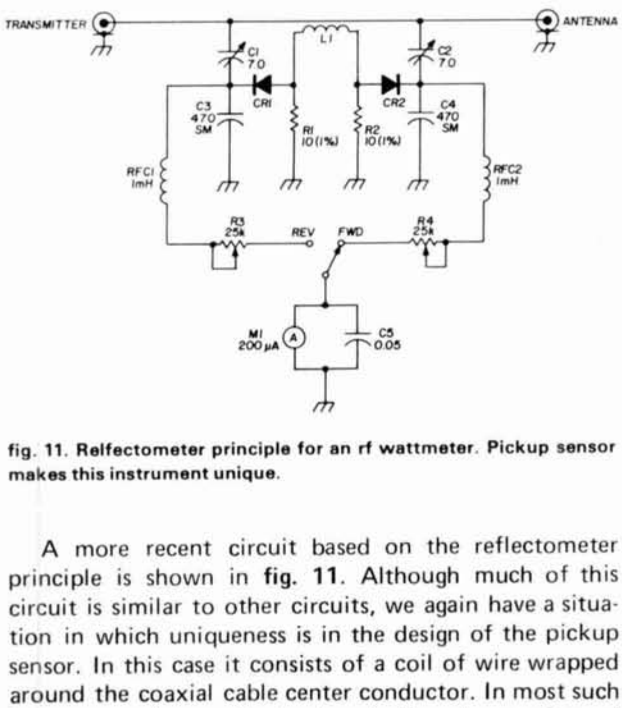
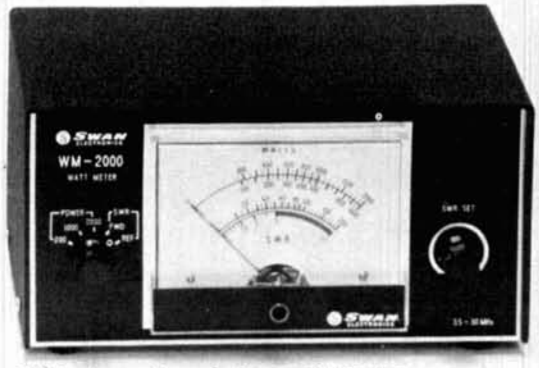


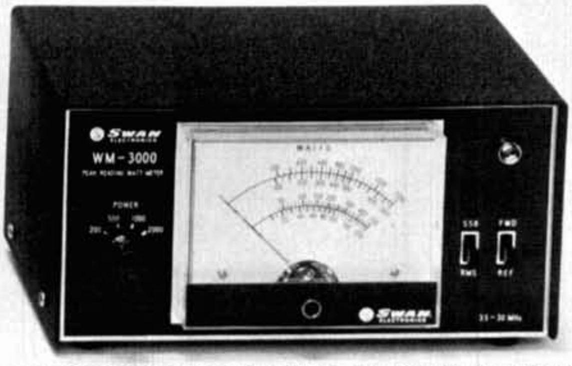
fig. 11. Relfectometer principle for an rf wattmeter. Pickup sensor makes this instrument unique.

A more recent circuit based on the reflectometer principle is shown in **fig. 11**. Although much of this circuit is similar to other circuits, we again have a situation in which uniqueness is in the design of the pickup sensor. In this case it consists of a coil of wire wrapped around the coaxial cable center conductor. In most such instruments the conductor is actually a heavy piece of solid bus wire and the pickup coil is wound on a toroid form. The wire conductor is threaded through the center of the toroid. This, in fact, is the basis of most low cost rf wattmeters on the market today.

Although it's a little excessive to expect any one amateur to own all of these different types of antenna instrumentation, it must be noted that every one of them is unique in its own way and all are eminently useful for doping out or troubleshooting reluctant antenna systems.

ham radio





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More Details? CHECK—OFF Page 126

sync generator

for a black-and-white 525-line television system

The National MM5320 IC is featured in this inexpensive circuit

This article describes a sync generator that supplies horizontal and vertical drive, blanking, and sync signals for a standard black and white interlaced 525-line TV system. It is simple and easy to build on the PC boards available,* or it can be reduced in size and mounted inside your TV camera. Almost all functions are performed within a single IC, the National MM5320 (about $5).

The PC board, loaded with parts, weighs in at only 4 ounces (113g). However, this lightweight board contains the equivalent of a 40 tube sync generator of the past and requires no maintenance to keep it running. Power consumption is only a few watts.

circuit description

The schematic is shown in **fig. 1**. The master oscillator consists of a 2.04545-MHz crystal, Y1, and a Motorola MC4024P multivibrator IC, U1. Actually, two multivibrators are contained in U1, so the spare is available at pin 12 on the PC board, **fig. 2**.

The circuit works well and delivers a symmetrical square wave at TTL level with almost any crystal. A disadvantage is that no provision is included for pulling the crystal to exact frequency. It's not necessary to do this with this sync generator — just plug in any HC6U crystal with the correct frequency.

The MM5320 has more features than needed for this application, such as gen-locking capability with circuits to reset both horizontal and vertical counters and a field-index output identifying the odd field at a 30-Hz rate. The MC5320 can be programmed to operate with either a 1.26-or 2.04545-MHz crystal. Therefore, certain pins are jumpered for this application, as shown in the PC-board layout, **fig. 2**.

The MM5320 also has burst-flag pulses, so it has color capability if a few circuits are added. Basically, a color generator would have a stable crystal-controlled master oscillator operating at 14.31818 MHz. This frequency is divided twice to derive the 3.58-MHz color subcarrier, which is processed to obtain a 2 volt p-p sine wave with adjustable phase shift.

The master oscillator also could drive a 7:1 counter to obtain 2.045 MHz, which is needed to drive the MM5320. The specification sheet for the MM5320 shows a minimum pulse width requirement of 190 nanoseconds. The 7:1 divider output is a 70-nanosecond pulse, so some means, such as a one-shot, could be used to square up the wave before it is applied to the MM5320. The master oscillator should remain within 4.0 Hz of 14.31818 MHz, so it can be seen that a few small problems need to be worked out. For black-and-white TV, we can ignore these difficulties and use the 2.04545-MHz crystal.

The circuit is simple. Buffering should be used between the mos-generator chip and output loads to

*An undrilled epoxy board is available from Bert Kelley, 2307 South Clark Avenue, Tampa, Florida 33609, for $6.00 postpaid in U.S.A.

By Bert Kelley, K4EEU, 2307 South Clark Avenue, Tampa, Florida 33609

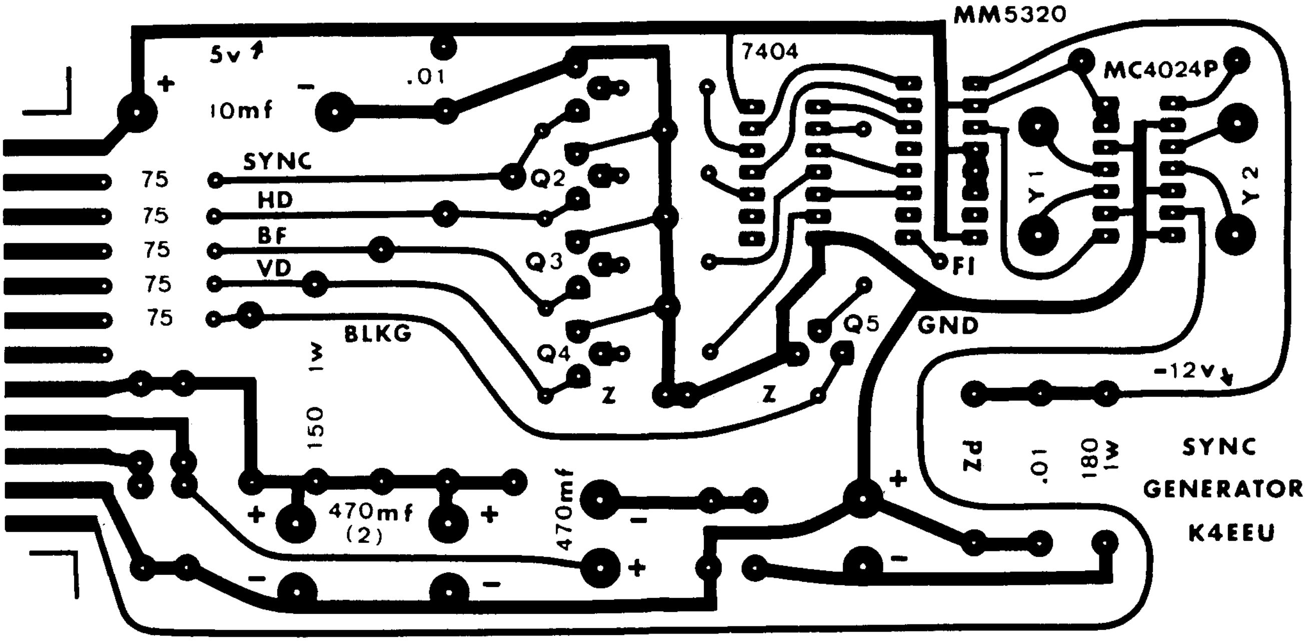


fig. 2. PC-board layout for the TV sync generator. Boards are available for $6.00 postpaid in U.S.A. from the author.

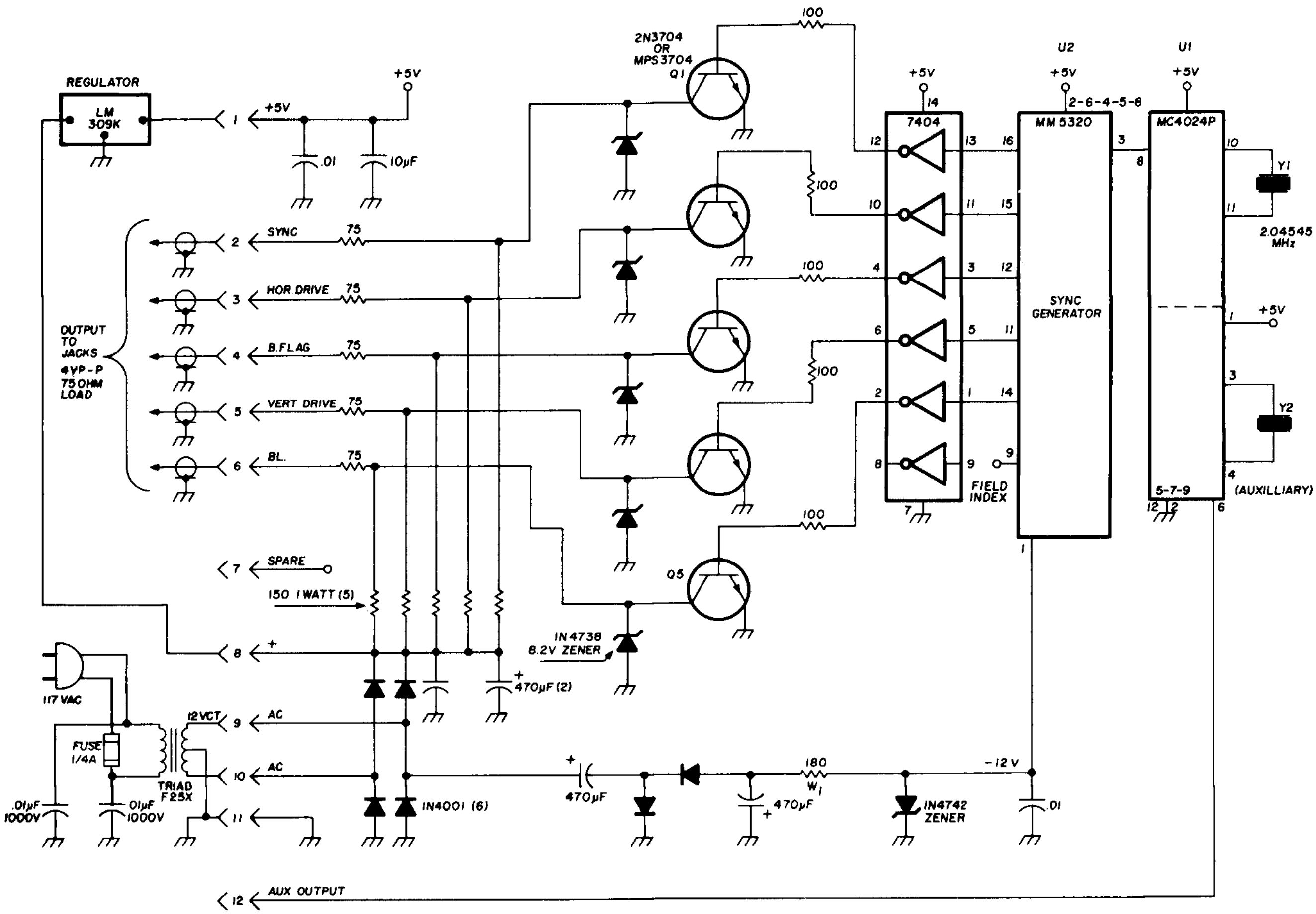
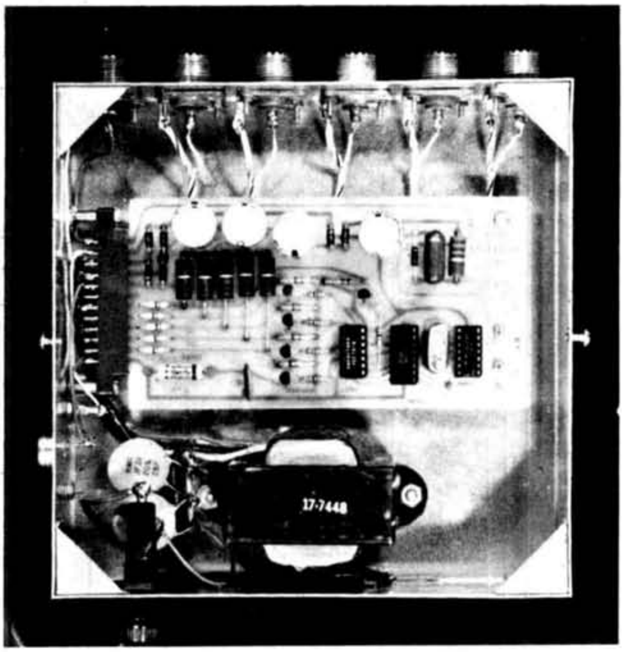


fig. 1. Sync-generator schematic. Circuit provides standard 4 volts p-p to a 75-ohm source impedance.

keep transients from damaging the chip. The circuit shown provides a 4-volt p-p to a 75-ohm load at 75-ohm source impedance. The voltage at the transistor collectors is limited by 8.2V zener diodes; when the transistor saturation voltage is deducted, the p-p voltage swing is almost exactly 8 volts to give 4 volts under load (negative-going pulses).

construction

The circuit board, which is 3 by 6 inches (77 by 153mm), fits an Amphenol 142-012-01 12-pin con-



Underchassis view of the TV sync generator showing power supply, right; PC board, center; and peripheral wiring. An LMB chassis is used, which also provides a heat sink for the regulator circuit.

nector. (Alternatively, connections can be soldered to the output pins.) IC sockets should be used, at least for the mos chip. Use 150-ohm 1-watt resistors where indicated. The 470-μF filter capacitors are of the radial-lead type. I used SO-239 output connectors; however, ordinary phono connectors would be all right (and much cheaper).

The circuit board should be mounted in a shielded enclosure such as the LMB 772, which also serves as a heat sink for the LM309K regulator.

The MM5320 is available from Nexus, and other parts are available from James Electronics.*

ham radio

*Nexus Company, P.O. Box 3357, San Leandro, California 94578 and James Electronics, 1021 Howard Avenue, San Carlos, California 94070.

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microcomputer interfacing: logical instructions

Most microcomputers manipulate information eight bits at a time. For example, the 8080A chip can move eight bits from internal register to internal register, between the accumulator and an external I/O device, and from internal register to memory. In additon, it can also perform arithmetic and logical operations, including add, subtract, and compare, and logical instructions AND, OR, Exclusive-OR, and complement. In this column, we will be concerned with the logical operations.

A truth table is the basic form that governs the one-bit logic operations. As a tabulation, it shows the relationship between all possible combinations of input logic levels and the subsequent outputs in a manner that completely characterizes the circuit functions.[1]

The truth tables for the AND, OR, Exclusive-OR, and complement operations are:

AND			OR			Exclusive-OR			complement	
B	A	Q	B	A	Q	B	A	Q	A	Q
0	0	0	0	0	0	0	0	0	0	1
0	1	0	0	1	1	0	1	1	1	0
1	0	0	1	0	1	1	0	1		
1	1	1	1	1	1	1	1	0		

These truth tables are called *one-bit tables* because the data words, A and B, each contain only a single bit.

When discussing logic instructions, it is useful to employ *Boolean symbols*. Such symbols originate from the subject of *Boolean algebra*, which is the mathematics of logic systems. This particular form of mathematics was originated in England by George Boole in 1847. Alphabetic symbols such as A, B, C, . . ., Q are used to represent logical variables with 1 and 0 representing logic states. Boolean algebra did not become widely used until 1938, when Claude Shannon adapted it to analyze multi-contact networks for telephone networks.

You should learn the basic Boolean symbols that are used in Boolean algebra computations, and thus all digital logic. These symbols include the following:

- $+$ which means *logical addition* and is given the name OR
- $\bullet$ which means *logical multiplication* and is given the name AND
- $\oplus$ which is given the name *Exclusive-OR* or XOR
- $\overline{A}$ which means *negation* and is given the name NOT

The negation symbol is a solid bar over a logical variable. Thus, the Boolean statement for a 2-input AND gate is $Q = A \bullet B$, or simply $Q = AB$, where the equality symbol ($=$) means that both variables or groups of variables are the same. It is useful to summarize the symbol operations for the three gates that are being considered:

AND	OR	Exclusive-OR	complement
$0 \bullet 0 = 0$	$0+0=0$	$0+0=0$	$\overline{0}=1$
$0 \bullet 1 = 0$	$0+1=1$	$0+1=1$	$\overline{1}=0$
$1 \bullet 0 = 0$	$1+0=1$	$1+0=1$	
$1 \bullet 1 = 1$	$1+1=1$	$1+1=0$	

Multi-bit logic operations are treated as many one-bit logic operations, therefore no new principles of logic are involved. The corresponding bits of one binary word logically operate on the corresponding bits of the second binary word to produce an overall multi-bit logic result. The length of the binary words can be any number of bits: two bits, eight bits, thirty-two bits, etc. Since the 8080A microprocessor performs multibit logic operations on eight-bit words, all of our examples will involve full bytes.

Consider the eight-bit logic variable, A. The individual bits in this variable can be labeled as A7, A6, A5, A4, A3, A2, A1, and A0, with A0 being the least significant bit (2^0 bit) and A7 being the most significant bit (2^7 bit). Also, consider the eight-bit logic

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

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

variable, B, which has individual bits that can be labeled as B7, B6, B5, B4, B3, B2, B1, and B0. The logic operation, $A \bullet B = Q$, means the following eight one-bit logic operations:

A0 • B0 = Q0	A4 • B4 = Q4
A1 • B1 = Q1	A5 • B5 = Q5
A2 • B2 = Q2	A6 • B6 = Q6
A3 • B3 = Q3	A7 • B7 = Q7

The result of the logic operation is the logic variable, Q, which has a least significant bit of Q0 and a most significant bit of Q7. In other words, *multi-bit logic operations are performed bit by bit via a series of one-bit logic operations*. It is easier to perform multi-bit logic operations if the multi-bit binary words are placed one under the other. Thus, if $A = 11011111_2$ and $B = 00100011_2$, then A • B is

$$\begin{array}{r} 11011111_2 \\ \bullet 00100011_2 \\ \hline 00000011_2 \end{array}$$

or $Q = 00000011_2$. We have performed a logical AND, and have used the relationships $0 \bullet 1 = 0$ and $1 \bullet 1 = 1$ in deriving the final result.

One of the more important uses for multi-bit logic operations is in situations in which the on-off state of external devices must be monitored. Consider the following system of eight devices:

bit position	devices	logic state information
Bit 0	pressure sensor	1 = pressure above setpoint 0 = pressure at or below setpoint
Bit 1	temperature sensor	1 = temperature above setpoint 0 = temperature at or below setpoint
Bit 2	velocity sensor	1 = velocity above setpoint 0 = velocity at or below setpoint
Bit 3	flow rate sensor	1 = flow rate above setpoint 0 = flow rate at or below setpoint
Bit 4	concentration sensor	1 = concentration above setpoint 0 = concentration at or below setpoint
Bit 5	valve A	1 = valve A open 0 = valve A closed
Bit 6	valve B	1 = valve B open 0 = valve B closed
Bit 7	power	1 = power on 0 = power off

We can call the group of eight bits the *status byte* for our system of eight devices. At any instant of time, the status byte will have a specific value. For example, the status byte 11100010_2 signifies that the pressure is at or below the setpoint, the temperature is above the setpoint, the velocity is at or below the setpoint, etc.

A group of logical instructions will permit you to determine the following characteristics about the previously listed external devices.

1. Which devices are on, open, or above the setpoint?
2. Which devices are off, closed, or at or below the setpoint?
3. Since the last time we checked, which devices have gone from on to off, open to closed, or above the setpoint to at or below the setpoint?
4. Since the last time we checked, which devices have gone from off to on, closed to open, or at or below the setpoint to above the setpoint?

In other words, using logical instructions you can determine not only the current state of the external devices, but also what changes have occurred since the last time that the devices were interrogated. Assume that we have just interrogated all eight devices and have found the current status byte to be 11101010_2, where the least significant bit, bit 0, is on the far right. One second ago, the status byte was 11101001_2. We wish to know what the current state of each device is, which devices have changed state during the last second, and in which direction. The steps that are employed to answer the questions are as follows:

1. Examine the current status byte. Determine the status of each external device from the logic state of its status bit.

The current status byte (CSB) is 11101010_2. From this value, we conclude that the pressure, velocity, and concentration sensors are all at or below their respective setpoints; the temperature and flow rate sensors are above their respective setpoints; and that value A, value B, and power are all on.

2. Perform an Exclusive-OR operation between the prior status byte (PSB) and the current status byte (CSB). A logic 1 in the result will indicate that the logic state of that device has changed.

The logical operation that we wish to perform is,

$$PSB \oplus CSB = Q1$$

where $PSB = 11101001_2$, $CSB = 11101010_2$, and Q1 is the result of the Exclusive-OR operation. Thus,

$$\begin{array}{rl} 11101001 & \text{PSB} \\ \underline{\oplus 11101010} & \text{CSB} \\ 00000011 & \text{Q1} \end{array}$$

and $Q1 = 00000011_2$. We conclude that only the pressure and temperature sensors have changed state.

3. Perform an AND operation between Q1 and the prior status byte (PSB). A logic 1 in the result indicates a device that has changed state from logic 1 to logic 0.

The logical operation that we wish to perform is,

$$PSB \bullet Q1 = Q2$$

where $PSB = 11101001_2$, $Q1 = 00000011_2$, and Q2 is the result of the AND operaton.

$$\begin{array}{ll} 11101001 & PSB \\ \underline{\bullet 00000011} & Q1 \\ 00000001 & Q2 \end{array}$$

Thus we can conclude that the pressure sensor has changed from being above the setpoint to now at or below the setpoint (logic 1 to logic 0 transition).

4. Negate (or complement) Q2, then ADD this complemented result with Q1. A logic 1 in the result indicates a device that has changed state from logic 0 to logic 1.

The logical operation that we now perform is,

$$Q2 \bullet Q1 = Q3$$

Since Q2 is 00000001_2, the complemented value of Q2 must be 11111110_2. The result of the AND operation is obtained as follows:

$$\begin{array}{ll} 11111110 & Q2 \\ \underline{\bullet 00000011} & Q1 \\ 00000010 & Q3 \end{array}$$

The result, $Q3 = 00000010_2$, permits us to conclude that the temperature sensor has changed from at or below the setpoint to above the setpoint (logic 0 to logic 1 transition).

The reason that we perform a series of logical operations is to determine what type of corrective actions must be applied to our system if it is not operating properly. If the temperature is below its setpoint, we may have to turn on a heater. If the concentration of the reactant is too high, we may have to turn valve B off and temporarily halt the flow of reactant into the system. If the pressure is above the setpoint, we may have an emergency condition and must shut the power off to the entire system. With a properly interfaced microcomputer, all such decisions can be easily and quickly made under software control and the necessary corrective actions initiated. However, you must be aware of the fact that mechanical and electromechanical devices typically have response times that are much longer than the decision times for a microcomputer. These response times must be taken into account in our software design, and are important considerations in the field of digital controls.

reference

1. R. F. Graf, *Modern Dictionary of Electronics*, Howard W. Sams & Co., Indianapolis, 1972.

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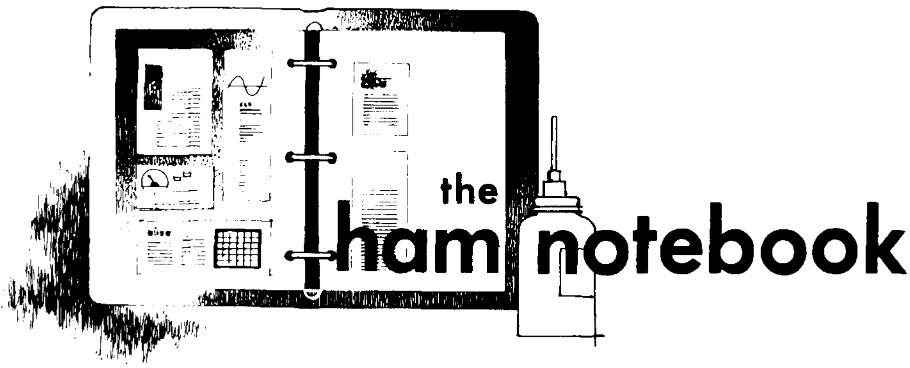

low-impedance headphones for Heath HW-16

The Heathkit HW-16 transceiver represents a very good value in amateur radio equipment, offering the beginning ham a transmitter and receiver combination designed for CW, and contains everything necessary to "get on the air" except antenna, speaker and key. The HW-16 also makes a good back-up rig in case your main station equipment is out of service or if your General class license is running out and you need some CW contacts to brush up on your code speed and fulfill the renewal requirement.

The problem arises when trying to use 8-ohm headphones to avoid disturbing the rest of the family. Low-impedance headphones do not mute the speaker. Fortunately, the solution is simple, quick, and costs nothing. If you are building the kit, you might want to wire in the new circuit from the start.

Fig. 1A is a reproduction of the audio output circuit diagram which appears in the Heath HW-16 construction and operation manual.

You will see that plugging headphones into this circuit does not disconnect the speaker, but places the speaker in series with the phones. With 8-ohm phones and an 8-ohm speaker, one-half of the audio still goes to the speaker. You will also note that to use earphones of any kind, you must have an 8-ohm load plugged into the speaker jack, because the two jacks are in series.

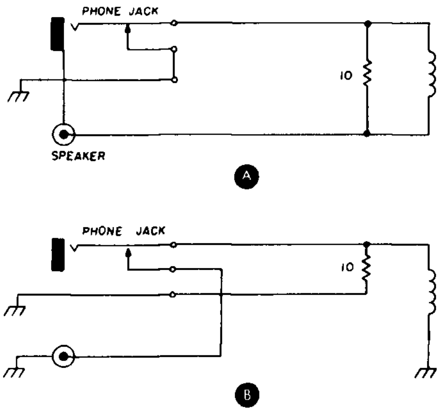

fig. 1. Schematic of the audio output circuit of the Heathkit HW-16 transceiver showing phone jack in series with the speaker jack as originally wired. (B). Revised schematic showing simple modification of the HW-16 audio circuit permitting phones or speaker to be used interchangeably.

Rewire the circuit as shown in **fig. 1B**, to disconnect the speaker completely when phones are plugged in. Note that in this circuit, however, you can use phones *or* speaker interchangeably. Since I've rewired my HW-16 this way, I operate more often because my family doesn't complain that audible CW notes interfere with their favorite television program.

John Pawlicki, WN8WJR

improved ssb reception with the Collins R392

A serious deficiency of the Collins R392, an otherwise excellent general-coverage receiver, is the severe distortion produced by strong ssb signals. While turning down the rf gain control solves the problem, it is extremely inconvenient when dealing with a variety of signals of different strengths — as when tuning across an amateur band.

The R392 was designed primarily for a-m and fsk use and uses a simple diode detector. Experiments with a product detector and outboard audio stages produced no significant improvement. Observing the signals at the i-f output connector on the front panel with an oscilloscope revealed that strong ssb signals caused considerable flat-topping in the stages ahead of the detector. When the rf gain control was backed off to the point where clipping did not occur, the detected audio sounded fine. This indicated that the problem was due to the agc system (designed for a-m signals) more so than the detector.

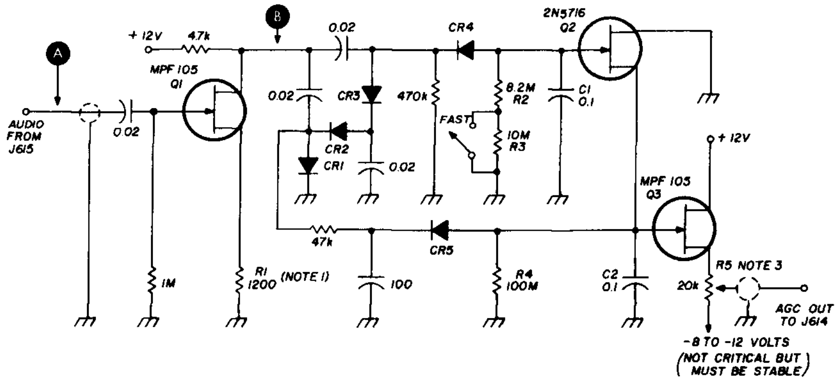

fig. 2. Hang agc circuit for the Collins R392 communications receiver. R4 is five 20-megohm resistors in series. Resistor R1 is adjusted for 7 Vdc at rest point B. Adjust R5 for zero out at agc out with no signal at the input (point A). All diodes are 1N459A or 1N914.

Ssb reception may be improved using the audio-derived hang agc circuit of **fig. 2**. The detector/audio module of the R392 conveniently has test points at the detector output, at which the audio signal may be picked up, and at the agc line, at which the externally-derived agc voltage may be introduced, with no modifications.

A hang-type circuit derived from that described by Hartke* is used. Q1 provides some amplification and high input impedance to prevent loading of the audio circuit at J615. The output of Q1 is rectified by CR1 and gated by CR5 onto C2, the agc hang capacitor. Simultaneously, CR1, CR2, CR3, and CR4, in a voltage quadrupler configuration, develop a control voltage gated to C1 via CR4. This control voltage keeps Q2 pinched off as long as a signal is present and permits C2 to remain charged and maintain the negative agc voltage. Upon removal of the signal, C1 discharges through R2/R3 and – after the hang time, when the pinch off voltage of Q2 is reached – it conducts and quickly discharges C2, restoring the receiver gain.

Source follower Q3 acts as a high input impedance buffer between C2 and the receiver agc line. R4 is required to provide a dc return for the gate of Q3. The time constant R4C2 is so long relative to R2C1 that the presence of R4 causes no significant discharge of C2 during the hang period. R5 serves to adjust the dc offset at the agc terminal. The minus supply voltage is not critical, as long as it is stable (I happen to have an 8-volt zener on hand). Strong signals charge C2 negatively, reducing the drain current of Q3 and causing the source voltage (drop across R5) to move in a negative direction.

installation

No major modification of the R392 is required. The agc circuitry was built on a small piece of perforated board and mounted in the external power supply unit. Connections to the detector output and the agc line were made at test points J615 and J614 respectively, on the audio module (lower deck assembly). In my case, lengths of miniature shielded cable were fitted with pin plugs to match the J614 and J615 sockets and the other ends wired to unused pins on the receiver power socket, J103. The connections to the agc module in the power supply were thus made via the power connectors, again using miniature shielded cable. Pin J on the receiver power connector is unused and is directly available. Several other pins are only needed when the original matching transmitter and microphone/headset are used. Therefore, the wire to pin C was unsoldered, capped, and this pin on J103 was used for the other agc connection.

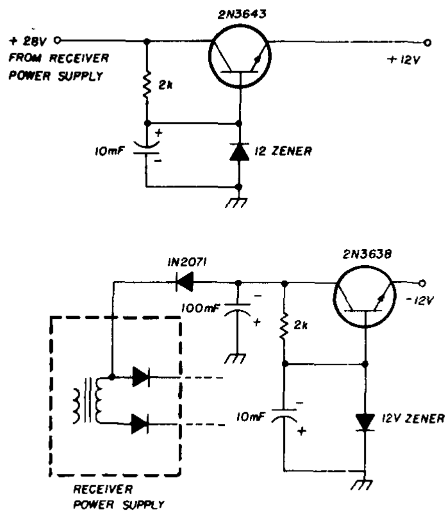

fig. 3. Power supply for the R392 hang agc system (fig. 1).

The front panel must be removed to gain access to the rear of the power connector. However, this is a relatively simple procedure following the instructions in the manual.† With the panel off, capacitor C104, 0.47 mF, which is switched into the original agc circuit by the bfo switch, should be disconnected. This improves the attack time of the new agc circuit. This is the larger of the two capacitors mounted on the inside of the front panel near the bfo on-off switch.

The required voltages are derived from the power supply unit shown in **fig. 3**. Simple zener diode regulators would also suffice since the current demands are only a couple of milliamps.

adjustment

Adjustment is simple. First, with no input signal to Q1, source resistor R1 is selected to provide a static drain voltage of about 7 volts to ensure maximum dynamic range at this stage. Next, with the connection to J614 also open, R5 is adjusted for zero volts at the agc output terminal. The circuit is then ready to go.

The added agc required no butchering of the original circuitry and has resulted in a dramatic improvement in ssb reception performance.

A.D. Lightstone, VE3LF

*J.L. Hartke, W1ERJ, "Solid-State Hang AGC Circuit for SSB and CW," *ham radio*, September, 1972, page 50.

†"Field and Depot Maintenance Manual, Radio Receiver R-392/URR," *Army Technical Manual TM 11-5820-334-35.*

test probe accessory

A volt-ohm-milliammeter (VOM) may be the most-used piece of equipment owned by the average radio amateur or experimenter. All voms are supplied with flexible insulated wire leads that have either pointed test probes or alligator clips at their ends, but not both.

Each type has its preferential use, but unfortunately not always in a compatible manner. Pointed test probes are handy for poking around to check resistors, capacitors, shorts, and the like; but for troubleshooting and checking voltages at tube sockets and so on, I prefer an alligator clip on the negative lead so that it may be clipped on to the chassis. Alligator clips are also handy for connecting to a circuit that may require monitoring for a period of time.

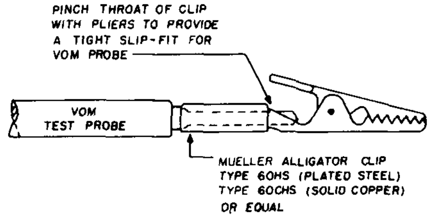

fig. 4. Pointed-tip test probe is easily converted to clip-type test probe by simple addition of an alligator clip.

My favorite vom had pointed test probes and, due to my experimental endeavors, I fashioned a bunch of patch cords with alligator clips on both ends. A few of these in red and black were left over, so I began experimenting with them. I found that I could slip the pointed test probes into the clips (Mueller) and obtain a reasonable fit. Now I have the advantage of using my test probes as is, or I can slip an alligator clip on when I want, as shown in **fig. 4**.

Ken Cornell, W2IMB

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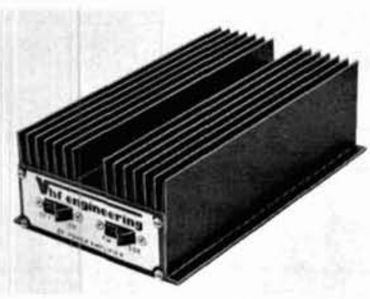
NEW products

synthesized two-meter fm

"The radio that goes where you do" is a feature of Wilson Electronics' new WE800 radio. The 800-channel, 2-meter portable synthesized radio has an internal nicad battery pack which allows installation as a mobile in the automobile or portable with shoulder strap and rubber duck. The unit provides 12 watts output in service, or 2 watt output as a portable with the internal nicad battery pack or ten AA nicads).

The WE800 has a low drain cmos synthesizer, with 45 milliamps on receive, and 450 milliamps on 2-watt transmit. Frequency range is 144-148 MHz in 5-kHz steps, 600 kHz offset up or down. Subaudible tones are available as an option. For more information, contact Wilson Electronics Corporation, 4288 S. Polaris, Las Vegas, Nevada 89103.

80-watt 450-MHz amplifier



A new broadband, high efficiency four-mode amplifier for the 420-450-MHz amateur band has been introduced by VHF Engineering. This new amplifier, the Blue Line BLE 10/80, will deliver 80 watts output in either class C or linear mode with a nominal 20 watts input. It is designed to be used with fm, a-m, ssb, or CW rigs in the 10-watt class. A similar amplifier, the Blue Line BLE 30/80, is designed to be used as an amplifier for rigs in the 30-watt class.

The Blue Line series of amplifiers from VHF Engineering are high efficiency, broadband, stripline amplifiers which have been designed for long life and reliable operation. Because of their unique, broadband design, they contain no tunable or adjustable components. Tuning or adjustment will not be required during the lifetime of the units. Automatic transmit/receive switching is provided through the use of sensing circuitry which detects the modes. They are designed for 12-14 Vdc operation in base station or mobile service.

These new four-mode Blue Line amplifiers are manufactured by VHF Engineering, 320 Water St., Binghamton, New York 13902. The BLE 10/80 is priced at $289.95 and the BLE 30/80 is $259.95. Both units are wired and tested.

antenna rotor control conversion kit

Have you had rotator damage? Removed the rotator? Waited for parts? No more! *Autobrak* is a complete conversion kit, including punched and finished cabinet for all HAM-M series 1, 2, and 3, rotator control units. *Autobrak* reduces the inherent problem of damaged rotator components due to instant brake engagement; it allows the antenna array to come to a coasting stop before brake engagement, thereby reducing stress on rotator components.

Other features include zener-regulated meter circuitry, adjustable brake delay, and handsome up-to-date styling compatible to most ham gear. The *Autobrak* is priced at $39.95 (plus shipping and handling, $1.75 in the United States). For more information write to Kampp Electronics, Post Office Box 43, Wheaton, Illinois 60187.

vhf frequency counter

Sencore, manufacturers of high quality test equipment for radio-TV service and communications, is offering a new FC45 frequency counter with continuous frequency checking capability from audio through vhf (and uhf bands with an optional PR47 uhf prescaler).

The FC45 audio through vhf counter is highly sensitive, with 25 millivolts average throughout the band, for circuit servicing with a pickup loop that does

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not upset the circuit during frequency tests. The FC45 is highly accurate with a one part in one million tolerance, using a high-accuracy temperature-controlled oven. Accuracy is better than FCC requirements on all bands including uhf. An eight-digit display with all pushbutton action makes the FC45 extremely easy to use. An exclusive crystal checker is provided as an integral part of the FC45. The company has stated that the price of $395 is also far below any other vhf frequency counter on the market today; the optional PR47 prescaler, selling for only $125, is simply connected into the input cable and the FC45 readings multiplied by ten. If used with any other frequency counter, a separate PA202 Power Adapter at $9.95 is used. For more information write to Sencore, Inc., 3200 Sencore Drive, Sioux Falls, South Dakota 57107.

ASCII keyboard encoder

A new ASCII Keyboard Encoder is now available from Radio Shack in their popular Archer Project-Board kit form. With the Project–Board concept you purchase the printed-circuit board with a complete, step-by-step assembly instruction manual and parts list. All necessary parts for the assembly of the ASCII Keyboard Encoder, including a 63-key computer control keyboard, are available from Radio Shack, or the builder may use parts from his own junkbox.

The completed keyboard encoder can be used to provide inputs to all types of equipment designed to operate with ASCII inputs, such as TV typewriters, minicomputers, microprocessors, electric typewriters, or any other devices which require positive or negative ASCII encoded alpha-numerical characters.

Features of the Archer ASCII Keyboard Encoder include a repeat key to control all characters and symbols, a negative- or positive-going data valid strobe, latch outputs (stores last key code), shift and shift lock capability, true or false ASCII outputs, and six extra control keys. The encoder is able to handle an output of 833 characters per minute (cpm) and has a repeat key rate of 208 cpm. An external power source of 5 Vdc at about 500 mA is required to power the encoder.

The Archer ASCII Keyboard Encoder Project-Board with complete assembly and parts manual is priced at $14.95. All parts needed to assemble the encoder, including the circuit board, manual and keyboard, but excluding hardware and case, are available from Radio Shack for $57.80. Archer Project-Boards are available exclusively from Radio Shack stores and dealers in all 50 states and Canada.

tape antennas

Tualatin Valley Labs is presently manufacturing indoor antennas for two meters. Available is a three-element beam at $9.95, and a vertical, ½-wave J pole design for $7.95.

The antenna elements are made from ¾ inch (2cm) wide aluminum tape designed specifically for rf applications. The adhesive backing on the tape is also conductive, allowing antenna connections to be made without the use of solder. Installation can be on any flat, dry surface, such as walls, doors, or inside closets. The tape alone is also available for your own experimentation in 10-foot (3m) lengths for $7.50 per 10 feet; 25 feet and 50 feet lengths are available for $15.00 and $25.00 respectively. For more information, write to Pat Adamosky, Tualatin Valley Labs, 18285 N.W. Parkview, Portland, Oregon 97229.

receive multi coupler

Did you ever wish you had a way to use one antenna to feed two vhf receivers? Well, it's here! The *P13 Receive Multicoupler Unit*, from Hamtronics, Inc., is a dual-output fet preamplifier which provides low-noise gain and power splitting to drive two receiver inputs from one antenna.

The unit gives about 15 dB of gain in each of the two channels. If desired,

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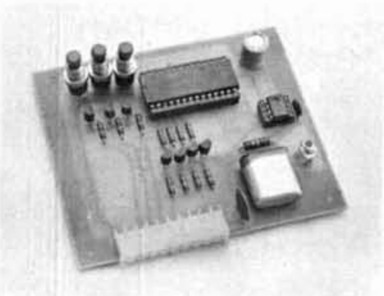
outputs can be on somewhat different parts of the band with some reduction in gain. Standard models are available for any segment of the 26-230 MHz vhf range.

Price of the P13 Receive Multicoupler kit is $12.95. The P24 wired and tested model is $24.95, and the P50 model, which is housed in a die-cast box with BNC connectors, is $49.95.

For a complete catalog of vhf and uhf modules, send a self-addressed stamped envelope to Hamtronics, Inc., 182 Belmont Road, Rochester, New York 14612.

real-time clock



Does your computer know what time it is? If not, you should add this new hardware peripheral from TED that will allow your computer to keep track of the time of day for control, timer and game applications. The *Real Time Clock* employs the latest cmos technology, resulting in high reliability, small size and low power consumption. The inputs and outputs are TTL-compatible and provide simple connection to your system with four wires from a parallel output port, ground, positive 5 volts, and positive 12 volts.

The *Real Time Clock* may be used with any computer system and includes complete hardware and software documentation and features crystal-controlled accuracy with a trimmer to allow exact frequency setting. Push buttons allow rapid and easy time setting. The clock can be operated from a separate power supply or battery, permitting it to keep the correct time, even when the rest of your computer system is off.

The crystal oscillator operates at a frequency of 3.579600 MHz and is divided by a factor of 59,600 to yield an exact 60-Hz frequency which is then applied to a digital clock chip containing the components to generate

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a six-digit 12- or 24-hour time display. The time information from the clock is transferred in BCD form to the computer, one digit at a time. The digit-select inputs determine which of the six digits representing the time is present on the BCD output pins. A short program in machine language or BASIC may be used to transfer the time information from the *Real Time Clock* to the computer.

IC sockets are used for all of the ICs, and components are mounted on top-quality G-10 printed circuit board material. The unit is available in wired and tested form *only*, at a special introductory price of $39.95 plus $0.75 postage. Wisconsin residents please add 4% sales tax. For additional information, write TED, Box 4122, Madison, Wisconsin 53711.

two-meter frequency synthesizer

VHF Engineering announces their new *SYN II Synthesizer*, a high-quality synthesizer designed for use in virtually all two-meter rigs available on the market today.

The SYN II Synthesizer is designed to operate over a frequency range of 140 to 149.995 MHz in 5 kHz steps and is compatible with two-meter equipment using transmit crystals in the 6, 8, 12, or 18 MHz range, and receive crystals in the 15 or 45 MHz range. The synthesizer may be used with either fm or phase-modulated transmitters and with mobile or base transceivers. The SYN II features unique i-f programming which permits the unit to be used with receivers having i-fs in the range of 100 kHz to 30 MHz. Detailed programming instructions are given in the construction manual so that the builder may select the standard 10.7 MHz i-f or any other i-f frequency in the permissible range, permitting use of the SYN II with older commercial equipment as well as currently available two-meter units. Standard repeater offsets of +600 kHz and −600 kHz are provided, along with

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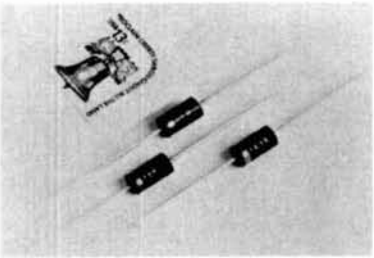
three user-selectable offsets in 100-kHz steps, permitting the user to operate into standard repeaters or with unique offsets. A modification kit is available for MARS and CAP offsets.

The synthesizer kit consists of high-quality epoxy-glass circuit boards, computer grade components, thumb-wheel switches, stylized cabinet, and a detailed construction manual. The kit is complete and requires no additional components. The SYN II is available from dealers nationwide or direct. The price of the kit is $169.95; wired and tested $239.95; programmed to your equipment $249.95. For additional information, write VHF Engineering, 320 Water St., Binghamton, New York 13902.

transient voltage suppressors



TRW has introduced a series of zener transient voltage suppressors (TVP) that protect circuits, systems, and equipment from voltage surges. The TVP1500 series rapidly changes the impedance value from a very high standby value to a very low conducting value when subjected to high energy transients. It shunts potentially damaging effects by clamping the voltage at some predetermined level.

The device features 1500 watts peak pulse power, a voltage breakdown range of 8.2 through 200 volts and a fast recovery time of approximately 1 picosecond. Series TVP1500, packaged in a polymer silicone case capable of withstanding temperature extremes to 400°C, also has a reverse standoff voltage range of 6.6 through 171 volts. All the devices in the series are 100 per cent surge tested, and can be used in telecommunication, automotive, computer, and dc power supply applications. The device is also available in 500 and 1000-watt peak pulse power ratings.

For more information write John Gamet, TRW Capacitors, 301 West "O" Street, Ogallala, Nebraska 69153.

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NEW! FM144-10SXRII

KDK

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

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

- **FREQUENCY RANGE:** Receive: 144.00 to 148.995 MHz, 5KHz steps (1000 channels). Transmit 144.00 to 148.995 MHz, 5KHz steps (1000 channels) + MARS-CAP.*
- **FULL DIGITAL READOUT:** Six easy to read LED digits provide direct frequency readout assuring accurate and simple selection of operating frequency.
- **AIRCRAFT TYPE FREQUENCY SELECTOR:** Large and small coaxially mounted knobs select 100KHz and 10KHz steps respectively. Switches click-stopped with a home position facilitate frequency changing without need to view LED'S while driving and provides the sightless amateur with full Braille dial as standard equipment.
- **FULL AUTOMATIC TUNING OF RECEIVER FRONT END:** DC output of PLL fed to varactor diodes in all front end R-F tuned circuits provides full sensitivity and optimum intermodulation rejection over the entire band. **No other amateur unit at any price** has this feature which is found in only the most sophisticated and expensive aircraft and commercial transceivers.
- **TRUE FM:** Not phase modulation — for superb emphasized hi-fi audio quality second to none.
- **FULLY REGULATED INTEGRAL POWER SUPPLIES:** Operating voltage for all circuits, i.e., 12v, 9v and 5v have independently regulated supplies. 12v regulator effective in keeping engine alternator noises out and protects final transistor from overload.
- **MONITOR LAMPS:** 2 LED'S on front panel indicate (1) incoming signal-channel busy, and (2) un-lock condition of phase locked loop.
- **DUPLEX FREQUENCY OFFSET:** 600KHz **plus** or **minus**, 5KHz steps. Plus simplex, any frequency.
- **MODULAR COMMERCIAL GRADE CONSTRUCTION:** 6 unitized modules eliminate stray coupling and facilitate ease of maintenance.
- **ACCESSORY SOCKET:** Fully wired for touch-tone, phone patch, and other accessories.
- **RECEIVE:** .25 uv sensitivity. **15 pole filter** as well as monolithic crystal filter and **automatic tuned** LC circuits provide superior skirt selectivity.
- **AUDIO OUTPUT: 4 WATTS.** Built in speaker.
- **HIGH/LOW POWER OUTPUT:** 15 watts and 1 watt, switch selected. Low power may be adjusted anywhere between 1 watt and 15 watts, fully protected — short or open SWR.
- **PRIORITY CHANNEL:** Instant selection by front panel switch. Diode matrix may be owner re-programmed to any frequency (146.52 provided).
- **DUAL METER:** Provides "S" reading on receive and power out on transmit.
- **OTHER FEATURES:** Dynamic microphone, mobile mount, external speaker jack, and much, much, more. Size: $2\frac{1}{8}$ x $6\frac{1}{2}$ x $7\frac{1}{2}$. All cords, plugs, fuses, mobile mount, microphone hanger, etc., included. Weight: 5 lbs.

NEW! 6 METER FM50-10SXRII

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

NEW! TONE ENCODER/ DECODER

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

SUMMER SPECIAL
FM144-10SXRII
$389.00
VALUE $599.00
Regulated AC/PS
Model FMPS-4R ... $49.00

Touch-Tone Pad
MODEL FMTP-1
... $59.00

Manufactured by one of the world's most distinguished Avionics manufacturers, Kyokuto Denshi Kaisha, Ltd.
First in the world with an all solid state 2 meter FM transceiver.

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East: Sanford Communications, Inc. Colonia, N.J.
West: Consumer Communications, Inc. Seattle, Wash.

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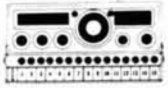
NEW!
FMSC-2 SCANNER
FOR KDK FM-144



14 CHANNEL PROGRAMABLE
INTRODUCTORY PRICE **$99**

*NEW

TEMPO 2020
A BRILLIANT NEW SSB TRANSCEIVER PROVIDING AN UNBEATABLE COMBINATION OF ADVANCED ENGINEERING AND UNIQUE OPERATING FEATURES.
PLEASE WRITE FOR COMPLETE INFORMATION.

NEW AM/FM ANALYZER
SIGMA AF-250L

- INTRODUCTORY PRICE **$199**
- DEVIATION/MODULATION METER – FM: 0-20 KHz, AM: 0-100%.
- SIZE: 5½"(h) x 10¼"(w) x 7¼"(d). WEIGHT: 7 lbs.
- FREQUENCY: 1.8MHz-520MHz

ALSO MODEL AF-251LW WITH BUILT IN 125 WATT CALIBRATED WATT METER & DUMMY LOAD. PRICE **$289**
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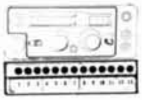
STANDARD
NEW 2 METER FM TRANSCEIVERS
Model SRC-146A
SPECIAL SALE!

• SRC-146A	$298.00
• 4 XTALS 34/94 AND 94/94	N/C
• USA-2 DELUXE BASE CHARGER	$40.00
• PT3644 LEATHER CASE	$10.00
• AT 19 RUBBER ANT AND WHIP	$6.00
• NI-CADS	$30.00
REGULAR	$384.00
OUR PRICE	**$279.00**

*NEW!! TOUCH TONE PAD COMPLETELY WIRED & READY TO PLUG IN $69.00

NEW! 7400 SCANNER
FOR KENWOOD TR-7400A



14 CHANNEL PROGRAMABLE
INTRODUCTORY PRICE **$99**

STANDARD TRANSCEIVER MODEL SRC-146A

*NEW!
NEW!! TOUCH TONE PAD COMPLETELY WIRED & READY TO PLUG IN $69.00
COMPLETE BACK ASSEMBLY

NEW! TWO NEW SCANNERS!
FMSC-1 SCANNER FOR KDK FM 144 **AND**
7400 SCANNER II for Trio-Kenwood TR-7400A

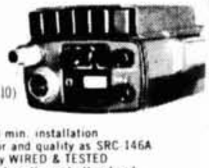
FMSC-1 SCANNER **$169**
7400 SCANNER II **$189**

• Full scan 146 and 147 MHz consecutively or 1 MHz, or any MHz range • Scan rate: 1 MHz/2 seconds (adjustable) • Controls: Scan/Hold, Latch/Delay, 600 KHz offset (off, up, down), program/1 MHz • Simple installation.

NEW— CDR HAM II ROTATORS Reg. $159.95 $125.00

the indispensable BIRD 43 THRULINE WATTMETER
AUTHORIZED BIRD DISTRIBUTOR
Please write for special deal

*NEW



$39.95
NEW!! ADD 5 CHANS (TOTAL 10) TO SRC-146A

- Simple 10 min. installation
- Same color and quality as SRC-146A
- Completely WIRED & TESTED
- ALSO usable with most other hand helds $39.95

YAESU FT-101E TRANSCEIVER'S
Please write for special deal.

SIGMA RF-2000 SWR & POWER METER
INTRODUCTORY PRICE **$29**
CAL. PWR. SCALES 200W/2000W
- FREQ. RANGE 3.5 - 150 MHz
- PLEASE DO NOT CONFUSE THE RF-2000 WITH SIMILAR APPEARING LOWER PRICED UNITS – RF-2000 IS AN INDIVIDUALLY CALIBRATED PROFESSIONAL QUALITY INSTRUMENT –

UNEQUALED AT MANY TIMES THE PRICE.

SIZE: 7"(w) x 2½"(h) x 2½"(d)

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210X-215X AND 350-XL

PLEASE WRITE FOR SPECIAL BONUS AND PACKAGE OFFERS.

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2 DAY AIR SHIPMENT ANYWHERE IN U.S. $35. ALASKA AND HAWAII SLIGHTLY HIGHER.

- FULL BAND COVERAGE 160-10 METERS INCLUDING MARS.
- 2000 WATTS P.E.P. SSB INPUT. 1000 WATTS INPUT CONTINUOUS DUTY, CW, RTTY AND SSTV.
- TWO EIMAC 3-500Z CONSERVATIVELY RATED FINALS.
- ALL MAJOR HV AND OTHER CIRCUIT COMPONENTS MOUNTED ON SINGLE G-10 GLASS PLUG IN BOARD. HAVE A SERVICE PROBLEM? (VERY UNLIKELY) JUST UNPLUG BOARD AND SEND TO US.
- HEAVY DUTY COMMERCIAL GRADE QUALITY AND CONSTRUCTION SECOND TO **NO OTHER UNIT AT ANY PRICE!**
- WEIGHT: 90 lbs. SIZE: 9½"(h) x 16"(w) x 15¾"(d).

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NEW!

- 25 watts output
- 600 kHz and 1 MHz offsets built in
- Please write for complete information and

SPECIAL INTRODUCTORY PACKAGE PRICE

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FMPS-4R	Regulated AC/PS	$49
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FMTP-2	Touch Tone Pad with 10 Number Programable Memory	$99
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FMTD-1	Private Call Decoder for use with and Programed by Any Touch Tone Pad	$129
SC-12A	Audible Tone Encoder Decoder	$119
FMSC-1	Scanner – Random, Any Range	$169
FMSC-2	Scanner – Programable, 14 Channels	$99
MARS-CAP*	Option Kit – Any Frequency, Any Split	$12
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Owners Manual (Extra)		$5.00
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flea market

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HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing). Repeat insertions of hamfest ads pay the non-commercial rate.

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SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N.H. 03048.

QSL'S CATALOG 30¢. RTTY paper 4½ inch roll white, case of 12, $23.00 plus shipping. PO Box 11184, Phoenix, AZ 85061.

QRP TRANSMATCH with Preamp for HW7 Ten-Tec. Send stamp for details to Peter Meacham Associates, 19 Lorreta Road, Waltham Mass. 02154.

TELEVISION CAMERA, new with zoom lense, 2:1 interlace, ALC-$149 or swap for DMM or 2M transceiver, W6DOM, 6017 Majorca Ct., San Jose, CA 95120 (408) 997-0132.

CANADIANS 1,000,000 surplus parts. Bargains galore. Free catalog. Etco-HR. Box 741, Montreal, H3c 2v2.

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PORTAPAK the accessory that makes your mobile really portable. $67.50 and $88.00. Dealer inquires invited. P.O. Box 67, Somers, Wisc. 53171.

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BECKMAN/BERKELEY counter manuels. Most Models. $5.00 each postpaid. W9AMZ, 2009 Hopi, Mount Prospect, Il 60056.

NEW 4-Varactor, 3-transistor UHF TV tuner. Tunes thru 440 MHz for ATV or repeater applications. $7.95 ppd. Workshop, Box 393H, Bethpage, NY 11714.

NEW 60/600 MHz 7-digit portable frequency counter, battery and or electric $299.00. 500 MHz prescaler ÷ 10 or ÷ 100 $59.95. CBSE, Box 1356, Cocoa Beach, Fla. 32931.

PRESCALER BOARD for 650 MHz 11090, Dec. 75 hr. board, instructions, parts source, $3.00. Also CSJ counter boards $15.00 (see 76 Flea Market). RTC Electronics, PO Box 2514, Lincoln, NE 68502.

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

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

HT 100 — Wanted. Also pre 1935 QST's wanted. Al Blank, W1BL, 727 Pine St., Bristol, CT 06010.

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HOMEBREWERS — We are your Oneida Electronics distributor. Free part list, write Tommy, WD4EKJ, Comm-Tech, 104 North Third, Richmond, Ky. 40475.

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

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

SALE: CX-7, $900. 2K-3, $575. GLB Synthesizer — still A kit $75. KAAR 2 mtr, 6 channel, 100 watt mobile, brand new, never used, $150. HT220, 2 channel, pl case, touchtone, charger, $325. Kraft 2 channel RC, $75. Kraft 5 channel RC, $200. Gonset 100 watt 2 mtr mobile amp $75. LA area pick-up. Egbert, 17333 Tramonto, Pacific Palisades, CA 90272.

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

CASH for your ham gear. Beacon Comm., 617-267-1975.

FREE Catalog. Solar Cells, Nicads, Kits, Calculators, Digital Watch Modules, Ultra onics gstrobes, LEDS Transistors, IC's, Unique Components. Chaney's, Box 27038, Denver, Colo. 80227.

WANTED: old keys, Navy,RR etc., also old microphones. K4NBN, Del Popwell, 1946 Sweetbriar, Jacksonville, Fla. 32217.

TRANSMITTER TECHNICIANS — Voice of America has opportunities for qualified technicians at VOA stations in California, North Carolina, and Ohio. Duties include operations/maintenance of high power shortwave transmitters and related facilities on shift basis. Minimum qualifications: 3-years broadcast chief engineer 5 to 50 KW, or 3-years supervisor of operations/maintenance high power military transmitting plant, or equivalent. U.S. citizenship required. Salary $15-19,000. Submit standard government application form, SF-171 to: VOA Personnel Office, Code 05-77, 330 Independence Avenue, S.W., Washington, D.C. 20547. AN EQUAL OPPORTUNITY EMPLOYER.

STOP LOOKING For a good deal on amateur radio equipment — you've found it here — at your amateur radio headquarters in the heart of the Midwest. We are factory-authorized dealers for Kenwood, Drake, Collins, ICOM, Ten-Tec, Atlas, Regency, Tempo, Swan, Midland, Alpha, Standard, Dentron, Hy-Gain, Mosley, Cushcraft, and CDE, plus accessories. Thousands of thrifty hams from coast to coast already know us and we invite you to join them by writting or calling us today for our low quote and trying our personal and friendly Hoosier service. HOOSIER ELECTRONICS, P. O. Box 2001, Terre Haute, Indiana 47802. (812)-238-1456.

CRYSTALS — S.A.S.E. my list. K8LJQ, 355 Mower Rd., Pinckney, Mich. 48169.

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

RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, Ca. 94806.

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flea market

DISCOUNTS — KLM, Larsen, Brimstone, Stereo Systems, Police Monitors, etc. Catalog available. Bankamericard & Mastercharge. 201-962-4695 Narwid Electronics, 61 Bellot Road, Ringwood, N.J. 07456.

TRAVEL PAK QSL KIT — Send call and 25¢; receive your call sample kit in return. Samco, Box 203, Wynantskill, N.Y. 12198.

FREE: COLLINS PM-2 power supply and MM-1 Microphone when you buy my fb condx KWM-2A for $1150. Round emblem, no trades. John, K4GBL, (912) 883-7373.

2-METER CRYSTALS, $3.50 each, for all popular rigs. In stock. Immediate delivery. Send cash or money order, we pay postage. Rolin Distributors, P.O. Box 436, Dunellen, N.J. 08812.

TRANSMITTING TUBES, HV and filament xfmrs, rotary inductors, transmitting capacitors. Send stamp for flyer. T.S. Marinich Electronics, 102 Bell St., Weirton, W. Va. 26062.

SB-200 Mint spare tube $250.00. W2JYL, Long Island, NY 516-261-1824.

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

RYCOM 2174A VLF receiver. 0-420kHz, rack mount; A.C. power, $65.00. JERROLD 602 Sweep Generator, 4-112 MHz; $35.00. Includes manual & UPS surface shipment. Both good condition. M.O. only. D. McMillan K7RKU, PO Box 1042, Astoria, Oregon 97103.

MOBILE IGNITION SHIELDING provides more range with no noise. Available most engines. Many other suppression accessories. Literature, Estes Engineering, 930 Marine Dr., Port Angeles, WA 93862.

P.C. DRILLS carbide #67, 60, 52, $2.25 ea. 3/$6 P.P. R&R Assoc., 12827½ Washington Bl., L.A., Cal. 90066.

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

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

CASH for any Collins unit, 618T, 490T, modules, parts, accessories. Air Ground Electronics, P. O. Box 416, Kearny, N. J. 07032.

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

WANTED: Swan TV2B — Two meter transverter with 50 M.C. IF. Price? Richard Mollentine, WAØKKC, 7139 Hardy, Overland Park, KS 66204.

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

WANTED — Heath SB-620, Hammarlund HQ-140 & HQ-180, Autek QF-1. Will pay cash or swap Rider's Radio, Sams Photofacts, receiving tubes, or magazines. Donald Erickson, 6059 Essex, Riverside, California 92504. 714-687-5910.

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FOR SALE: SB-110A with HP-23A. Six meter SSB/CW transceiver. Excellent condition. U Ship. $250.00. Price is firm. Bob, WB4HJN, Lot #8, Ledbetter Trailer Park, Auburn, AL 36830.

More Details? CHECK—OFF Page 126

More Details? CHECK—OFF Page 126

flea market

RTTY: E-51 AFSK generator — wide/narrow shift — wired/tested — $34.95 — kits available. Stamp for Info. Com — Tech Electronics, P.O. Box 73, Rensselaer, N.Y. 12144.

TX62 2/6M/AM/CW (75W) $55.00. AM1178/432MC Tripler/AMP (250W) HB/PWR $125.00. 1330 Curtis St., Berkeley, CA 94702.

Coming Events

MEMPHIS IS BEAUTIFUL IN OCTOBER! The Memphis ARRL-sponsored Hamfest, bigger and better than the 4,500 who attended last year, will be held at State Technical Institute, Interstate 40 at Macon Road, on Saturday and Sunday October 1 and 2. Demonstrations, displays, MARS meetings, flea market, ladies flea market, too! Hospitality room, informal dinners, XYL entertainment, many outstanding prizes. Dealers and Distributors welcome. Contact Harry Simpson W4SCF, PO Box 27015, Memphis, TN 38127 for further information.

550 AMATEUR RADIO CLUB FLEAMARKET Sunday August 28, 1977, 9-5pm. Oakland American Legion Hall Oak St., Oakland, N.J. Admission $1, Tables $3, Tailgate $2. Talk In — WR2AHD, 146.49-147.49, 146.52. Dealers Inq. invited. Beverages available. For further info. contact 550 A.R.C., PO Box 364, Oakland, N.J. 07436. Or call Rick Anderson, WB2QOQ, 201-684-8569.

BETTER THAN EVER — 1977 EDITION Golden Spread Hamfest and Flea Market- Holiday Inn West Amarillo, Texas Aug. 12, 13 & 14. Six big tech sessions. Commercial exhibits. Family recreation. Two Hospitality Hours. Big pre-registration prize and super Grand Prize, others. $3.00 advance, $4.00 at door. For info. pre-registration packet, P.O. Box 10221, Amarillo, Texas 79106.

MELBOURNE, FLORIDA, SEPTEMBER 10-11. The 12th Annual Melbourne Hamfest will be held Saturday and Sunday, from 9 a.m. to 5 p.m. each day in the airconditioned Melbourne Civic Auditorium located on Hibiscus Boulevard, Donation is $2.50 per person. Full program includes forums, meetings, auction, swap tables, commercial exhibits, awards, prizes, etc. Contact K4HPT, 2749 Herford Road, Melbourne, FL 32935 for swap table reservations. FCC exams on Saturday, donation not needed for exams. Form 610 must be filed with FCC, Room 919, 51 S.W. First Avenue, Miami, FL 33130, not later than August 31, 1977. Hamfest talk-in on 25/85 and 52/52. Sponsored by Platinum Coast Amateur Radio Society. For more info write P.O. Box 1004, Melbourne, FL 32901.

HARRISBURG ARC HAMFEST — Monday July 4. Indian Echo Park, 2 miles West of Hershey. Admission $3.00. Sell no charge.

FLORIDA: The BOLD CITY HAMFEST sponsored by the Jacksonville Range Association will be held at the Jacksonville Beach Auditorium AUGUST 6-7 . . . Vacation at our Hamfest — 'FLORIDA'S FRIENDLIEST' . . . Visit our special 'SOLAR' and 'QRPp' forums. Send request for information and tables to HAMFEST COORDINATOR, Jacksonville Range Association, P.O. BOX 10623, Jacksonville, FL 32207 . . . For Motel reservations call RAMADA INN toll free 1-800-228-2828.

ELMIRA, NY annual Hamfest Saturday, Sept. 24. Chemung Co. Fairgrounds. Flea Market, Dealer Displays, Tech. Talks, Talk-in 10/70 and 146.52. Advance tickets $2. At gate $2.50. Write WA2SMM, 320 West Ave., Elmira, NY 14904.

CEDAR LAKES (W.VA) HAMFEST on August 14. A Flea Market, food and recreation. Talk-in on 146.52 and 31/91. The call is WD8JNU. Space for commercial displays is available. Located 3 miles off I77 at Ripley W.Va., at the site of the Arts and Crafts Fair. For more info contact WB8TJA, P.O. Box 631, Ravenswood, W.Va. 26164 or call 304-273-3190.

INTERNATIONAL VHF FM GUIDE — the finest repeater directory/licensing guide ever. 1977 edition covers Europe, Australia, Canada and South Africa. Only $3 by air from G3UHK, Julian Baldwin, 41 Castle Drive, Maidenhead, SL66DB, England.

NORTH ALABAMA HAMFEST will be held Sunday August 21, at Calhoun Community College in Decatur, Alabama. For information write North Alabama Hamfest Association, P. O. Box 9, Decatur, Alabama 35602.

THE ST. CHARLES Hamfest, August 28, 1977 at Diermanns Lake, 4 miles south of OFallon, Mo. on highway K, will be bigger and better with improved facilities. Prizes, flea market, dealers refreshments and plenty of parking. Tickets still $1.00. For advance tickets or info sase to Dan Corbin, 1512 Sundowner, St. Charles, Mo. 63301.

More Details? CHECK—OFF Page 126

flea market

HAMFESTERS 43rd Annual Picnic and Hamfest. Sunday August 14, 1977, Santa Fe Park, 91st and Wolf Road, Willow Springs, Illinois, Southwest of Chicago. Exhibits for OM's and XYL's, Famous Swappers Row. Tickets at gate $2.00, advance $1.50. For advance tickets send check or money order to Bob Hayes W9KXW, 18931 Cedar Ave., Country Club Hills, Ill. 60477.

FORT WASHINGTON, PA. August 14, 1977. The Mt. Airy VHF Radio Club (the Packrats) are holding their annual family picnic in the Flourtown area of the Ft. Washington State Park on Sunday, August 14, 1977 (rain date 21 August). Talk-in via W3CCX/3 on 52.525, 146.52 and 222.98/224.58 MHz.

THE CARY AMATEUR RADIO CLUB will hold the 5th annual, Mid-Summer Swapfest, Saturday, July 16, 10:00 to 3:00 at the Cary Lions Club Shelter (near Raleigh). Auction at 1:00. No commissions or selling fees. Registrations for prizes. Info: CARC, Box 53, Cary, NC 27511. Talk-in on .28/.88.

CENTRAL STATES VHF SOCIETY, 11th Annual Conference, August 19, 20, 21 at The Breckenridge Inn, Kansas City, Missouri. Information from Orville Burg, Ill. K5VWW.

ANNUAL UTAH HAMFEST and steak fry July 16th, at Saratoga resort. Saratoga is located between Salt Lake City and Provo on Utah Lake. Swap tables, CW contest, homebrew contest, Oscar demo, womens activity, steak fry and many more ham games. On grounds camping is available with lodging near by. Registration is $2.00 for UARC members, $5.00 for non-members and $1.00 for children under twelve. Registration includes choice steak, discount on rides, hot dog for kids, drawings and all other activities. 9:00 am til after dark. Talk-in on 16/76. For more info contact John Dehnel c/o The Utah Amateur Radio Club, 1547 Redondo, Salt Lake City, Utah 84105.

WARREN, OHIO, HAMFEST — August 21, 1977. Moved again! Trumbull K.S.U. Branch Campus on Route 45 at Warren Outerbelt. Best site in our 20 years. Bigger flea market; all close-in parking; parks & lakes nearby. Displays; talk-in; $2 door prize registration. Arrowsigns lead from I-80; I-90; Ohio 5; 11; 45. Details? QSL: Hamfest, Box 809, Warren, Ohio 44483.

WESTERN PENNSYLVANIA — The 40th annual Hamfest of the South Hills Brass Pounders and Modulators will be held on August 7, 1977, from noon till 6:00 P.M. at St. Clair Beach, Upper St. Clair Township, 5 miles south of Mt. Lebanon, on Rte 19. Swap and shop, picnic space and swimming for the family. Mobile check-in on 29.0 and 146.52. Information and pre-registration at $1.50 per ticket ($2.00 at door) from Rich Eckenrode, WB3AAC, 1410 Bellaire Pl., Pittsburgh, Pa. 15226. Vendors must register.

THE "ORIGINAL FM HAMFEST" sponsored by the Fort Wayne Repeater Association, Incorporated of Fort Wayne, Indiana will be held on AUGUST 7. This function has been moved to bigger and better grounds — including 5,400 square feet of exhibit area (air conditioned) and acres of flea market. Location of Hamfest: Allen County Police Department Reserve Center, 3022 Easterday Road, Fort Wayne, Indiana. Talk-in WA9EAU on 16/76, 52/52 and 52.525. Write K9LSB for further information.

CHARLES TOWNE HAMFEST, Charlestown, S. C. July 9 & 10th, 1977. Saturday July 9th the Charles Towne Hospitality Room will be at the Heart of Charleston Motor Inn starting at 7:30 PM. Sunday July 10th the Flea Market and Swapfest will be at the Gaillard Municipal Auditorium starting at 8 AM. Complete details by writing to, Charles Towne Hamfest Committee, Box 4555, Charleston Heights, S.C. 29405.

"GREATER LOUISVILLE HAMFEST is Sunday Sept. 25, 1977 at Kentucky State Fairgrounds with exits off either I-65 or I-264. Indoor exhibitors area and Flea Market air conditioned. Also an outdoor flea market. Ladies Bingo, Meetings and Forums, refreshments available. Admission is $2.00 adults, 12 and under free. Flea market venders pay admission price plus $2.00 per space indoor or $1.00 per space outdoor. For more info or motel/camping contact Denny Schnurr, K4GOU 2415 Concord Dr., Louisville, Ky. 40217 (502-634-0619)."

ILLINOIS. Hamfest, Sunday, Aug. 14 at Santa Fe Park (near Chicago), 91st & Wolf Rd., Willow Springs. Prizes, displays, swapper's row, games, clowns, refreshments. Info from Vincent D. Pronites, WA9EOM, 8131 S. McVicker Ave., Burbank, IL.

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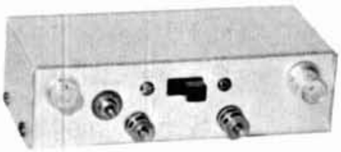

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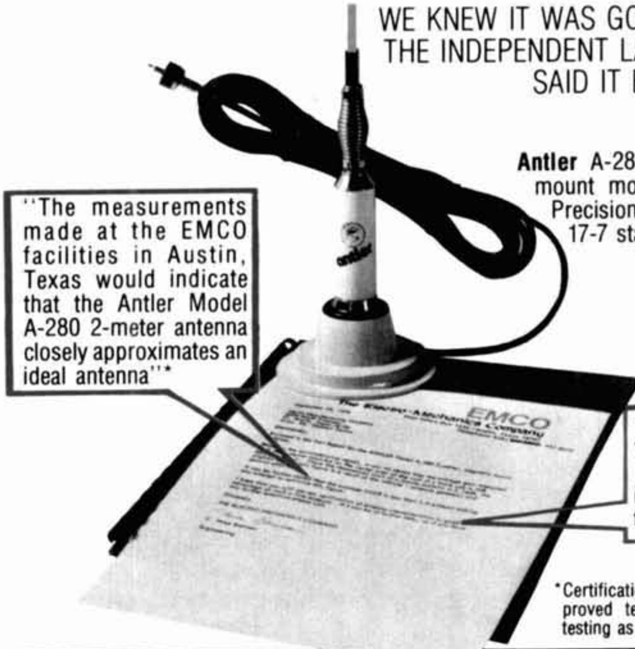

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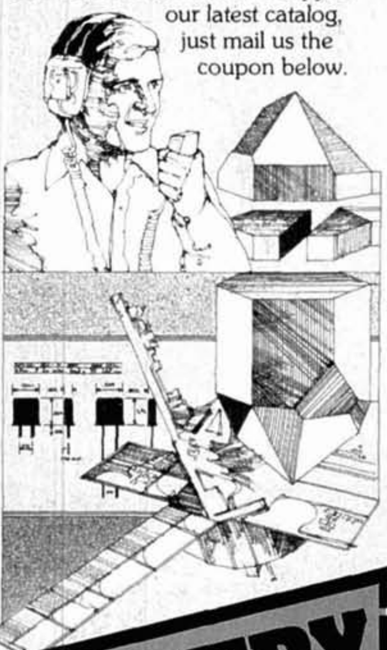

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CD4024	1.50	CD4081	.45	74C164	3.25
CD4025	.25	CD4511	2.50	74C173	2.60
CD4027	.69	CD4518	2.50	74C193	2.75
CD4028	1.65	MC14566	3.00	74C195	2.75
CD4029	2.90	74C00N	.39	MC4044	4.50
CD4030	.65	74C02N	.55	MC14016	.56

LINEAR

LM300H	.80			LM1351N	1.65
LM301H	.35			LM1414N	1.75
LM301CN	.35	78MG	1.75	LM1458C	.65
LM302H	.75	LM370N	1.15	LM1496N	.95
LM304H	1.00	LM373N	3.25	LM1556V	1.85
LM305H	.95	LM377N	4.00	LM2111N	1.95
LM307CN	.35	LM380N	1.39	LM2901N	2.95
LM308H	1.00	LM380CN	1.05	LM3065N	.69
LM308CN	1.00	LM381N	1.79	LM3900N	.55
LM309H	1.10	LM382N	1.79	LM3905N	.60
LM309K	.99	NE501K	8.00	LM3909	1.25
LM310CN	1.15	NE510A	6.00	LM5556N	1.85
LM311H	.90	NE531H	3.00	MC5558V	1.00
LM311N	.90	NE536T	6.00	LM7525N	.90
LM318CN	1.50	NE540L	6.00	LM7535N	1.25
LM319N	1.30	NE550N	.79	8038B	4.95
LM320K-5	1.35	NE555V	.39	LM75450	.49
LM320K-5.2	1.35	NE560B	5.00	75451CN	.39
LM320K-12	1.35	NE561B	5.00	75452CN	.39
LM320K-15	1.35	NE562B	5.00	75453CN	.39
LM320T-5	1.75	NE565H	1.25	75454CN	.39
LM320T-5.2	1.75	NE565N	1.75	75491CN	.79
LM320T-8	1.75	NE566CN	1.25	75492CN	.89
LM320T-12	1.75	NE567H	1.95	75494CN	.89
LM320T-15	1.75	NE567V	1.50	**RCA LINEAR**	
LM320T-18	1.75	LM703CN	.45	CA3013	2.15
LM320T-24	1.75	LM709H	.29	CA3023	2.56
LM323K-5	9.95	LM709N	.29	CA3035	2.48
LM324N	1.80	LM710N	.79	CA3039	1.35
LM339N	1.70	LM711N	.39	CA3046	1.30
LM340K-5	1.95	LM723H	.55	CA3059	3.25
LM340K-6	1.95	LM723H	.55	CA3060	3.25
LM340K-8	1.95	LM733N	1.00	CA3080	.85
LM340K-12	1.95	LM739N	1.00	CA3081	2.00
LM340K-15	1.95	LM741CH	.35	CA3082	2.00
LM340K-18	1.95	LM741CN	.35	CA3083	1.60
LM340K-24	1.95	LM741-14N	.39	CA3086	.85
LM340T-5	1.75	LM747H	.79	CA3089	3.75
LM340T-6	1.75	LM747N	.79	CA3091	10.20
LM340T-8	1.75	LM748H	.39	CA3102	2.95
LM340T-12	1.75	LM748N	.39	CA3123	2.15
LM340T-15	1.75	LM1303N	.90	CA3130	1.39
LM340T-18	1.75	LM1304N	1.19	CA3140	1.25
LM340T-24	1.75	LM1305N	1.40	CA3600	1.75
LM350N	1.00	LM1307N	.85	RC4194	5.95
LM351CN	.65	LM1310N	2.95	RC4195	3.25

74LS00 TTL

74LS00	.29			74LS139	1.95
74LS02	.29			74LS151	1.55
74LS03	.29	74LS74	.49	74LS153	1.89
74LS04	.35	74LS75	.69	74LS157	1.55
74LS05	.35	74LS76	.49	74LS162	2.25
74LS08	.29	74LS83	1.75	74LS163	2.25
74LS10	.29	74LS85	2.49	74LS164	1.95
74LS13	.69	74LS86	.49	74LS175	1.95
74LS14	1.75	74LS90	1.25	74LS181	3.69
74LS20	.29	74LS92	1.25	74LS190	2.85
74LS26	.39	74LS93	1.25	74LS191	2.85
74LS27	.39	74LS95	1.95	74LS192	2.85
74LS28	.39	74LS96	1.89	74LS193	2.85
74LS30	.29	74LS107	.59	74LS194	1.89
74LS32	.39	74LS109	.59	74LS195	1.89
74LS40	.39	74LS112	.59	74LS257	1.75
74LS51	.29	74LS132	1.25	74LS260	.55
74LS55	.29	74LS136	.59	74LS279	.79
74LS73	.49	74LS138	1.89	74LS670	3.95

CLOCK CHIPS

MM5309	6 Digit, BCD Outputs, Reset PIN	$9.95
MM5311	6 Digit, BCD Outputs, 12 or 24 Hour	4.95
MM5312	4 Digit, BCD Outputs, 1 PPS Output	4.95
MM5314	6 Digit, 12 or 24 Hour, 50 or 60 Hz	4.95
MM5316	4 Digit, Alarm, 1 PPS Output	6.95
MM5318	Video Clock Chip, For Use With (MM5841 - $9.95)	9.95
CT7001	6 Digit, Calendar, Alarm, 12 or 24 Hour	5.95

DATA HANDBOOKS

7400	Pin-out & Description of 5400/7400 ICS	$2.95
CMOS	Pin-out & Description of 4000 Series ICS	$2.95
Linear	Pin-out & Functional Description	$2.95

ALL THREE HANDBOOKS $6.95

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T201 Black Bracelet LED $19.95

T237 White w/bracelet $29.95
T236 Yellow w/bracelet $34.95
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- Displays hour, minute, second, month & day
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- Choose LED or LCD styles
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LADIES LED

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- Automatic time and scorekeeping
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Channel F — additional cartridges — $19.95 ea.

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#8113 - Blackjack (1 or 2 players)
#8114 - Spitfire (1 or 2 players)
#8115 - Space War
#8111 - Tic-Tac-Toe/Shooting Gallery/Doodle/Quadra-Doodle

DISCRETE LEDS

.125" dia.			.190" dia.		
XC209	Red	10/$1	XC111	Red	10/$1
XC209	Green	4/$1	XC111	Green	4/$1
XC209	Orange	4/$1	XC111	Yellow	4/$1
XC209	Yellow	4/$1	XC111	Orange	4/$1

.200" dia.			.185" dia.			.200" dia.			.085" dia.
XC22	Red	10/$1	XC526	Red	5/$1	XC556	Red	10/$1	MV50 - Red - 6/$1
XC22	Green	4/$1	XC526	Green	4/$1	XC556	Green	4/$1	
XC22	Yellow	4/$1	XC526	Yellow	4/$1	XC556	Yellow	4/$1	INFRA-RED LED
XC22	Orange	4/$1	XC526	Orange	4/$1	XC556	Orange	4/$1	¼" x ¼" x 1/16"
SSL-22	RT	4/$1	XC526	Clear	4/$1	XC556	Clear	7/$1	Flat 5/$1.00

SPECIAL * — XC556 Red 100/$8.00 1000/$60.00 — SPECIAL *

DISPLAY LEDS

DL707 UL338

TYPE	POLARITY	HT		TYPE	POLARITY	HT	
MAN 1	Common Anode	270	2.95	MAN 3640	Common Cathode-orange	300	1.75
MAN 2	5 x 7 Dot Matrix	300	4.95	MAN 4710	Common Anode-Red	400	1.95
MAN 3	Common Cathode	125	3/1.00	DL701	Common Anode-red ±	300	.99
MAN 4	Common Cathode	187	1.95	DL704	Common Cathode	300	.99
MAN 7	Common Anode	300	1.25	DL707	Common Anode	300	.99
MAN 7G	Common Anode-green	300	1.95	MAN 4740	Common Anode-Red	400	.99
MAN 7Y	Common Anode-yellow	300	1.95	DL741	Common Anode	600	1.50
MAN 52	Common Anode-green	300	.99	DL 747	Common Anode	600	2.25
MAN 64	Common Anode-red	400	.99	DL 750	Common Cathode	600	2.49
MAN 74	Common Cathode	300	1.50	DL 338	Common Cathode	110	.50
MAN 82	Common Anode-yellow	300	.99	FND70	Common Cathode	250	.75
MAN 84	Common Cathode-yellow	300	.99	FND503	Common Cathode	500	1.00
MAN 3620	Common Anode-orange	300	1.75	FND507	Common Anode	500	1.00

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HP 5082-7300 Multi-Digit Series

- ⅛" Ht. • Common Cathode • Dip Package
- 3 to 5 volts @5 mils per segment
- 7 segment Monolithic • Red Display

2 Digit	$ .79
3 Digit	.89
4 Digit	.99
5 Digit	1.19

IC SOLDERTAIL — LOW PROFILE (TIN) SOCKETS

	1-24	25-49	50-100		1-24	25-49	50-100
8 pin	$.17	.16	.15	24 pin	$.38	.37	.36
14 pin	.20	.19	.18	28 pin	.45	.44	.43
16 pin	.22	.21	.20	36 pin	.60	.59	.58
18 pin	.29	.28	.27	40 pin	.63	.62	.61
22 pin	.37	.36	.35				

SOLDERTAIL STANDARD (TIN)

14 pin	$.27	.25	.24	28 pin	$.99	.90	.81
16 pin	.30	.27	.25	36 pin	1.39	1.26	1.15
18 pin	.35	.32	.30	40 pin	1.59	1.45	1.30
24 pin	.49	.45	.42				

SOLDERTAIL STANDARD (GOLD)

8 pin	$.30	.27	.24	24 pin	$.70	.63	.57
14 pin	.35	.32	.29	28 pin	1.10	1.00	.90
16 pin	.38	.35	.32	36 pin	1.75	1.40	1.26
18 pin	.52	.47	.43	40 pin	1.75	1.59	1.45

WIRE WRAP SOCKETS (GOLD) LEVEL #3

10 pin	$.45	.41	.37	24 pin	$1.05	.95	.85
14 pin	.39	.38	.37	28 pin	1.40	1.25	1.10
16 pin	.43	.42	.41	36 pin	1.59	1.45	1.30
18 pin	.75	.68	.62	40 pin	1.75	1.55	1.40

50 PCS. RESISTOR ASSORTMENTS $1.25 PER ASST.

SPECIAL

		10 OHM	12 OHM	15 OHM	18 OHM	22 OHM	
ASST. 1	5 ea.	27 OHM	33 OHM	39 OHM	47 OHM	56 OHM	1/4 WATT 5% — 50 PCS.
		68 OHM	82 OHM	100 OHM	120 OHM	150 OHM	
ASST. 2	5 ea.	180 OHM	220 OHM	270 OHM	330 OHM	390 OHM	1/4 WATT 5% — 50 PCS.
		470 OHM	560 OHM	680 OHM	820 OHM	1K	
ASST. 3	5 ea.	1.2K	1.5K	1.8K	2.2K	2.7K	1/4 WATT 5% — 50 PCS.
		3.3K	3.9K	4.7K	5.6K	6.8K	
ASST. 4	5 ea.	8.2K	10K	12K	15K	18K	1/4 WATT 5% — 50 PCS.
		22K	27K	33K	39K	47K	
ASST. 5	5 ea.	56K	68K	82K	100K	120K	1/4 WATT 5% — 50 PCS.
		150K	180K	220K	270K	330K	
ASST. 6	5 ea.	390K	470K	560K	680K	820K	1/4 WATT 5% — 50 PCS.
		1M	1.2M	1.5M	1.8M	2.2M	
ASST. 7	5 ea.	2.7M	3.3M	3.9M	4.7M	5.6M	1/4 WATT 5% — 50 PCS.

SPECIAL

ASST. 8R Includes Resistor Assortments 1-7 (350 PCS.) **$7.49 ea.**

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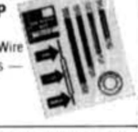
HOBBY-WRAP TOOL-BW-630

- Battery Operated (Size C)
- Weighs ONLY 11 Ounces
- Wraps 30 AWG Wire onto Standard DIP Sockets (.025 inch) Complete with built-in bit and sleeve

$34.95
(batteries not included)

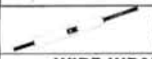
WIRE-WRAP KIT — WK-2-W

WRAP • STRIP • UNWRAP



- Tool for 30 AWG Wire
- Roll of 50 Ft. White or Blue 30 AWG Wire
- 50 pcs. each 1", 2", 3" & 4" lengths — pre-stripped wire.

$11.95



WIRE WRAP TOOL WSU-30

WRAP • STRIP • UNWRAP - $5.95

WIRE WRAP WIRE — 30 AWG

25 ft. min. $1.25 50 ft. $1.95 100 ft. $2.95 1000 ft. $15.00
SPECIFY COLOR — White - Yellow - Red - Green - Blue - Black

Plastic Push Button Switch

- 18 AWG Solid Wire - 5" Long
- .50 (wide) X .60 (high) ⅜-27 Thread
- 8 AMP @ 14 Volt - 1 AMP @ .10 Volt

		1-9	10-Up
J-188-1	Push On-Push Off	.59	.49
J-188-2	Normally Open	.59	.49
J-188-3	Normally Closed	.59	.49

DIP SWITCHES

SPST Slide Action

#206-4	(8 pin dip)	4 switch unit	$1.75 ea.
#206-7	(14 pin dip)	7 switch unit	$1.95 ea.
#206-8	(16 pin dip)	8 switch unit	$2.25 ea.

TV GAME CHIP SET — $18.50

Includes AY-3-8500-1 chip and 2.010 mhz crystal — if purchased separately would cost $21.90.

ZENERS — DIODES — RECTIFIERS

TYPE	VOLTS	W	PRICE	TYPE	VOLTS	W	PRICE
1N746	3.3	400mm	4/1.00	1N4005	600 PIV	1 AMP	10/1.00
1N751A	5.1	400m	4/1.00	1N4006	800 PIV	1 AMP	10/1.00
1N752	5.6	400m	4/1.00	1N4007	1000 PIV	1 AMP	10/1.00
1N753	6.2	400m	4/1.00	1N3600	50	200m	6/1.00
1N754	6.8	400m	4/1.00	1N4148	75	10m	15/1.00
1N959	8.2	400m	8/1.00	1N4154	35	10m	12/1.00
1N965B	15	400m	4/1.00	1N4305	75	25m	20/1.00
1N5232	5.6	500m	.28	1N4734	5.6	1w	.28
1N5234	6.2	500m	.28	1N4735	6.2	1w	.28
1N5235	6.8	500m	.28	1N4736	6.8	1w	.28
1N5236	7.5	500m	.28	1N4738	8.2	1w	.28
1N456	25	40m	6/1.00	1N4742	12	1w	.28
1N458	150	7m	6/1.00	1N4744	15	1w	.28
1N485A	180	10m	6/1.00	1N1183	50 PIV	35 AMP	1.60
1N4001	50 PIV	1 AMP	12/1.00	1N1184	100 PIV	35 AMP	1.70
1N4002	100 PIV	1 AMP	12/1.00	1N1185	150 PIV	35 AMP	1.50
1N4003	200 PIV	1 AMP	12/1.00	1N1186	200 PIV	35 AMP	1.80
1N4004	400 PIV	1 AMP	12/1.00	1N1188	400 PIV	35 AMP	3.00

SCR AND FW BRIDGE RECTIFIERS

C36D	15A @ 400V	SCR	$1.95
C38M	35A @ 200V	SCR	1.95
2N2328	1.6A @ 200V	SCR	.50
MDA 980-1	12A @ 50V	FW BRIDGE REC.	1.95
MDA 980-3	12A @ 200V	FW BRIDGE REC.	1.95

TRANSISTORS

				PN4249	4/$1.00
MPS A05	5/$1.00			PN4250	4/$1.00
MPS A06	5/$1.00	PN3567	3/$1.00	2N4400	4/$1.00
2N2219A	3/$1.00	PN3568	4/$1.00	2N4401	4/$1.00
2N2221	4/$1.00	PN3569	4/$1.00	2N4402	4/$1.00
2N2222A	5/$1.00	2N3704	5/$1.00	2N4403	4/$1.00
2N2369	5/$1.00	2N3705	5/$1.00	2N4409	5/$1.00
2N2369A	4/$1.00	2N3706	5/$1.00	2N5086	4/$1.00
2N2484	4/$1.00	2N3707	5/$1.00	2N5087	4/$1.00
2N2906A	4/$1.00	2N3711	5/$1.00	2N5088	4/$1.00
2N2907A	5/$1.00	2N3724	$.65	2N5089	4/$1.00
2N2925	5/$1.00	2N3725	$1.00	2N5129	5/$1.00
2N3053	2/$1.00	2N3903	5/$1.00	2N5138	5/$1.00
2N3055	$.89	2N3904	4/$1.00	2N5139	5/$1.00
MJE3055	$1.00	2N3905	4/$1.00	2N5209	5/$1.00
MJE2955	$1.25	2N3906	4/$1.00	2N5951	5/$1.00
2N3392	5/$1.00	2N4013	3/$1.00	C106B1SCR	2/$1.00
2N3398	5/$1.00	2N4014	3/$1.00	2N5432	$2.00
		2N4123	10/$1.00		

CAPACITOR CORNER

50 VOLT CERAMIC DISC CAPACITORS

	1-9	10-49	50-100		1-9	10-49	50-100
10 pf	.05	.04	.03	.001µF	.05	.04	.035
22 pf	.05	.04	.03	.0047µF	.05	.04	.035
47 pf	.05	.04	.03	.01µF	.05	.04	.035
100 pf	.05	.04	.03	.022µF	.06	.05	.04
220 pf	.05	.04	.03	.047µF	.06	.05	.04
470 pf	.05	.04	.035	.1µF	.12	.09	.075
100 VOLT MYLAR FILM CAPACITORS							
.001mf	.12	.10	.07	.022mf	.13	.11	.08
.0022	.12	.10	.07	.047mf	.21	.17	.13
.0047mf	.12	.10	.07	.1mf	.27	.23	.17
.01mf	.12	.10	.07	.22mf	.33	.27	.22
+20% DIPPED TANTALUMS (SOLID) CAPACITORS							
.1/35V	.28	.23	.17	1.5/35V	.30	.26	.21
.15/35V	.28	.23	.17	2.2/25V	.31	.27	.22
.22/35V	.28	.23	.17	3.3/25V	.31	.27	.22
.33/35V	.28	.23	.17	4.7/25V	.32	.28	.23
.47/35V	.28	.23	.17	6.8/25V	.36	.31	.25
.68/35V	.28	.23	.17	10/25V	.40	.35	.29
1.0/35V	.28	.23	.17	15/25V	.63	.50	.40
MINIATURE ALUMINUM ELECTROLYTIC CAPACITORS							
Axial Lead				**Radial Lead**			
.47/50V	.15	.13	.10	.47/25V	.15	.13	.10
1.0/50V	.16	.14	.11	.47/50V	.16	.14	.11
3.3/50V	.15	.13	.10	1.0/16V	.15	.13	.10
4.7/25V	.16	.14	.12	1.0/25V	.16	.14	.11
10/25V	.15	.13	.10	1.0/50V	.16	.14	.11
10/50V	.16	.14	.12	4.7/16V	.15	.13	.10
22/25V	.17	.15	.12	4.7/25V	.15	.13	.10
22/50V	.24	.20	.18	4.7/50V	.16	.14	.11
47/25V	.19	.17	.15	10/16V	.14	.12	.09
47/50V	.25	.21	.19	10/25V	.15	.13	.10
100/25V	.24	.20	.18	10/50V	.16	.14	.12
100/50V	.35	.30	.28	47/50V	.24	.21	.19
220/25V	.32	.28	.25	100/16V	.19	.15	.14
220/50V	.45	.41	.38	100/25V	.24	.20	.18
470/25V	.33	.29	.27	100/50V	.35	.30	.28
1000/16V	.55	.50	.45	220/16V	.23	.17	.16
2200/16V	.70	.62	.55	470/25V	.31	.28	.26

More Details? CHECK—OFF Page 126

For the SERIOUS CW/SSB OPERATOR...

500 Hz crystal filter now included
for extra fine cw reception – it's not an option

While operating cw, you may receive with either the 2.1 kHz or the 500 Hz crystal filter —
they are front panel selectable!

Full rated at 250 watts cw input
Ideal for the upgraded Novice/Technician class licenses

the TR-4Cw is a great pile-up puncher for DX chasing and contesting —
it runs more power output than most transceivers run input!

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the New Drake TR-4Cw
sideband/cw transceiver

The serious DX chaser, contester, and traffic handler knows the value of power when the going gets rough in heavy QRM. He also knows the value of sharp receiving selectivity to pull the other guy through the roar.

The NEW Drake TR-4Cw is a great combination of transmit power and receive selectivity. It runs 300 watts PEP input on ssb, and 260 watts input on cw. It is very conservatively rated at these power levels.

When transmitting cw, you're not "locked" into the cw receiving filter as in some rigs. You have your choice of either filter — they're front panel selectable. The cw filter is standard — not an option.

Drake TR-4Cw, Ssb/Cw Transceiver
Drake AC-4, Ac Supply
Drake DC-4, Dc Supply
Drake RV-4C, Remote VFO
Drake 34PNB, Noise Blanker

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Output Power	100 W PEP max
Rx Pre-Amp Gain	20 dB typ
N.F.	2.5 dB typ
Power Supply	115 V A.C.
Size:	10″ x 6″ x 7″

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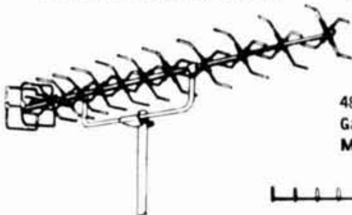

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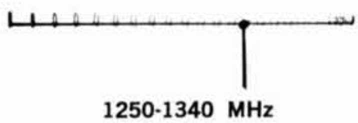

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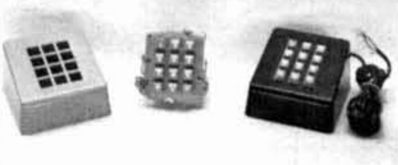

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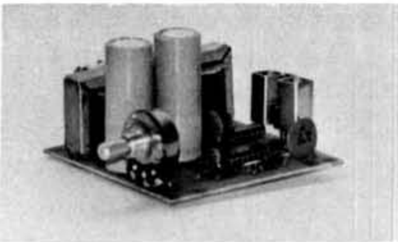

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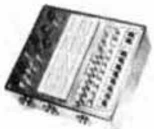

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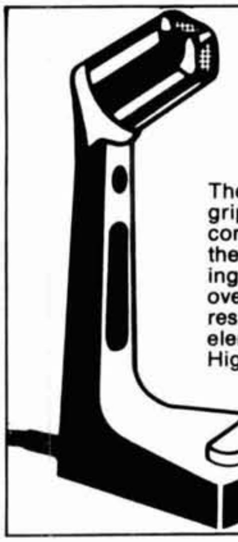

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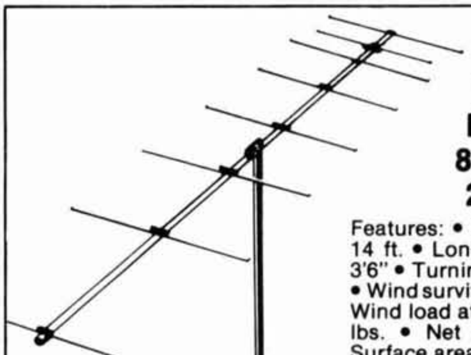

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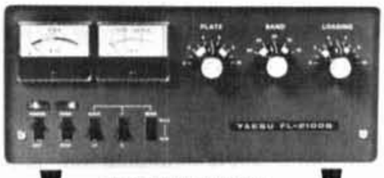

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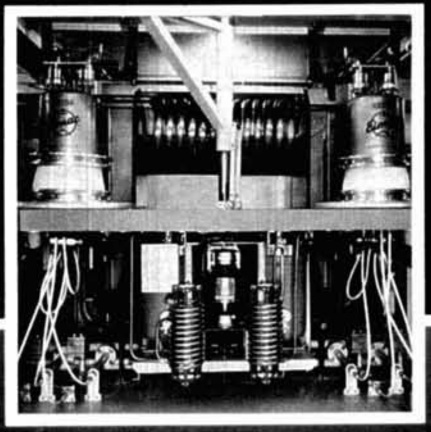

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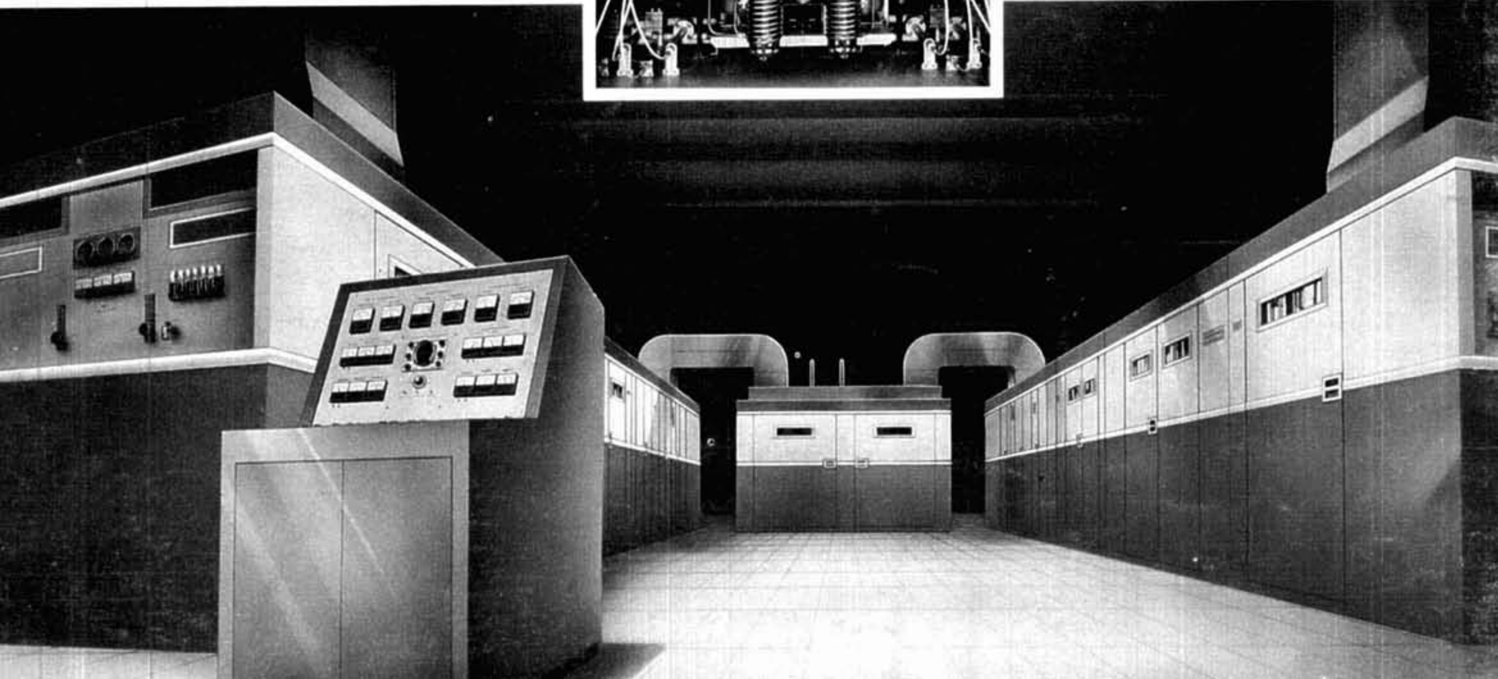

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