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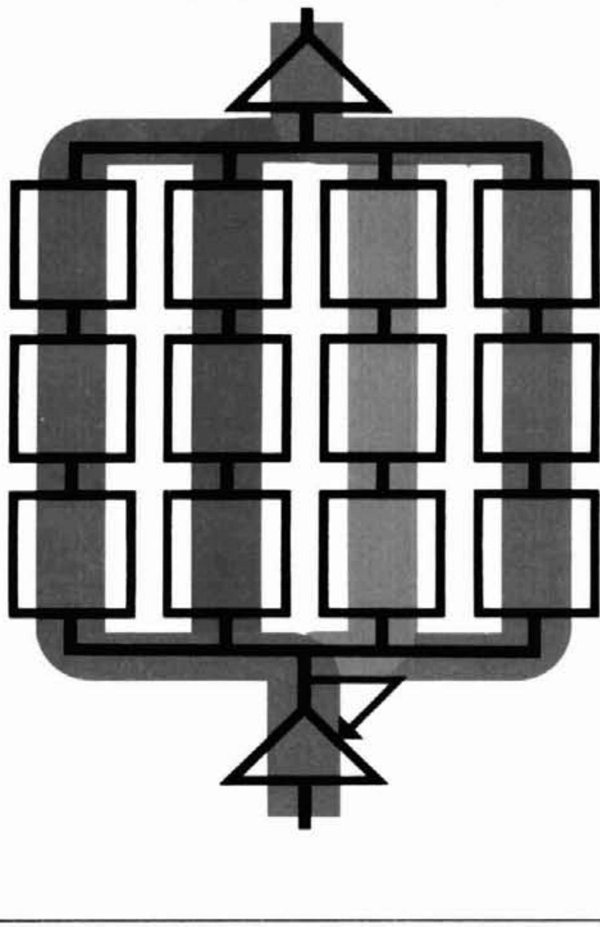
ham radio magazine

SEPTEMBER 1979

split band SPEECH PROCESSOR



*Shown with accessory touch tone pad

PROVEN DEPENDABILITY ...THE TEMPO S-1

KEEPS WORKING WHEN THE GOING GETS TOUGH

By now most of you have heard the same words of praise on the air that we (gratefully) receive over and over. The quality that is built into the S-1 has been attested to by the outstanding performance and dependability of the thousands of units in daily use. The high level of innovative engineering that brought forth the Amateur world's first hand held synthesized radio also designed into this compact beauty exciting performance and features at a very affordable price. A price that also includes a ni-cad battery pack, charger, and a telescoping whip antenna. The optional touchtone pad shown in the illustration adds greatly to its convenience. In addition we offer superior quality 30 and 80 watt solid state matching power amplifiers that give the S-1 the flexibility of operating as a portable, mobile, or base station rig. Remember...the Tempo S-1 is the original and proven 800 channel synthesized hand held transceiver. Don't be fooled by substitutes.

This S-1 was in a motorhome that was totally destroyed by fire. When the owner probed around in the ashes he found his S-1, burned almost beyond recognition. BUT STILL OPERATING. Since then we cleaned it up, replaced the case and controls, tuned it, and now it's back on the air...as good as new.

The Tempo line also features a fine line of extremely compact UHF and VHF pocket receivers. They're low priced, dependable, and available with CTCSS and 2-tone decoders. The Tempo FMT-2 & FMT-42 (UHF) provide excellent mobile communication and features a remote control head for hide-away mounting.
The Tempo FMH-42 (UHF) and the *NEW* FMH-12 and FMH-15 (VHF) micro hand held transceivers provide 6 channel capability, dependability plus many worthwhile features at a low price. FCC type accepted models also available.
Please call or write for complete information. Also available from Tempo dealers throughout the U.S. and abroad.

SPECIFICATIONS

Frequency Coverage	144 to 148 MHz
Channel Spacing	Receive every 5 kHz, transmit Simplex or ±600 kHz
Power Requirements	9.6 VDC
Current Drain	17 ma-standby 500 ma-transmit
Batteries	8 cell ni-cad pack included
Antenna Impedance	50 ohms
Dimensions	40 mm x 62 mm x 165 mm (1.6" x 2.5" x 6.5")
RF Output	Better than 1.5 watts
Sensitivity	Better than .5 microvolts

Price... $349.00

With touch tone pad $399.00

SUPPLIED ACCESSORIES

Telescoping whip antenna, ni-cad battery pack, charger

OPTIONAL ACCESSORIES

Touch tone pad (not installed): **$39** • Tone burst generator: **$29.95** • CTCSS sub-audible tone control: **$29.95** • Rubber flex antenna: **$8** • Leather holster: **$16** • Cigarette lighter plug mobile charging unit: **$6** • Matching 30 watt output 13.8 VDC power amplifier (S30): **$89** • Matching 80 watt output power amplifier (S80): **$149**

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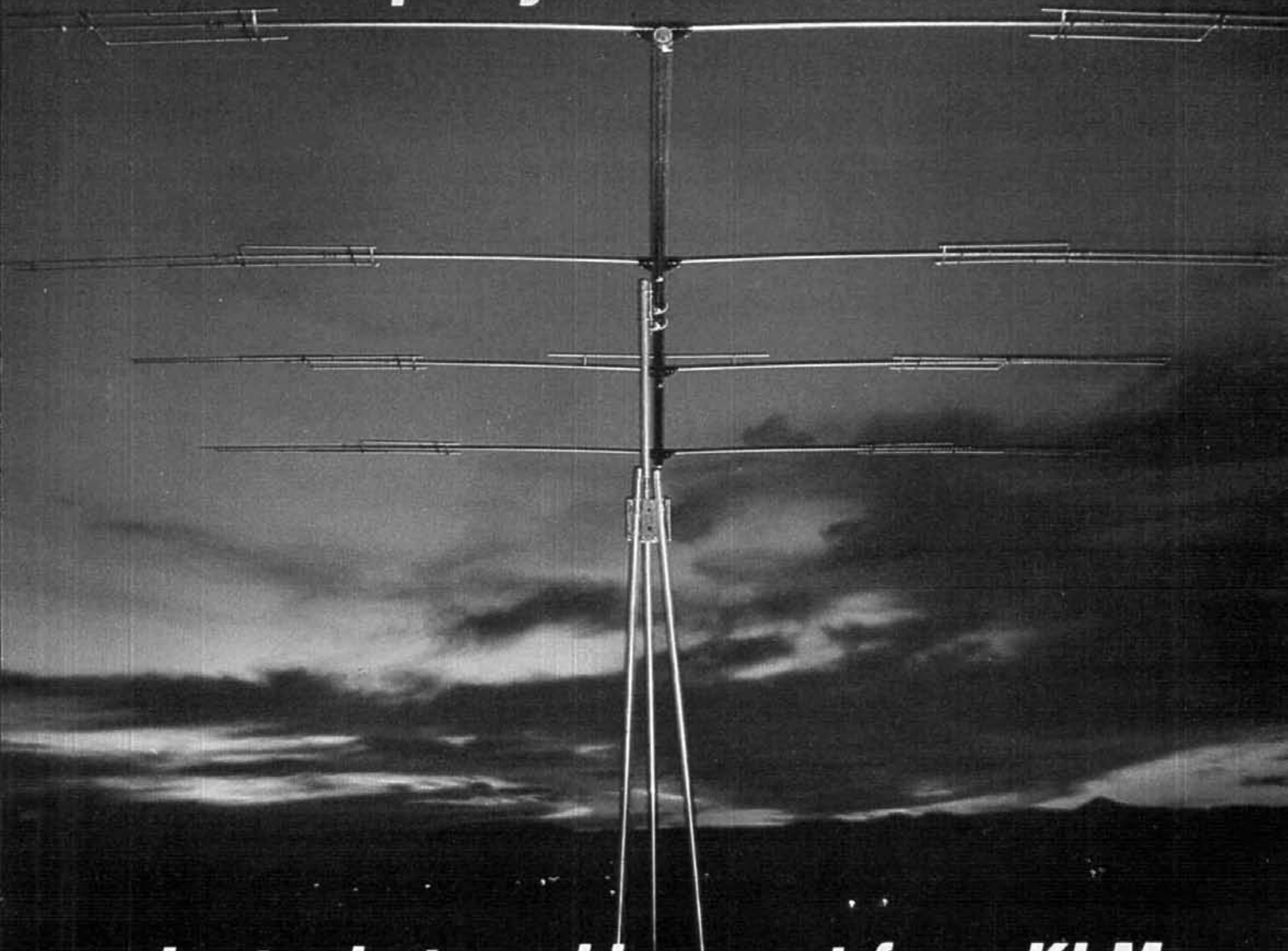

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More Details? CHECK—OFF Page 126

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SEPTEMBER 1979

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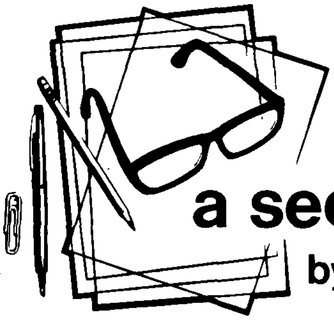


a second look

by Jim Fisk

If you're active on the high-frequency amateur bands, you have probably formed your own idea of what it would be like to operate from a foreign country. You don't need many DX entries in your logbook before you begin to see some trends: power input, types of equipment that are preferred in various places, and the antennas that are the most popular. Have you ever wondered how those same DX operators visualize American radio amateurs?

Writing in a recent issue of *Break-In*, the official journal of the New Zealand Association of Radio Transmitters, Harry Bourne, ZL1OI, provided some of the answers. While making contacts with more than 2500 amateurs in all callsign districts of the United States and Canada on 15 and 20 meters, Harry collected a good deal of interesting data on transmitter input power and antennas. He found, for example, that 13 per cent of the stations used less than 100 watts, 59 per cent used between 100 and 500 watts, and 28 per cent of the operators used more than 500 watts; he also found that the average power input on the 14-MHz band is higher than on 21 MHz.

In the antenna department, ZL1OI's survey showed that 48 per cent of the American amateurs use Yagi beams at heights of 30 to 80 feet (10-25 meters), 21 per cent use verticals (either ground mounted or as elevated ground planes), 13 per cent run quads, often at rather low heights above ground, and 13 per cent depend on half-wave dipoles. The remaining 5 per cent use a variety of antenna types including Zepps, delta loops, vee beams, rhombics and indoor antennas.

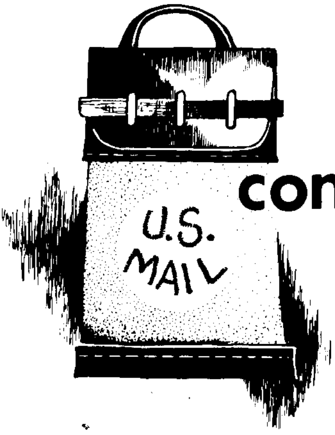
ZL1OI's logbook reveals further interesting results; signal reports, for example, confirm that antennas have a far greater effect on signal strength than transmitter input power — and it is much more effective to improve the antenna than it is to increase power. This will come as no surprise to serious DXers, but it's reassuring to have it confirmed by a DX station. And the excellent propagation conditions we've been experiencing for the past few months have made it possible for amateurs to achieve good DX results with low input powers, especially if they have a good antenna system. One afternoon not too long ago I hooked up with a G3 who was running 150 milliwatts input on CW; he reduced power to 35 mW and we easily exchanged signal reports on ssb. That's roughly 100,000 miles per watt! And just recently I worked 7X2BK on 28 MHz using 200 mW and a 3-element beam.

When propagation conditions are good and the high-frequency bands are as hot as they have been so far this year, directional antennas are not so important for increasing signal strength as they are for reducing interference from directions other than that of the desired station. With a power input of 200 watts, excellent DX results can be obtained with simple vertical or dipole antennas, or single quad or delta loops. If you're unable or unwilling to install a larger or more sophisticated antenna system, you may not be able to crack that big DX pileup on your first call, but with good operating techniques and patience you'll be able to work any station in the world on CW. On phone it's more difficult, but only because the competition is tougher and the interference is horrendous!

If you want to improve your station performance, the message is clear: spend your budget on your antenna *system*, not a linear amplifier, and remember that includes not only the antenna, but the ground system and the transmission line. If you're using inexpensive coaxial line, or cable that's several years old, you may be surprised to find that you can greatly increase your effective radiated power by simply installing RG-213/U or other high-quality coax.

If your budget won't allow a new antenna, try to increase the height of the one you already have; you may be able to double your signal strength by raising your antenna above nearby objects. And if your antenna is ground mounted, increase the number of radials; aluminum electric-fence wire is ideal and costs about a penny a foot. Unless you're already using a Yagi on a 100-foot (30-meter) tower, dollars invested in your antenna system will give you more bang for the buck than dollars spent in any other part of your ham station. Keep that in mind as you get your station ready for the coming DX season. Nearly all the propagation forecasters agree that band conditions this fall and winter will be better than they have been in twenty years — and conditions may not be as good for another twenty!

Jim Fisk, W1HR
editor-in-chief



comments

propagation predictions

Dear HR:

I have been advised that orders for the government publications on *Ionospheric Predictions* cited on page 30 of my article in the April issue of *ham radio* are no longer available from the Superintendent of Documents. I scouted around and, courtesy W9OWZ, discovered that photocopies can be obtained from National Technical Information Service, Post Office Box 1553, Springfield, Virginia 22151. Here are ordering information and prices:

Volume I COM-73-50654 "General Instructions"	$ 3.00
Volume II COM-73-50655 "Sunspot Number = 12"	$11.75
Volume III COM-74-50041 "Sunspot Number = 110"	$11.75
Volume IV COM-74-50042 "Sunspot Number = 160"	$11.75

Henry G. Elwell, Jr., N4UH
Cleveland, North Carolina

voltage-regulator noise

Dear HR:

I very much enjoyed W1HR's article on Gunnplexers in the January issue.

However, I would like to bring something to your readers' attention in reference to the suggested 723 voltage regulator. This regulator employs internal zener regulation, and zeners being inherently noisy, can contribute to system signal-to-noise ratio (SNR) degradation. I have been able to increase the signal-to-noise of a studio-transmitter link (STL) receiver by just short of 3 dB and unmask a VCXO's actual distortion of less than 0.2 per cent by simply by-passing pin 5 of the 723 with a 10 μF capacitor and placing a 47-kilohm resistor between pins 5 and 6. Motorola indicates this addition in one of their application notes; however, its importance is not stressed, nor followed on in other application notes.

I have experienced no such problems with W1HR's suggested Fairchild device, the 78MG. I understand the 78MG regulators are not internally zener regulated. This should be considered by those who are looking toward the ultimate in noise figures, distortion, and SNR.

Dave Clingerman, W6OAL
RF Project Engineer
Moseley Associates, Inc.

Dear HR:

Over the past few years, I have monitored the 160-190 kHz band listening for the large number of stations that are supposed to be running beacons and scheduled transmissions. Since I have never positively identified any of these signals, it appears that either the wrong frequency was being monitored or the signal was too far down into the noise.

A discussion with N6GN concerning this problem resulted in the idea of using the sixth subharmonic of a one megahertz crystal as a standard operating frequency. The resulting 166.666...kHz signal would be very exact since the 5th, 10th, or 15th harmonic of the 1 MHz crystal could be set to be "zero beat" with one of the WWV signals.

I would be very happy to schedule anyone in the San Francisco area on 166.666 kHz.

Dick Bingham, N6HZ
4880 Burnside Rd.
Sebastopol, California 95472

anodize dyes

Dear HR:

The article on anodizing in the January, 1979, issue of *ham radio* mentions several sources of dyes which can be used, including "drugstore" fabric dyes. The trade name dyes, *RIT* are typical of this group, are low cost, and have the reputation of being repeatable. These dyes are used warm, 50-60°C; the dyed surface is then sealed by boiling.

Bob Haviland, W4MB
Daytona Beach, Florida

note of acknowledgment

The April, 1979, issue contained an article entitled "The Jammer Problem: Some Interesting Solutions," page 56. This article was adapted from "A Contribution to the Mathematical Theory of Big Game Hunting," by H. Petard, Princeton, New Jersey, which originally appeared in *American Mathematical Monthly* (1938). The article has since been reprinted in *A Random Walk in Science* (1973), compiled by R. L. Weber and edited by E. Mendoza.

We thought that Petard's article on big game hunting would make an interesting basis for an adaptation geared to the very real problem of intentional interference in the Amateur bands. Thanks to Jim Kirkpatrick, WB7BUP, for the background information on the original piece. **Editor**

KENWOOD

TECH TALK 8C79

AT-120 and AT-180

Antenna Tuners recommended for TS-120S and TS-180S All Solid-State HF Transceivers.

AT-120

AT-180

Why is the use of an antenna tuner so much more important with the TS-120S and TS-180S all solid-state HF rigs than with transceivers having tube-type final amplifiers?

Tube-type final amplifiers generally handle a broader range of load impedance than a transistor final. However, RF power into the antenna system will decrease with an increasing impedance mismatch, and tube life may even be shortened if the mismatch is extreme. Transistor final amplifiers, on the other hand, require a 50-ohm nonreactive load for efficient power transfer and are not very tolerant of high SWR. Therefore, protection circuits are used in the TS-120S and TS-180S to reduce RF power output significantly under high-SWR conditions, thus preventing damage to the solid-state devices.

In the TS-120S, an SWR detection circuit detects reflected-wave voltage, which is then amplified and applied to the ALC circuit as a protective voltage to control power output. Thus, as SWR increases, RF power output decreases continuously.

In the TS-180S, the final amplifier functions normally up to an SWR of 3:1, at which point the protection circuit drops RF power output significantly.

Using an antenna tuner such as the AT-120 to match the TS-120S or the AT-180 to match the TS-180S will lower the reflected power at the transceiver to avoid detection by the protection circuit, thus enabling the transceivers to produce full RF power output and even with rigs which have tube final amplifiers, we recommend an antenna tuner (such as the AT-200 to match the TS-520 or TS-820 Series) for optimum coupling to antennas with high SWRs.

A major advantage of using an all solid-state rig such as the TS-120S or TS-180S is the elimination of final-amplifier tuning and loading. It's great to be able to switch bands, dial up any frequency, and transmit immediately, especially when operating mobile or in a contest or chasing DX. Isn't this advantage lost if an antenna tuner has to be used?

We recommend using an antenna that has a low SWR (below 1.5:1) and that presents a proper impedance match (50-ohms) to the transceiver. Then the full advantages of using an all solid-state rig can be realized. Furthermore, the antenna will be more efficient, and power will not be reflected back to an antenna tuner.

However, many antennas are not broad enough to cover an entire band, and may have an SWR below 1.5:1 in just a portion of the band. The antenna may be cut for resonance in the middle of the portion of the band that is mostly used. When operating outside this portion, where SWR exceeds 1.5:1, the antenna tuner should be switched in.

Therefore, with a well-designed antenna, the antenna tuner may be switched out for most operating, and the full advantage of using a no-tune all solid-state rig may be realized. But for those occasions when operating in the band portions where the antenna is not resonant and reactance increases or when, for some reason, the antenna develops a high SWR or a poorly matched antenna is used, the antenna tuner should be switched in to obtain full RF power output.

During those occasions when the antenna tuner is needed with an all solid-state rig, it would probably be advantageous with a tube-type rig also for optimum power transfer to the antenna system. With a tube-type final, plate tuning and loading adjustments would be required in addition to adjusting the antenna tuner. With an all solid-state transceiver, only the antenna tuner would need adjusting during those occasions when it is required.

What are the primary features of the AT-120 and AT-180 antenna tuners?

The AT-120 antenna tuner is very compact (only 6 inches wide, 2-3/8 inches high and 6-1/4 inches deep) - perfect for mobile mounting with the mounting bracket provided - and operates on 80 through 10 meters. It consists of an antenna coupler and an SWR meter (which can be illuminated). Although much smaller, it complements the appearance of the TS-120S.

The bandswitch has a "THROUGH" position for switching the AT-120 out of the circuit. Input impedance (to the transceiver) is 50-ohms and output impedance (to the antenna system) covers 20 to 300 ohms, unbalanced. It handles 150 watts (120 watts on 80 meters). The SWR meter measures from 1.0:1 to 10.0:1.

The AT-180 antenna tuner matches the TS-180S (same height) and consists of a through-line watt and SWR meter, antenna selector switch, and, of course, an antenna coupler.

It operates on 160 through 10 meters, with a 50-ohm input impedance and an output impedance of 10 to 500 ohms (10 to 400 ohms on 160 meters), unbalanced. Switches allow up to 20 or 200 watts of forward or reflected power to be measured. (It is not intended for use at the output of a linear amplifier.) UHF-type connectors are provided for the input, two antenna outputs, and a dummy load, and a standoff connector is provided for a wire feedline.

With both tuners, the "R TUNE" (for resistance component) and "X TUNE" (for reactance component) controls are adjusted alternately with a CW carrier applied until minimum SWR or reflected power is obtained.

presstop

UNUSUAL PROPAGATION EFFECTS should result in September when an Atlas-Centaur rocket is launched from the Kennedy Space Flight Center in Florida. The exhaust gases from the giant rocket are expected to burn a large hole in the ionosphere's F region, and a group of Amateur Radio experimenters are planning to observe the effects on propagation during, and for several hours following, the launch some time in September.

The Resulting Hole in the ionosphere could be as large as 500 km across, starting from the north Florida coast and extending eastward along the launch trajectory. The paths from Puerto Rico, the Virgin Islands and the rest of the northern West Indies to the U.S. East Coast and Bermuda offer the best chances for HF observations.

For Specific Details concerning participation in the experiments, which are attracting much scientific interest, contact W1JR or W1BZT. A special certificate will be issued to all contributing participants.

FIRST 432-MHz HAWAII to mainland U.S. contact was made in July when KH6HME worked WB6NMT on ssb. Three more stations, W6YDF, WB6ESQ, and WB6WLR, also made the grade, with signals generally above the noise level.

FIRST WAS ON 432 MHz has just been achieved by WØYZS, with a Wyoming contact with WA7DKZ. Mike, who's been working toward this moment for 10 years, caught most of his recent states via moonbounce. Congratulations on an outstanding achievement!

A 6-METER HAWAII TO EAST COAST opening in late June provided many 50-MHz enthusiasts, particularly those in New England, with their first KH6 contacts; during the opening KH6IAA worked a large number of 1s, 2s, 3s, and 4s. Openings in July put KG4HC, KV4FZ, and KP4Q into a number of U.S. logbooks, and in early July JE2NQC worked VE7XF and heard several W6 and W7 calls.

A British Isles 6-Meter Beacon has been proposed by the Radio Society of Great Britain. It's to be located on the island of Angesea, off the northwest coast of Wales, and there's hope to have it operational by year's end.

A CALIFORNIA COURT DECISION against a CB operator has established a precedent that could work against Amateurs as well, and serves as a warning to carefully examine all restrictions before renting or buying a new home if you plan antennas. San Diego CB operator Jerry Lee Dunn was sued by a neighbor for violating the convenants, conditions, and restrictions (CCRs) of their subdivision, which absolutely prohibited outside antennas for any purpose. Dunn fought the suit on the grounds that the CCRs violated his First Amendment rights to free speech and was thus unenforceable.

In Its Decision Upholding the antenna prohibition, the Court of Appeals, Fourth District of California, found that Dunn's right to speak on the air was not itself restricted, and also required that he pay the other side's legal fees.

AMSAT HAS RECEIVED APPROVAL of the donation of the rocket motor that will boost the first Phase III spacecraft into a higher eliptical orbit. The launch date for Phase III-A still stands at March 5, 1980, with spacecraft delivery required by December 3rd in France for tests at the European Space Agency facilities. The general beacon will be on 145.810 MHz, plus or minus 2 kHz; the engineering beacon will be on 145.990 MHz, plus or minus 2 kHz.

AMSAT Reports That OSCAR 7 is available for use regardless of mode. It's been hard to keep it in scheduled modes, so which one you use it in, A or B, will depend on which mode it's jumped to when you find it. Wednesdays, of course, are still reserved for experiment days.

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

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

EXPERTISE ON BIOLOGICAL EFFECTS OF RF is being sought by ARRL for a new ad hoc committee which will help prepare League comments on the FCC Notice of Inquiry (Docket 79-144, August Presstop), and later provide on-going advice in this increasingly sensitive area. Qualified amateurs interested in serving on this important committee should contact ARRL president Harry Dannals, W2HD.

W2PV's YAGI ANTENNA SERIES, originally slated to begin in this issue of ham radio, has been slightly delayed. Look for the first section in the December issue; it is well worth waiting for.

New OMNI/SERIES B Filters The Crowd

The new OMNI/SERIES B makes today's bands seem less crowded. By offering a new i-f selection that provides up to 16 poles of filtering for superior selectivity. And a new Notch Filter to remove QRM. No other amateur transceiver we know of out-performs it.

NEW I-F RESPONSE SELECTION. OMNI comes equipped with an excellent 8-pole 2.4 kHz crystal ladder i-f filter which is highly satisfactory in normal conditions. But when the going gets rough, the new OMNI/SERIES B, with optional filters installed, provides two additional special purpose i-f responses.

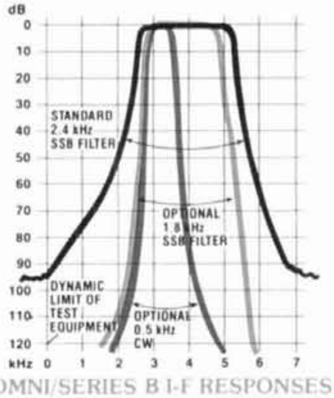
The 1.8 kHz crystal ladder filter transforms an unreadable SSB signal in heavy QRM into one that gets the message through. The 0.5 kHz 8-pole filter provides extremely steep and deep skirts to the CW passband window which effectively blocks out even the very strong adjacent signals.

Both of these filters can be front-panel switched in series with the standard filter to provide up to **16 poles** of filtering for near-ultimate selectivity. In addition, the standard CW active audio filters have three bandwidths (450, 300, and 150 Hz) to give even further attenuation to adjacent signals. In effect, OMNI/SERIES B has six selectivity curves—three for SSB and three for CW. That's true state-of-the-art selectivity.

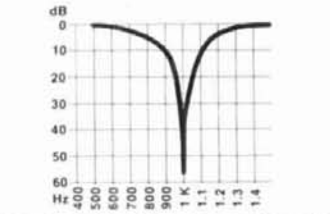
NEW NOTCH FILTER. A variable frequency notch filter in OMNI/SERIES B is placed inside the AGC loop to eliminate interfering carriers and CW signals without affecting received signals. Attenuation is more than 8 "S" units (over 50 db) for any frequency between 0.2 kHz and 3.5 kHz.

OMNI/SERIES B RETAINS ALL THE FEATURES THAT MADE IT FAMOUS.

All solid-state; 160-10 meters plus convertible 10 MHz and AUX band positions; **Broadband design** for band changing without tuneup, without danger;



OMNI/SERIES B I-F RESPONSES WITH STANDARD AND OPTIONAL FILTERS.



NOTCH FILTER PERFORMANCE ADJUSTED TO 1 kHz POINT.

TEN-TEC, INC. SEVIERVILLE, TENNESSEE 37862
EXPORT 5715 LINCOLN AVE., CHICAGO, ILL. 60646

Choice of readouts —OMNI-A for analog dial or OMNI-D for digital dial; **Built-in VOX and PTT** facilities; **Selectable Break-in,** instant or delayed receiver muting; **Dual-Range Receiver Offset Tuning,** ±5 kHz or ±0.5 kHz; **Wide Overload Capabilities,** dynamic range typically exceeds 90 dB and a PIN diode switched 18 dB attenuator is also included; **Phone Patch Interface Jacks; Adjustable ALC; Adjustable Sidetone; Exceptional Sensitivity; 200 Watts** input to final with full warranty on final transistors for first year, pro-rata for 5 years; **100% Duty Cycle** for RTTY, SSTV or sustained hard usage; **12 VDC Circuitry** for mobile use, external supplies for 117/220 VAC operation; **Front Panel Microphone and Key Jacks; Built-in 25 kHz Calibrator** in analog dial model; **Zero-Beat Switch; "S"/SWR Meter; Dual Speakers; Plug-In Circuit Boards; Functional Styling,** black textured vinyl over aluminum "clamshell" case, complementary nonreflective warm dark metal front panel; **Complete Shielding; Easier-to-use size:** 5¾"h x 14¼"w x 14"d; **Full Options:** Model 645 Keyer $85; Model 243 Remote VFO $139; Model 252MO matching AC power supply $139; Model 248 Noise Blanker $49; Model 217 500 Hz 8-pole Crystal Ladder CW Filter $55; Model 218 1.8 kHz 8-pole Crystal Ladder SSB Filter $55;

OMNI owners note: Your OMNI can be converted to a SERIES B model at the factory for just $50 (plus $5 for packing and shipping). The notch filter replaces your present squelch control and provision is made for the two additional optional filters; a partial panel with new nomenclature is provided. Contact us for details.

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

Experience the uncrowded world of OMNI/SERIES B. See your TEN-TEC dealer or write for full details.

Discover the unrivaled ASTRO 102BX with Dual PTO's

Plus IF Passband Tuning/Full Band Coverage, Speech Processing, 235 Watts PEP All Bands and Much More.

4-Function Meter
reads "S" units in receive, and selects forward power, reflected power, or ALC in transmit.

Visual Display of Passband Tuning
A series of 8 LED's indicate the equivalent band width and position of the passband from 0.6 to 2.7 KHz as the passband tuning knob is rotated.

300Hz Crystal CW Filter
Cascaded with the passband filters and tunable through the passband. Combined with notch filter yields unrivaled CW performance.

Full or Semi CW Break-in
A must for the avid CW operator!

2 Position CW Wave Shaping
Your choice of soft or hard CW rise and decay time!

Variable AGC Decay
Continuously variable to suit your preference.

Passband Tuning with 16 Pole Filters
Two 8-pole crystal filters in cascade provides a 1.4:1 shape factor at -100dB! The ultimate in selectivity.

Tunable Notch Filter
Provides the ability to "notch out" QRM. When used in conjunction with the passband tuning, provides the ultimate in removing interference.

Dual PTO's
Two independent high stability PTO's provide true split band operation. The digital counter reads the PTO selected, or in split band reads the PTO used for receive, then switches to the transmit frequency selected when the mic is keyed.

All Band Coverage using PLL and Synthesizer for Band Selection
Full coverage of 160 meters through all of 10 meters in 9 ranges . . . diode matrix programmed for now — or for the future!

The ASTRO 102BX with it's companion PSU-6 Power Supply, 1500Z Linear Amplifier and ST-2A Antenna Tuner provides a matched and highly efficient 1500 watt PEP or 1000 watt CW complete station to be complemented by a great Swan antenna.

All Solid State Quality American Construction

ASTRO 102BX

SWAN ELECTRONICS

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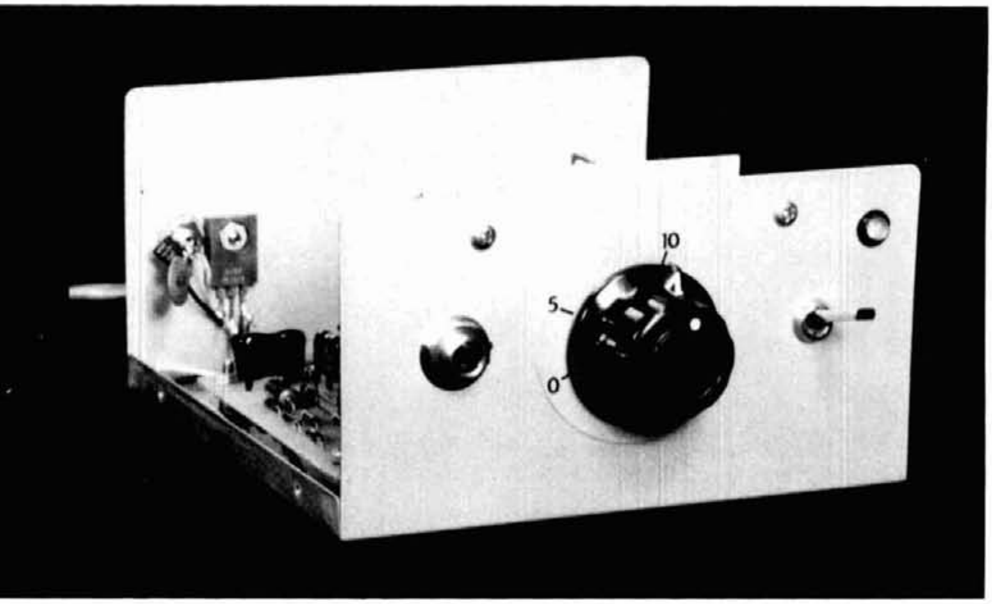

split-band speech processor

Design and construction details of a split-band audio speech processor that features up to 15 dB of clipping and low distortion

Speech processing, especially for SSB, can be a relatively inexpensive means of improving the effective "talk power" of a voice modulated transmitter. Much has been written about various devices and methods that can be used to gain this increase in effective talk power. The devices used have ranged from simple audio compressors to rf envelope clipper-filters. All of these devices attempt to reduce the peak-to-average ratio of the speech or rf waveforms, thereby overcoming the peak power limitations of the transmitter. Generally, the degree of improvement is proportional to the complexity of the processing method; the simpler circuits offer minimal improvement while the more complex effect substantial improvement.

This article will not attempt to present all the theory involved in speech processing; however, the interested reader is referred to excellent articles by Fisk,[1,2] Kirkwood,[3] Moxon,[4] and Schreuer[5] for more detailed overviews of the subject.

Until recently, rf envelope clipping has generally been accepted as the most effective SSB processing method. Distortion products are small, generally consisting only of intermodulation products. The primary disadvantage of rf processing is the circuit complexity involved, and the necessity of modifying the associated transmitter. When modifying the transmitter is out of the question, a processor using the audio-SSB-audio (Comdel) approach can be used. In this method, an SSB signal is generated, peak limited (clipped), filtered, and then demodulated back to an audio signal which then modulates the transmitter.

My initial efforts were directed toward designing and building a unit of this type. A breadboard model was constructed and evaluated under laboratory conditions. Performance was very good, and distortion was held to under 10 per cent at 20 dB of clipping. The circuit was, however, excessively complex. It required an audio preamplifier, two balanced modulators, an oscillator, a clipper, an rf amplifier, and an expensive mechanical or crystal filter.

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At this point, Jim Metzger, W7TKR, suggested that I try the split-band approach. He had done some work with the process with considerable success and Fisk[2] had written in glowing terms about a similar unit available commercially from Maximilian Associates. This was inducement enough to build a breadboard model for evaluation.

basic circuit

Fig. 1 is a simplified block diagram of the split band clipper. The input signal is applied to an agc-controlled preamplifier which then drives the first set of bandpass filters (BPFs). The filters split the audio spectrum into four narrow bands which are then clipped and directed into the second set of BPFs, where the harmonics generated by the clipping process are filtered off. These filtered signals then go to the combiner stage where they are reassembled into the desired output.

Input amplifier. The design of the input amplifier is not particularly critical. The gain required will depend on the output amplitude of the source, the gain (if any) of the BPFs, and the limiting threshold of the clipper stages. If a very low output microphone is used, low noise may be of some importance. If, as in my case, active bandpass filters are used, the amplifier will also have to exhibit low output impedance. Automatic gain control is also desirable, as it helps maintain a high average clipping level, which in turn insures maximum talk power improvement.

Bandpass filters. As pointed out by Fisk, the optimum design for BPFs is a compromise between several conflicting requirements. Overshoot or ringing due to the near squarewave input from the clipper must be minimized, skirt selectivity should be good, and phase shift through the passband must be smooth and predictable. The latter point becomes important when the design of the combiner is considered, as will be seen later. Other very important factors to be considered are circuit complexity and reproducibility.

After pondering all of the above points, I decided on a two-pole Butterworth active filter. The Butterworth is not optimum when considering only impulse response and phase shift; however, when used in a low-Q configuration, it is a good compromise between filters with these attributes and those possessing superior skirt selectivity.

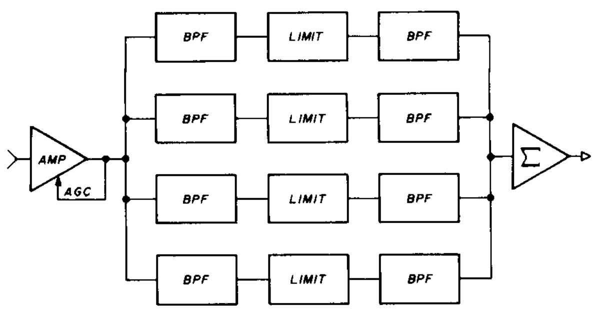


fig. 1. Block diagram of a split-band audio speech processor.

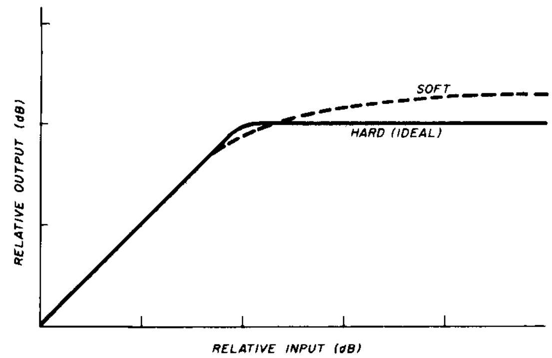


fig. 2. Comparison of "hard" vs "soft" limiting. Soft limiting is undesirable because of the uncertainty of the threshold point, making it hard to maintain constant output from the processor.

The final circuit is configured as a multiple-feedback type.[6] These filters are relatively insensitive to component variations, allowing the use of 5 per cent tolerance components and inexpensive operational amplifiers. Detailed design data for the selection of center frequency, gain, and Q will be given later.

Peak clipper. The clipper may seem to be one of the least critical parts of the circuit, but, in fact, its requirements are quite stringent. One of the most important factors in the performance of the clipper is that of clipping symmetry. Perfect symmetry insures that only odd harmonics are generated; second-order products would be too much for the two-pole filters to handle. An important point is that the only place clipping should occur is in the clipper. Clipping or limiting elsewhere in the circuit cannot be easily controlled and must be avoided. This may seem easy to do, but if the clipping threshold is too high, limiting may occur in a preceding stage when large amounts of clipping are in use. For example, if a clipping threshold of one volt is used and 20 dB of peak clipping is desired, the preceding stage must be able to have an output voltage swing of 20 volts peak-to-peak. If this stage is running off a single 12-volt power supply, this will of course be impossible.

Another important aspect is that of how "hard" the limiting is. Many of the circuits initially examined, which included limiting differential amplifiers, shunt-diode clippers, and operational amplifiers with shunt diode feedback, had rather "soft" limiting characteristics. That is, the threshold was ill-defined and the slope of the transfer function continued to change over a wide range of input levels. **Fig. 2** graphically shows the difference between hard and soft limiting.

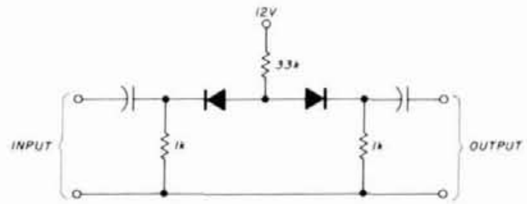


fig. 3. Circuit diagram of the amplitude limiter (clipper) used in the final design. With the resistor values shown, the output will be limited to approximately 300 mV p-p.

Soft limiting is undesirable because it makes it difficult to maintain a constant peak output level.

The circuit finally selected for this application, as best satisfying the above requirements as well as using a minimum of parts, is shown in **fig. 3**. This will be recognized as a variation of the old series automatic noise limiter used in receivers. By suitable selection of resistor values and bias voltage, the clipping threshold may be adjusted over a wide range.

The performance of this circuit is demonstrated in **fig. 4**. This is a multiple-exposure oscilloscope photograph taken of the output of the clipper. The inner, near sinusoidal, trace was obtained by increasing the input signal until a 3-dB increase caused only a 2-dB change in output. This point was defined as the clipping threshold. The middle trace represents a further input increase of 4 dB, and the outermost trace was obtained with a total input overdrive of 15 dB. The photograph shows the nearly flat peak output and the exceptional symmetry. A further test of symmetry was made by examining the frequency spectrum of the clipper output with a Hewlett-Packard 302A wave analyzer. With 15 dB of clipping, the second harmonic remained more than 40 dB below the fundamental output.

Combiner. The combiner has the job of taking the four BPF outputs and putting them back together again while maintaining their original phase relationships. Improper phasing will result in excessive passband ripple being generated. As described by Fisk, the Maximilian unit incorporates phase shift networks before the combiner to compensate for the phase shifts through the BPFs. As will be shown later, these networks can be eliminated by the judicious selection of filter characteristics and the use of a simple summing and differencing amplifier.

circuit description

Fig. 5 is the complete schematic of the system. The input is applied to Q1, an FET source follower, used to match high impedance microphones. The follower output drives U1, a Plessey SL1626 gain-controlled amplifier. This IC maintains a nearly constant output of slightly less than 100 mV RMS over an input range of 1 to 100 mV.

The SL1626 is used as recommended by the data sheet, except for the addition of R6 and C10, which are necessary to suppress a high-frequency oscillation. R4 lowers the sensitivity about 20 dB and may be unnecessary in some applications. Front panel adjustment of the clipping level is possible via R7.

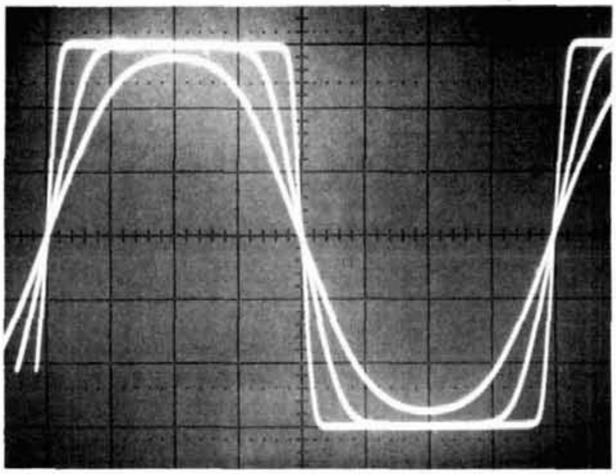
Amplifier U2A, one section of an LM324, develops a small amount of additional gain and serves as a low-impedance source to the following BPFs. The resistors used on the outputs of all the LM324s are necessary to eliminate cross-over distortion.[7]

All of the bandpass filters are operated at the same gain and Q; only the center frequency (f_o) differs from channel to channel. For simplicity, all capacitors are of the same value; the center frequency is adjusted by choice of resistor values. Using the given values, the overall frequency response will be approximately 350 to 3000 Hz at −6 dB, with no greater than 3 dB of passband ripple. If other cutoff frequencies are desired, **appendix 1** gives the equations necessary to calculate new values of f_o and Q. **Appendix 2** gives the equations for calculating the parts values for the individual filters.

The clipping stages, as described earlier, use a pair of forward-biased diodes. With the bias resistor values shown, the clipped output will limit at about 300 mV p-p. The shunt-bias resistor values are kept low enough to insure that the input impedances of the second BPFs remain fairly constant even when the clipping diodes turn off.

The second set of BPFs are identical to the first. Their outputs are combined in another section of an LM324, which delivers the system output through a resistive divider. By adjusting the resistor values, the output amplitude can be set approximately the same as that of the microphone, allowing the clipper to be

fig. 4. Performance of the clipper stage shown in fig. 3. The sinusoidal trace was made at the threshold point (1-dB clipping). The middle trace shows 5 dB of clipping, and the outer trace was made with 15 dB of clipping. The vertical sensitivity is 50 mV/div.



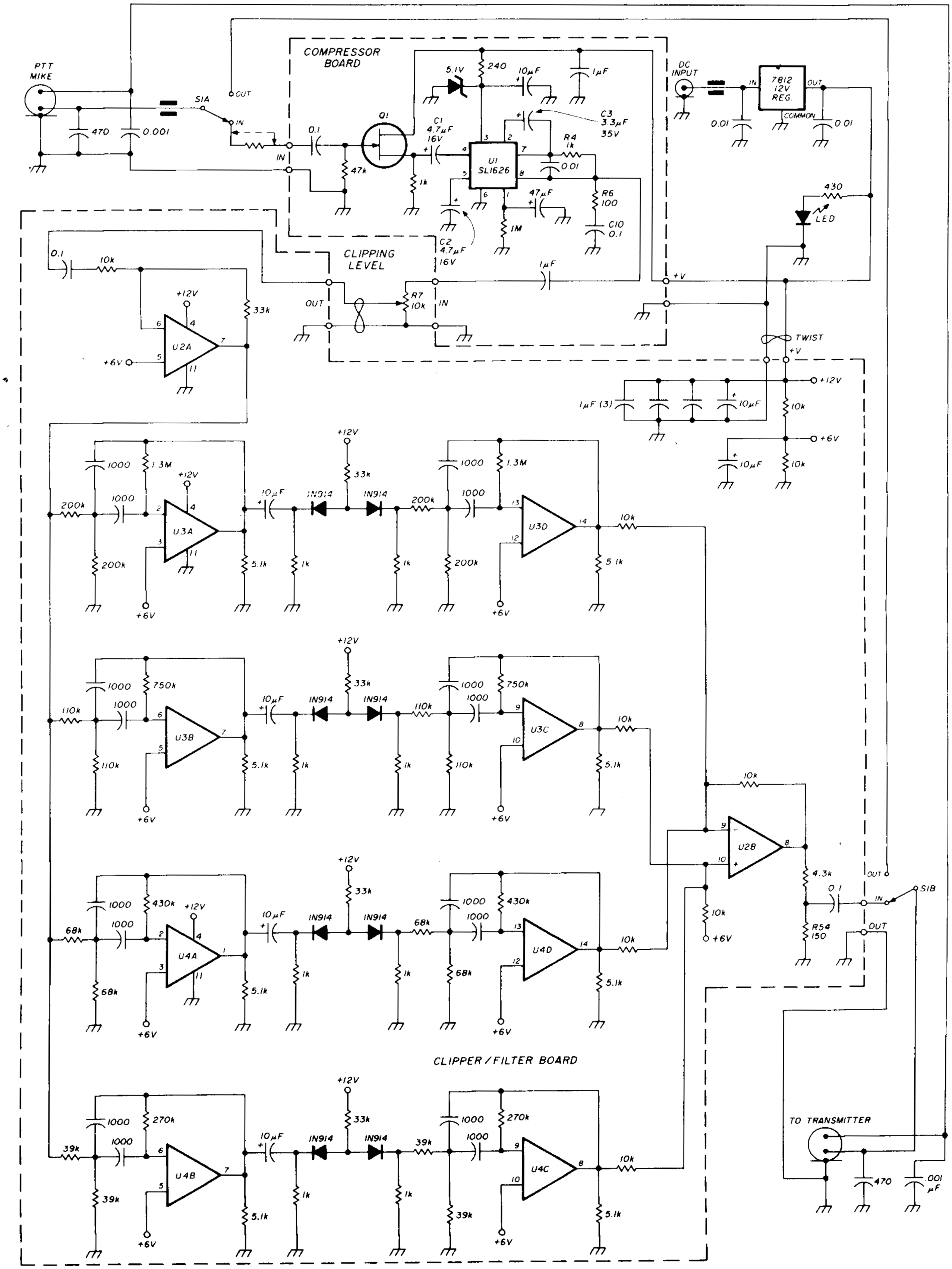


fig. 5. Complete schematic diagram of the split-band audio processor. Q1 is a 2N4392 or equivalent. U2, U3, and U4 are LM324s. C1 and C2 are dipped tantalum capacitors (RS 272-1409). C3 is also a dipped tantalum (RS 272-1408). All other polarized capacitors are tubular tantalums or electrolytics. The remaining capacitors are ceramics, with the exception of the 1000-pF capacitors, which are 5 per cent dipped micas. All resistors are 1/4-watt, 5 per cent, carbon composition.

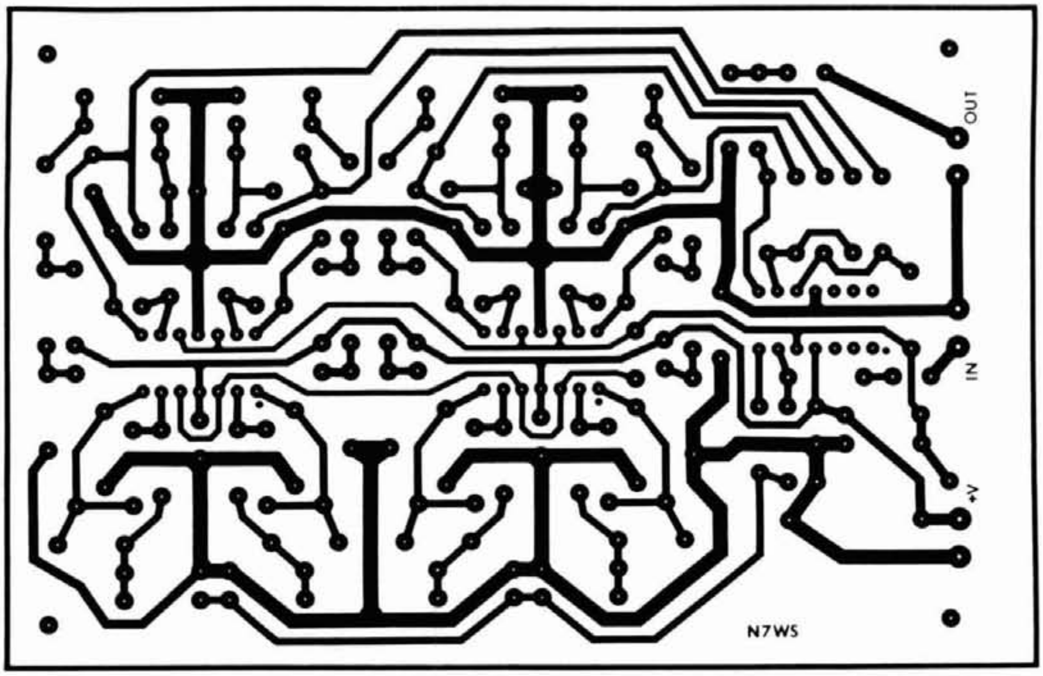


fig. 6. Full-size printed-circuit layout for the audio clipper board.

switched in and out without a gain change in the transmitter. Note that the BPF outputs are alternately connected to the plus and minus inputs of the combiner. The next section will demonstrate why this is done.

Assuming the equations shown in the appendix were used to determine the f_o and Q of the individual filters, adjacent filters will share a common −3 dB frequency. **Eq. 1** demonstrates how an input signal at this frequency is shifted +45 degrees in one channel and −45 degrees in the other:

$$\theta = 90 - \arctan\left(\frac{2Qf}{f_o} + \sqrt{4Q^2-1}\right) - \arctan\left(\frac{2Qf}{f_o} - \sqrt{4Q^2-1}\right) \quad (1)$$

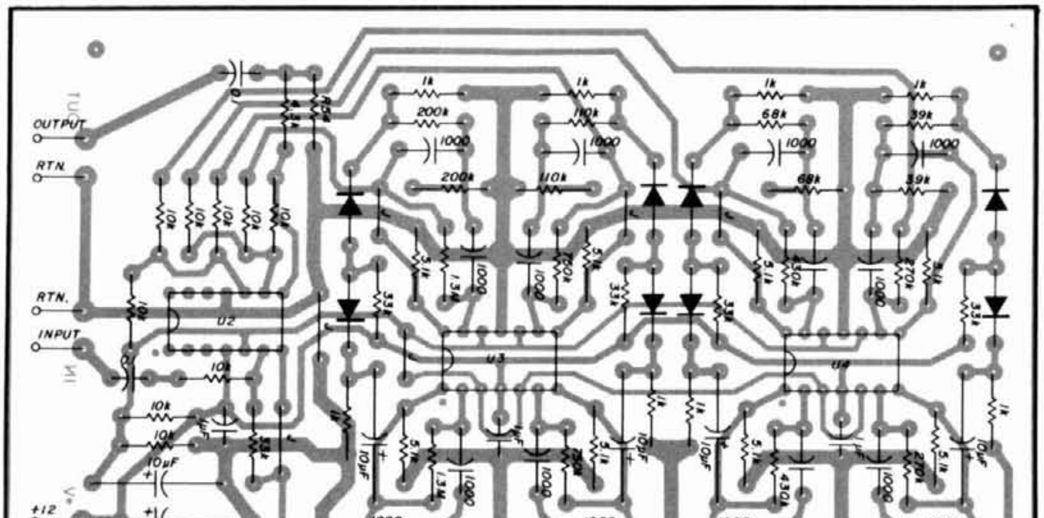


fig. 7. Component location on the audio clipper board.

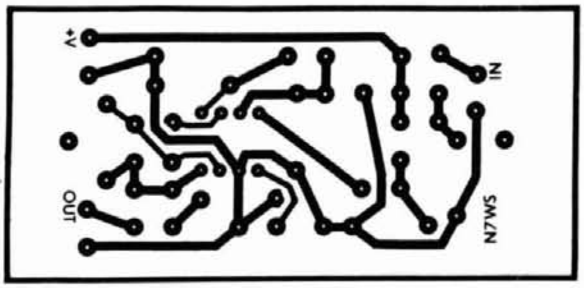


fig. 8. Full-size circuit board layout for the compressor board of the split-band processor.

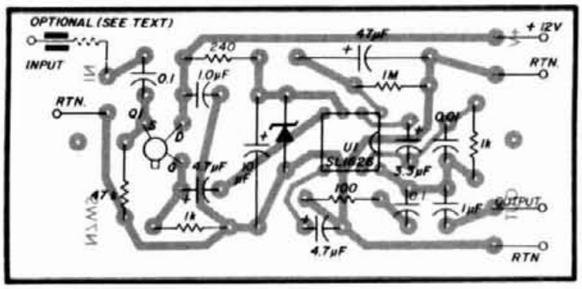


fig. 9. Parts placement diagram for the compressor circuit board.

where θ is the phase shift in degrees

f_o is the filter center frequency

f is the frequency of interest

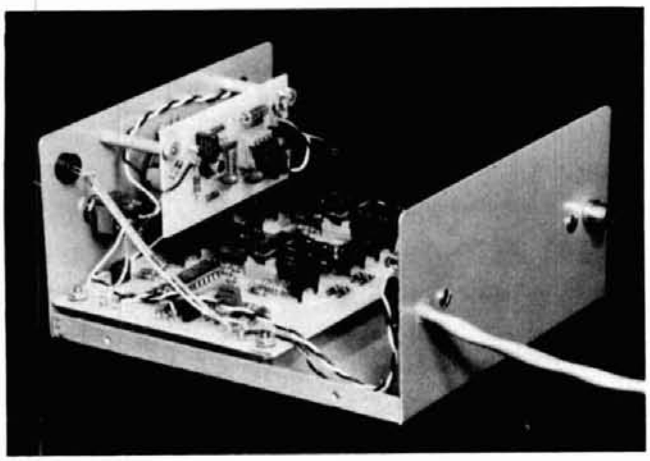
After cascading the two filters in each channel, this shift will be doubled to ± 90 degrees. Clearly, if these two signals are vectorially added, their sum will be zero because they are of equal amplitude but 180 degrees out of phase. A simple solution to this problem is to invert the phase of one signal. This is effectively what is done by the combiner.

Solving **eq. 1** for other frequencies will yield a phase error that increases with distance from the -3 dB point. This error is less important, however, because the amplitude difference also increases, so the larger signal dominates when the summation is made.

construction

For added versatility, the circuit is constructed on two etched circuit boards; the input compressor on one, the clipper-filter on another. This allows either one to be used alone in other applications. **Figs. 6** and **8** are full-size layouts of the foil sides of the two boards, while **figs. 7** and **9** show the component placement. These boards have been laid out with considerable attention to preventing ground loops. A hand-wired board should be built with the same attention.

fig. 10. View of the prototype split-band speech processor. The circuit boards are mounted using metal spacers and machine screws. Room is available for mounting an ac power supply; an external supply was used for this model.



The prototype shown in **fig. 10** was constructed in a Radio Shack enclosure (270-253). Sufficient space remains for the inclusion of an ac-operated power supply. **Fig. 11** is a schematic diagram of a suitable supply. Liberal use of ferrite beads and bypass capacitors on all leads entering the enclosure eliminates any chance of problems with rf interference.

performance

As **fig. 12** shows, the frequency response is very close to what was calculated, despite the use of 5 per cent components. By adjusting R7, the clipping level can be varied from 0 to 15 dB. Greater amounts of clipping can be had by increasing the gain of either U2 or the BPFs, or reducing the clipping stage bias to lower the clipping threshold.

Caution should be exercised before deciding on greater amounts of clipping, however. This could turn out to be too much of a good thing. Increased clipping does continue to reduce the peak-to-average ratio, but at the same time distortion increases rapidly. This is shown graphically in **fig. 13**. As pointed out by Moxon,[4] most of the improvement is obtained by the first 6 dB, with little to be gained by increased amounts. My on-the-air tests seem to indicate that 10 to 12 dB is about optimum with this system. All of this is rather subjective, but the whole topic of speech intelligibility and recognition is pretty subjective, so take it for whatever it's worth.

Total harmonic distortion was measured with an HP 331A distortion analyzer at various frequencies and clipping levels. The results of these measurements are shown graphically in **fig. 13**. As the figure indicates, distortion begins to rise rapidly as the clipping level approaches 15 dB.

These measurements were of necessity made with single frequency inputs which represent worst-case conditions. Because clipping is occurring on every

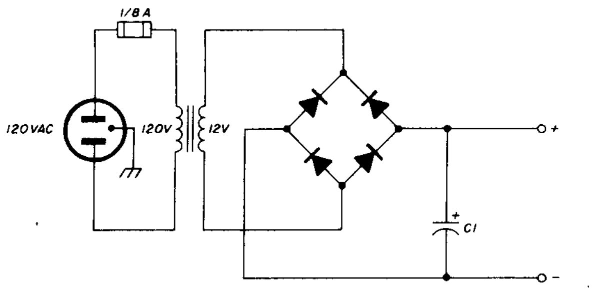


fig. 11. Schematic diagram of an ac power supply suitable for use with the processor. The transformer can be a Radio Shack 273-1385, the diode bridge a 276-1151, and C1 either 272-1019 or 272-1032.

half cycle, harmonic generation is maximum. With speech, clipping occurs much more randomly, with proportionally less total distortion.

On-the-air tests have been extremely gratifying. Reports have indicated substantial increases in apparent signal strength without noticeable distortion or loss of naturalness as long as the clipping level was held around the 10- to 12-dB point. Some loss of naturalness seems to occur above this point, but up to 15 dB, the sound is still not too objectionable. No tests have been run at levels in excess of 15 dB.

operation

Operation is very simple. The agc amplifier holds the clipping level constant, relaxing the operator requirements considerably. Some adjustment of the input sensitivity may be necessary if the microphone used has either a very high or very low output. While the dynamic range of the compressor will handle a higher input, the rise in background noise between speech pauses will be annoying to the listener. In this case, a series resistor may be added to the input which, in combination with R1, forms an attenuator. In the case of a very low-output mircophone, increasing the value of R4 will increase the sensitivity. Highest gain occurs with R4 omitted entirely.

On the output side, changing the value of R54 will control the maximum output level. This interacts with the audio gain control on the transmitter, so

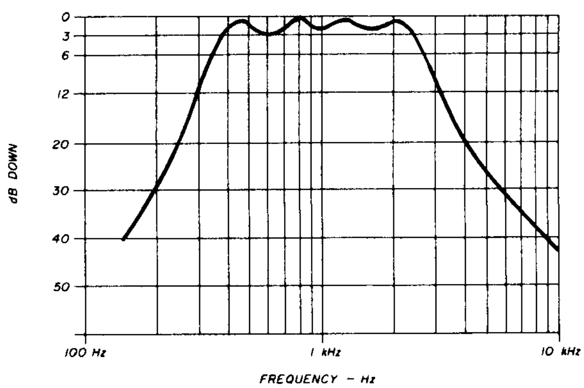


fig. 12. Measured audio response of the speech processor.

corrections can be made either place. I tried to pick a value that allowed the clipper to be switched in and out without having to readjust the microphone gain each time.

Finding the best setting for the microphone gain is best done with the aid of an oscilloscope on the transmitter output. With the clipping level set to maximum, adjust the transmitter gain so the peak output just approaches the level achieved with full carrier or excitation. If no oscilloscope is available, I find that just whistling into the microphone and setting the gain to the point that just activates the transmitter ALC works out very well. If you are not going to use the maximum amount of clipping available, then do the adjusting at the clipping level you intend to use. Even the best of clippers will not maintain a completely flat output vs input characteristic. Therefore, if you adjust your gain at 15 dB of clipping, then reduce it to 10 dB, your peak output will drop a little.

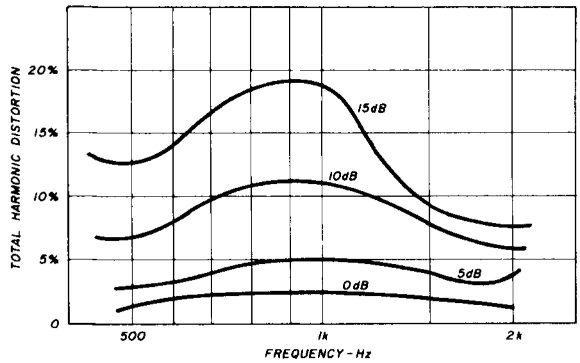


fig. 13. Total harmonic distortion vs clipping level. These curves were made with single tone inputs. Average distortion with speech input should be lower.

This effect can be explained as follows: As shown in **fig. 4**, sine waves subjected to 15 dB or so of clipping take on the appearance of pretty good square waves. As mathematical analysis can show, a square wave is composed of a fundamental frequency and all of its odd harmonics. We try to filter out these harmonics and retain only the fundamental. Unfortunately, the peak amplitude of this fundamental component is larger than the peak amplitude of the square wave by a factor of $\frac{4}{\pi}$, or 2.1 dB.[8] It is this factor that causes a continuing increase in output despite the use of a "perfect" limiter.

I want to express my thanks to Jim Metzger, W7TKR, for his technical advice, to Frank Baker for his circuit-board layout genius, and to Don Scheick and Norm Keopfer for their assistance in the preparation of the circuit boards. Additional thanks go to the many others who offered advice and encouragement, to Norma Putney for the typing of the manuscript, and to my wife, Terry, for the many hours spent away from family affairs during this project.

appendix 1

For new passband limits, the values for Q and f_o can be found as in the following example:

1. Define the low frequency − 6 dB point, f_L (350 Hz)
2. Define the upper frequency − 6 dB point, f_L (3000 Hz)
3. Find the multiplying coefficient, L

$$L^4 = \frac{f_H}{f_L} = \frac{3000}{350} = 8.571$$

$$L = \sqrt[4]{8.571} = 1.711$$

4. Find the individual filter cutoff frequencies

$$f_L = 350\,Hz$$
$$Lf_L = 599\,Hz$$
$$L^2f_L = 1025\,Hz$$
$$L^3f_L = 1753\,Hz$$
$$L^4f_L = 3000\,Hz$$

5. Find the individual center frequencies

$$f_{o1} = \sqrt{(350)(599)} = 458\,Hz$$
$$f_{o2} = \sqrt{(599)(1025)} = 784\,Hz$$
$$f_{o3} = \sqrt{(1025)(1753)} = 1340\,Hz$$
$$f_{o4} = \sqrt{(1753)(3000)} = 2293\,Hz$$

6. Determine required Q

$$Q = \frac{f_o}{BW}$$

$$Q_1 = \frac{458}{249} = 1.839$$

$$Q_2 = \frac{784}{426} = 1.840$$

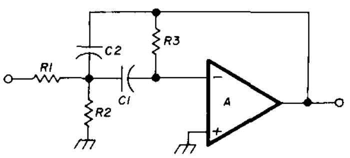
$$Q_3 = \frac{1340}{728} = 1.841$$

$$Q_4 = \frac{2293}{1247} = 1.839$$

Use $Q = 1.84$

appendix 2

The multiple feedback bandpass filter shown below may be designed by the following method (example in brackets):



Choose: $C = C_1 = C_2$

$$[C = 1000\,pF = 10^{-9}F]$$

Let: $H = \dfrac{|A_o|}{Q}$

$$[H = \frac{3.39}{1.84} = 1.84]$$

where A_o = desired gain

$Q = Q$ from **appendix 1**

Calculate: $K = 2\pi f_oC$

$$[2\pi \cdot 458 \cdot 10^{-9} = 2.878 \cdot 10^{-6}]$$

$$R_1 = \frac{1}{HK}$$

$$\left[\frac{1}{1.84 \cdot 2.878 \cdot 10^{-6}} = 188.8k\right]$$

$$R_2 = \frac{1}{K(2Q-H)}$$

$$[2.878 \cdot 10^{-6}(1.84) = 188.8k]$$

$$R_3 = \frac{2Q}{K}$$

$$\left[\frac{3.68}{2.878 \cdot 10^{-6}} = 1.28M\right]$$

This completes the calculations; the final step is to select the nearest 5 per cent standard resistor values. If, as in the above example, A_o equals Q^2, R_1 will equal R_2, which minimizes errors due to tolerance variations.

The following program, written for an HP 25 calculator, will speed the design of the BPF:

HP-25 Program Form

Title: Multiple Feedback Bandpass Filter

Switch to PRGM mode, press [f] [PRGM], then key in the program.

DISPLAY LINE	DISPLAY CODE	KEY ENTRY	X	Y	Z	T	COMMENTS
00							
01	31	↑					enter f_o
02	02	2					
03	61	X					
04	15 73	gπ					
05	61	X					
06	24 00	RCL 0					
07	61	X					defines K
08	23 06	STO 6					
09	24 05	RCL 5					
10	15 03	g ABS					
11	24 04	RCL 4					
12	71	÷					
13	23 07	STO 7					
14	24 06	RCL 6					
15	61	X					
16	15 22	g 1/x					defines R1
17	23 01	STO 1					
18	24 04	RCL 4					
19	02	2					
20	61	X					
21	24 07	RCL 7					
22	41	−					
23	24 06	RCL 6					
24	61	X					
25	15 22	g 1/x					defines R2
26	23 02	STO 2					
27	24 04	RCL 4					
28	02	2					
29	61	X					
30	24 06	RCL 6					
31	71	÷					defines R3
32	23 03	STO 3					
33	13 00	GTO 00					
34							
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REGISTERS	
R_0	C
R_1	R1
R_2	R2
R_3	R3
R_4	Q
R_5	A
R_6	K
R_7	H

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ham radio

antenna design for omnidirectional repeater coverage

The problem of good coverage with vhf antennas on towers with large cross-sectional areas is resolved in this article

This is the story of how one club obtained uniform coverage in all directions with a repeater antenna mounted on the side of a very wide tower. Perhaps the solution will help others with the typical problems of side-mounted antennas.

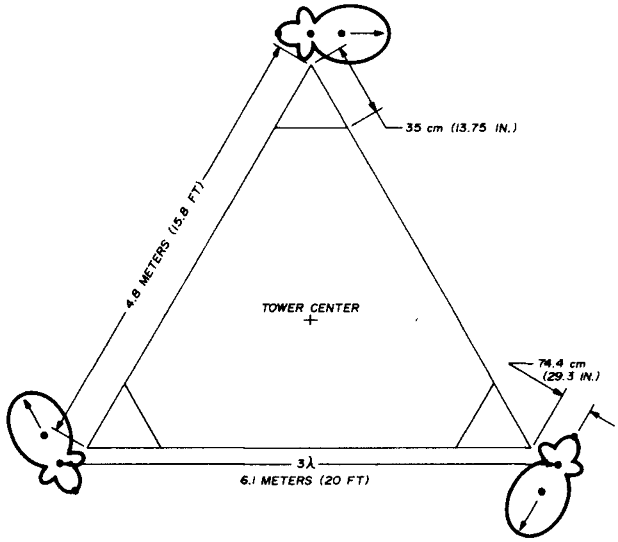
the problem

The difficulty that the Western Illinois Amateur Radio Club (WR9AEA) faced was not an unusual one on side-mounted repeater antennas. Coverage was not uniform in all directions; there were many peaks and many nulls. In some directions range was disappointingly short. Unless the repeater antenna is mounted on top of a structure this situation is typical, because a side-mounted antenna pattern usually has peaks and nulls resulting from the interference, reflections, and absorption of the structure. The local TV station, unfortunately, wouldn't let us put our array on the top of their tower above the TV antenna.

An interesting aspect of the WR9AEA problem was the large cross-sectional area of the TV broadcast tower we're using. The triangular shape is 4.8 meters (15 feet, 10 inches) on each side. Although this tower is very wide for a 244-meter (800-foot) structure, the problem and solution are relevant to both smaller and larger structures.

solution

The solution needed was some type of antenna array all the way around the supporting structure. Minimum coupling to the tower and uniform illumination of the horizon with good input vswr were required. A search of Amateur Radio reference materials yielded no answers. At this point the club



NOTES:
1. 63.5 mm (2-1/2 IN.) OD TUBING AT ALL TRIANGLE CORNERS.
2. ANTENNA HEIGHT: 223 METERS (731 FT) ABOVE GROUND.
3. OPERATING PARAMETERS:

	FREQUENCY (MHz)	WAVELENGTH, CM (IN.)
TRANSMIT:	147.03	204 (80.3)
RECEIVE:	147.63	203 (79.9)

4. RMS GAIN:
   2 LAYER ≈ 0.8 dB
   3 LAYER ≈ 3.8 dB

fig. 1. Tangential-fire antenna array using Yagis attached to a tower of large cross section. Note that the main lobe of each radiator is perpendicular to the tower and that free space exists in front of, and to the rear of, each pattern. The resultant radiation pattern of each antenna is summed so that the overall pattern is essentially omnidirectional.

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president, Tom, W9NJV, approached a local professional antenna engineer, Ron, W9NOO. Ron is very well respected for his many years of designing vhf and uhf broadcast antennas.

As usual, Ron knew what to do. He suggested a "tangential fire arrangement" for mounting antennas on the very large triangular tower. Of course, we didn't know what he was talking about; but as if often the case with someone who really knows his subject, Ron was able to make it simple for us.

description

By *"tangential"* Ron *meant that the radiators* would have their maximum radiation on a tangent, or at right angles, to the tower. This seems a little unusual at first, because we normally think in terms of *an antenna radiating straight out from a tower.* But here, if you're standing at the center of the tower, the maximum energy is pointed off to one side rather than straight out. To obtain constant signal amplitude *in all directions, one radiator is placed on each* leg of the tower. Notice from **fig. 1** that the main lobe of each radiator is prependicular to the tower and there is free space in front of, and to the rear of, each radiation pattern. The tower structure is off to one side of the radiator, so there's a minimum of coupling and distortion.

pattern sum

To obtain omnidirectional coverage it's necessary for the pattern from one radiator to add to the next, so that the resulting sum is as close as possible to a circle. **Fig. 2** illustrates this concept of the addition of the patterns. (In this figure the patterns are drawn to a very large scale, and the tower triangle to a very small scale, to represent the addition that takes place in the far field.)

The ideal individual radiation patterns would have a 6 dB beamwidth of 120 degrees. The half voltage (−6 dB) intensity of one radiator would then coincide with the half voltage radiation of the next. If the components from adjacent radiators are in phase, they will then sum to equal the full intensity. **Figs. 3** and **4** illustrate the development of this concept. Since the cosine function has a value of one-half ±60 degrees, the desired pattern shape is referred to as a cosine pattern.* The repeater antenna is vertically polarized, so our concern is the pattern

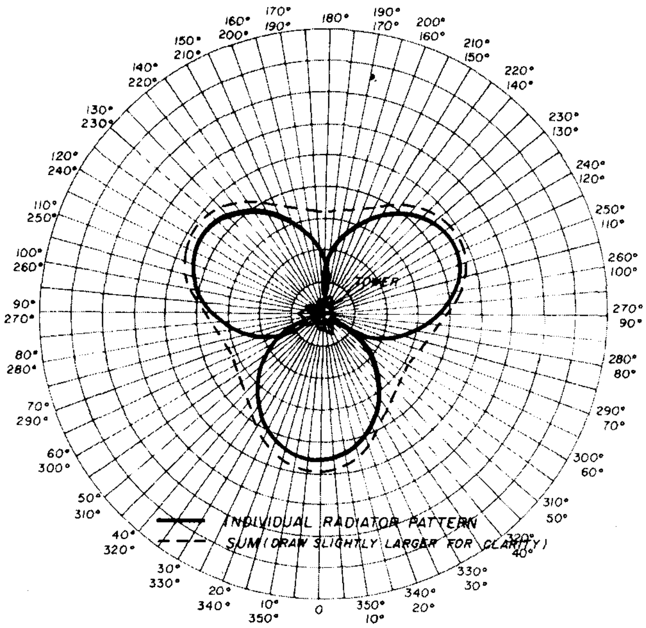


fig. 2. Development of the cosine radiation pattern resulting from three antennas fed with the proper phasing system. The sum of the patterns approaches a circle.

in the plane perpendicular to the radiating elements (**H** plane). Other patterns lend themselves to four or more radiators around a tower.[1, 2]

radiators

Ron told us that the desired cosine-shaped pattern is approximated by the typical short Yagi antenna. We decided to use on each leg of the tower a five-element Yagi manufactured locally. This beam

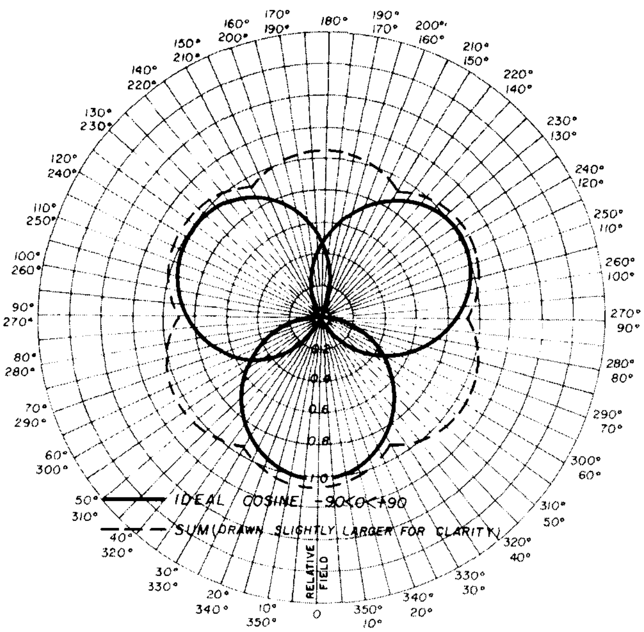


fig. 3. Ideal cosine pattern of three antennas fed in phase. The cosine of the angle, θ, lies between −90 and +90 degrees.

*Another variation of this concept is the cos² pattern, which was developed for vhf antennas on ballistic missiles. The same problem existed: the requirement for omnidirectional coverage with minimum attenuation from antennas mounted on the side of a huge mass of metal (the missile). Much time and effort went into the development of the cos² antenna, which is now standard for range safety and telemetry electronics on large rocket launch vehicles. Some of the early work on these antennas was done by the engineering department of the Convair division of General Dynamics for the Atlas missile in the late 1950s. Editor.

has standard dimensions with about 9 dB gain. It is very well constructed to take the rigors of being mounted 163 meters (535 feet) in the air. This was an important consideration, because nobody was interested in climbing up there — or paying a professional to go up there — in windy, cold weather to tighten a bunch of flapping aluminum.

spacing between radiators

For the amplitudes of the patterns of the radiators to add, it's necessary for the phasing and spacing to be correct. In our case, each Yagi was fed in phase

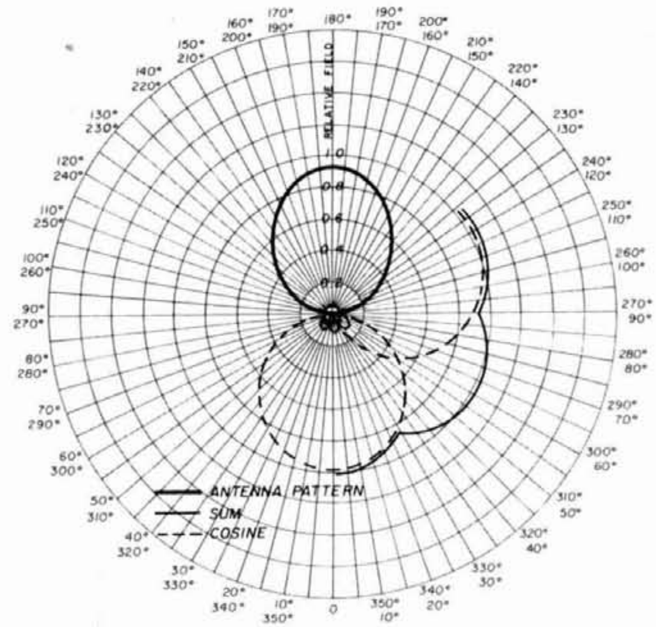


fig. 4. The development of the cosine pattern of n antennas fed in phase. This data is from a pattern recorder used during tests of the WR9AEA two-meter system.

through an equal length of feedline. The center of radiation (driven element) of each beam must be an integral number of free-space wavelengths apart. This requirement assures that the energy of each element will add correctly with energy from the next element. This is represented by the dimension $n \cdot \lambda$ (n times lambda), **fig. 5**. To suspend the Yagis at least one-half wavelength from the tower legs, the spacing worked out in our case to three wavelengths (see **fig. 1**). Ron pointed out that there are techniques for spacing the radiators at any multiple of one-third wavelength.[3]

gain of the array

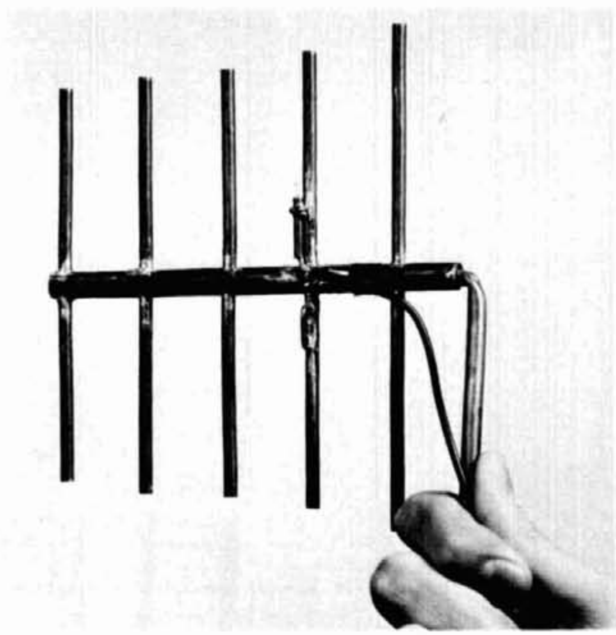
At this point some of us got enthusiastic about the gain of this concept. After all, with three 9 dB Yagis the gain should be high, right? Wrong. When the patterns add up to a circle, the average gain drops to that of a half-wavelength dipole. It was hard for some of us to get this through our thick heads, but the single stack or "bay" of three radiators around the tower yields to gain equal to that of a reference dipole.

Antenna-mounting hardware consists of stainless steel and heavy-gauge aluminum.

Ron pointed out that the addition of a second level, or bay, of three more Yagis, stacked one wavelength above, would double the gain and give 3 dB over a reference dipole. So we decided to build a two-bay system with three Yagis per bay.

scale-model tests

To make sure the thing would work, Ron and his collaborator, Joe Donovan, tested a scale model of



Scale model of one of the antennas used for tests.

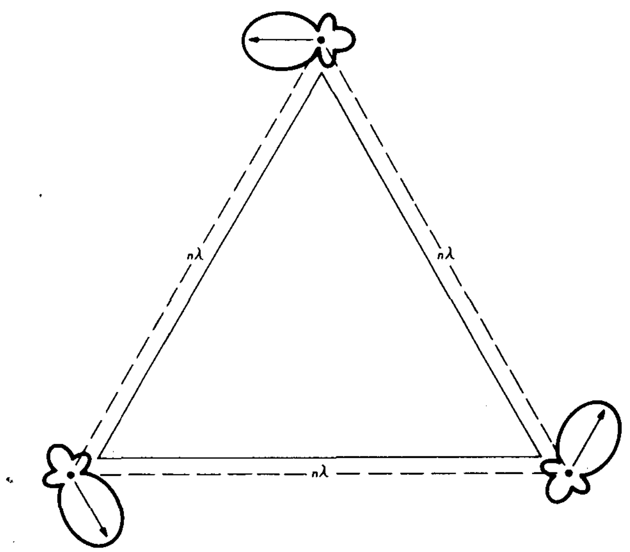


fig. 5. The center of radiation from each beam must be an integral number of free-space wavelengths apart so that the energy of each element (antenna) will add with energy from the next. This is represented by the dimension $n\lambda$, where λ is the spacing in wavelengths.

the tower cross section and Yagi elements. A convenient test frequency for their scale-model antenna was 955 MHz. At this frequency the models are small enough to be easily rotated by a powered turntable. A continuous plotter automatically recorded the pattern shape. **Fig. 6** shows the pattern with three Yagis pointed straight out, or a radial-fire arrange-

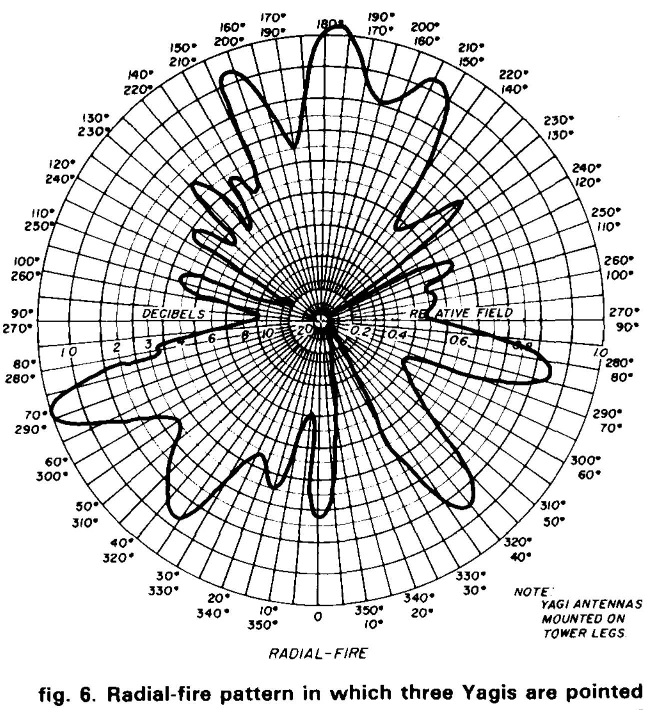


fig. 6. Radial-fire pattern in which three Yagis are pointed straight out from the tower. Note nulls and broad areas of low gain. The poor circularity is typical of many side-mounted vhf antennas.

ment. Note the nulls down to 20 dB below maximum and broad areas of poor gain. This type of poor circularity is typical of many side-mounted vhf antennas. **Fig. 7** shows the pattern of the tangential-fire configuration used for our new array. The circularity is ±3 dB or better. In other words, the gain in any direction is no more than 3 dB from the average.

power divider

A power divider to feed the six Yagis in phase from a single feedline was the next design task. A quarter-wavelength transmission-line transformer is

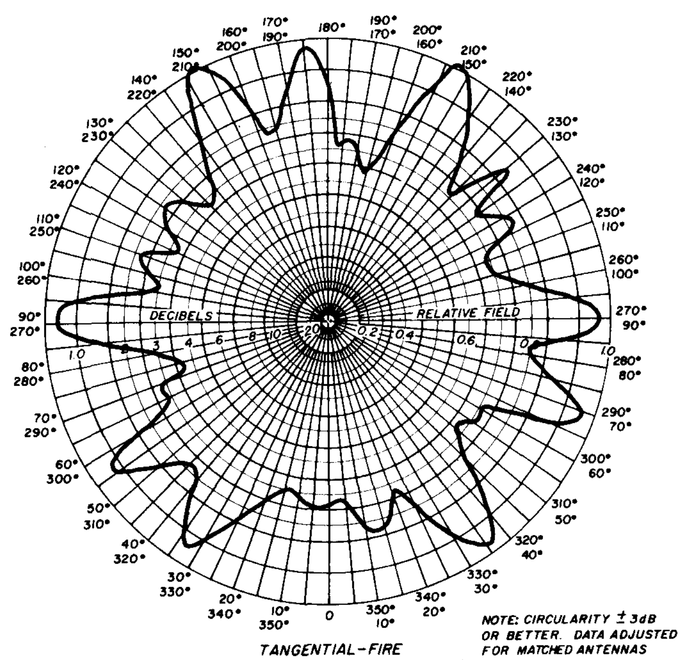


fig. 7. Radiation pattern of the tangential-fire arrangement used at WR9AEA. Circularity is ±3 dB or better, which means that the gain in any direction is no more than 3 dB from the average.

perhaps the most simple technique. If all six Yagis are matched to 50 ohms and fed through convenient, equal lengths of feedline, the feedlines can be paralleled at a single point. Six 50-ohm loads in parallel result in an impedance of 8.3 ohms. In other words, we need an impedance transformation of six to one.

The design curves in Chapter 22 of reference 4 shows about a ±5 per cent bandwidth at a vswr of 1.2 for a six-to-one transformation with a single 1/4-wavelength transformer. The usual equation, $Z = \sqrt{(Z1)\,(Z2)}$ or $Z = \sqrt{(50\ (8.3)}$, tells us that 1/4 wavelength of transmission line, with a characteristic impedance of 20.4 ohms, would match 8.3 to 50 ohms. However, the design curves also show that, by making the transformation in two steps, the bandwidth at a vswr of 1.2 can be increased to ±20 per

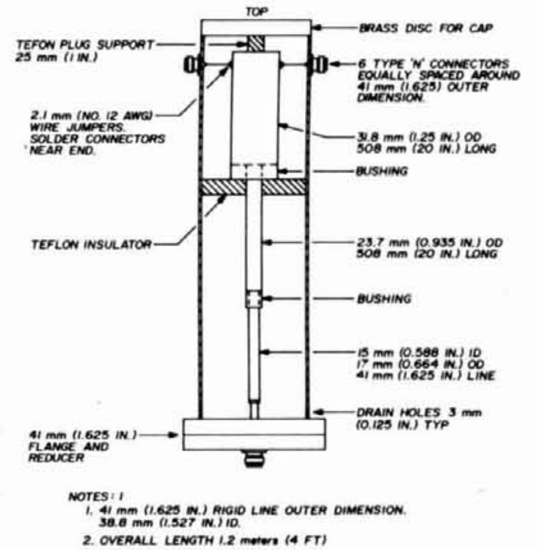


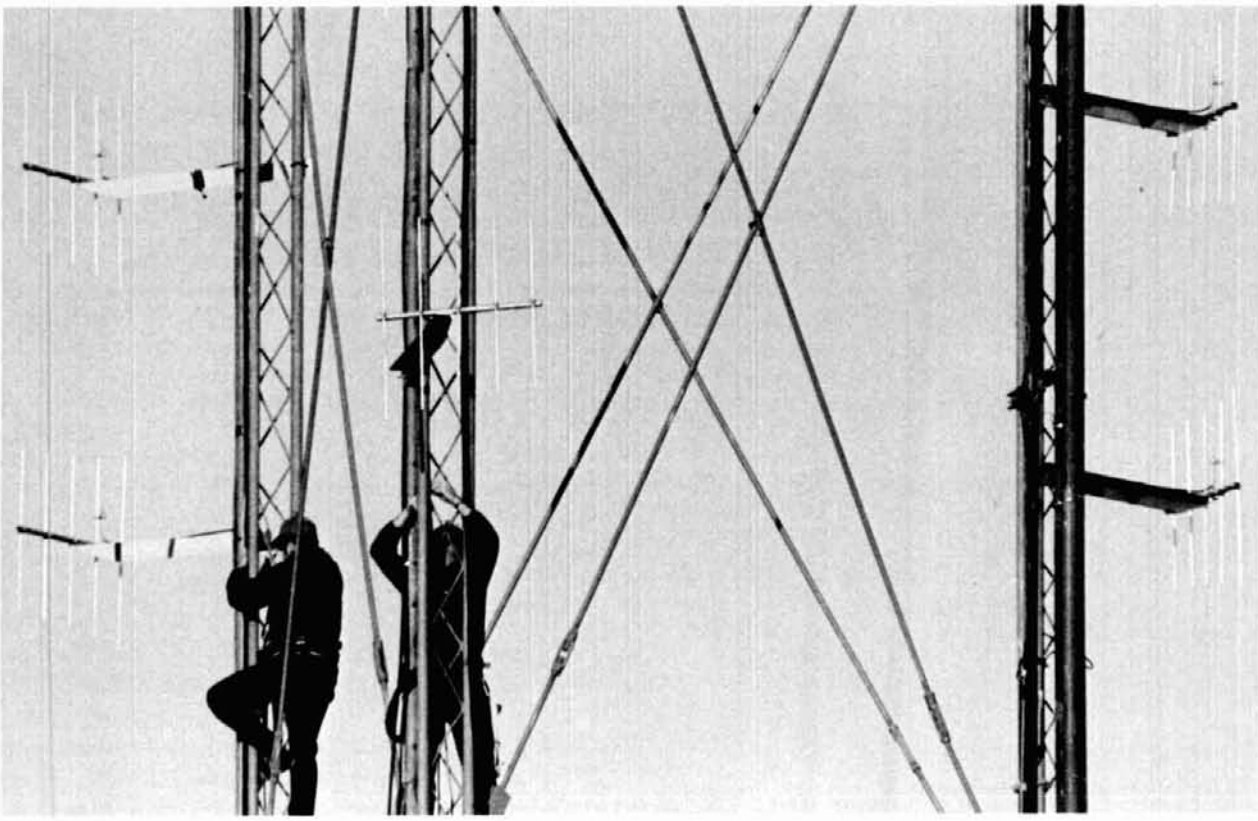
fig. 8. Construction details of the power divider used with the WR9AEA antenna.

cent. This configuration is less sensitive to inaccuracies and changes in the load impedance. Two 1/4-wavelength sections in series match the 8.3-ohm load to an intermediate value to 20.4 ohms, which is in turn matched to 50 ohms. One 1/4-wavelength section has an impedance of $\sqrt{(50)\ (20.4)} = 32\ ohms$ and the other $\sqrt{(20.4)\ (8.3)} = 13\ ohms$.

divider construction

Fig. 8 shows the construction of the power divider. The 1/4-wavelength sections are coaxial. Therefore the usual formula $Z = 138\ log\ (db)$ was used to calculate the ratio of the diameter of the outer to inner conductors. It was convenient to construct the outer shell from a piece of 41-mm (1.625-inch) rigid coax line. The 50-ohm type N input was constructed from a 41-mm (1.625-inch) flange, a 41-mm (1.625-inch) reducer and a short section of 41-mm (1.625-inch) inner conductor. The six outputs are type N connectors spaced equally around the circumference at the opposite end. The center conductors of the six type N outputs are connected in parallel with short lengths of 2.1-mm (no. 12 AWG) solid copper wire to the end of the last 1/4-wavelength inner conductor. Some routine lathe work was necessary to construct the inner conductors, brushings, Teflon supports, and end cap.

Rex, K9ZJV, put his workshop facilities to the task of constructing the divider. Initial testing showed a very flat vswr of about 1.22 over the whole 2-meter band. To bring the device up to professional stan-



The WR9AEA array on an fm broadcast antenna tower. Array is at the 20-meter (65-foot) level for testing. The heroes doing their thing for the cause are N9SN, left, and W9NWN.

dards, a stub was added to the input transmission line to reduce the vswr to less than 1.1 from about 142 to 151 MHz. See **fig. 9**.

full-scale tests

The Yagis, mounting hardware, feedlines, and power divider were then mounted on the TV tower at the 20-meter (65-foot) level. Jim, N9SN, and Dave, W9NWN, performed these tasks of installing and adjusting the antennas. This work provided a very important check of all parts of the system before the critical full-height installation. A check of the pattern was made by comparing the signal received from the array with that from a reference Yagi, hand-held out from the tower in the direction of the field-strength meter. Although this method of checking a pattern isn't accurate, seventeen measurements in all directions showed no major peaks or holes. Once all the minor mechanical bugs were corrected, a professional climber was hired to install the array several wavelengths below an fm broadcast transmitting antenna, approximately 163 meters (535 feet) high.

predicted coverage

Ed, W4HTP, calculated the predicted coverage using broadcast techniques. FCC 50/50 curves calculate coverage exceeding 50 per cent of the time in 50 per cent of the potential receiving locations. The calculations consist of two steps: prediction of field strength from the repeater transmitter and determination of field strength required by various configurations of fixed and mobile stations. The results are shown in **table 1**. (Reference 5 and 6.)

table 1. Predicted coverage of the WR9AEA two-meter repeater antenna array for various receiving antennas. Predictions are based on 50 per cent of the potential receiving locations for 50 per cent of the time. Distances are for receiving the repeater.

Example	receiving equipment	required field strength at 9 meters (30 ft)	Distance km	(miles)
1	1/4-wavelength rooftop mobile	+21 dBμV/m	77	(48)
2	5/8-wavelength rooftop mobile	+18 dBμV/m	85	(53)
3	Ringo at 9 meters (30 ft)	+1 dBμV/m	144	(90)
4	11-element beam at 12 meters (40 ft)	−12 dBμV/m	216	(135)
5	2 stacked 11-element beams at 24 meters (80 ft)	−21 dBμV/m	280	(175)

results

Results have been excellent. Coverage in all directions seems to bear out the predictions. Mobile coverage is 72-88 km (45-55 miles); fixed stations at 160 km (100 miles) check in regularly. There appear to be no holes in the pattern. All bad spots seem to be explained by local terrain. We hope our experience and the references will help other groups to obtain omnidirectional repeater service.

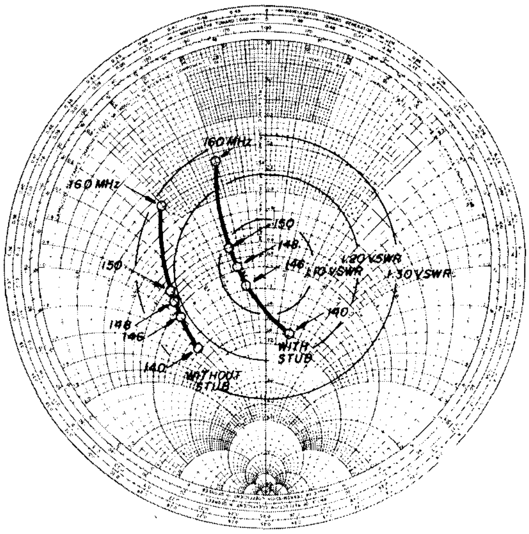


fig. 9. Response of the power divider with and without a stub on the transmission line.

acknowledgments

The Western Illinois Amateur Radio Club wishes to thank Ron Fisk, W9NOO, for his professional guidance. We also thank the club president, Tom, W9NJV, for his help during this project. Tom deserves public recognition for his constant, active leadership in getting everybody to work together. Photo credits to to Roger Humke, WA9KRG.

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ham radio

exposure to radio-frequency generating equipment: is it safe?

Review of the literature has produced some interesting observations on the hazards of EMR — here's a report of the latest information on this controversial subject

Those of us involved with electromagnetic radiation (EMR) for business or pleasure have, in the last few years, become disturbed by certain questions, which grow louder and more insistent as time passes. These questions address the safety, or lack of safety, associated with exposure to EMR. The principal target of these inquiries has been in the area of microwave radiation, but there has been increased interest in the matter of safety within the more usual communication frequencies: uhf, vhf, and high frequency. These questions have been amplified, perhaps more than they deserve, by the press; and it is becoming harder and harder to ignore them.[1] You find yourself wondering whether the push-to-talk button and antenna on a handheld vhf transceiver might better be replaced by a trigger and gun barrel respectively.

While under most conditions we can't "feel" EMR, we're becoming more aware of its presence. This is an unwelcome occurrence, since most of us have always thought of shortwave radiation as being nothing if not safe. We all live most of our lives within a constant "fog" of EMR of many frequencies. In fact, we're constantly affected by the information transmitted by radio waves. The suspected damage that may be caused by rf energy varies from the "blahs" to cataracts, heart disease, cancer, impotence, and birth defects. What's worse, while most of this is unproven, it hits the lay press with the impact of fact.

By Steve Kraman, MD, WA2UMY, 2901-B Candlelight Way, Lexington, Kentucky 40502

Even the more responsible reviews of the subject offer no assurance that we may continue our use of radio with impunity.[2,3]

Distrust of the establishment also has had a hand in promoting doubt among the rf-irradiated public. We mistrust our government when we hear that the Russians have a standard for safe rf fields that is one thousand times smaller than ours, and that they believe that many adverse mental health and physical effects can be produced in man by low-dose rf exposure. We read that the Russians have been beaming microwave radiation (for questionable reasons) at the U.S. embassy in Moscow for years; then we hear that more cases of cancer may occur among the embassy workers than can be explained by chance. When the United States government chalks this up to chance, we wonder.[4]

The purpose of this article is not to reassure you that there's nothing to worry about; not quite enough information is available for that yet. Nor is the article intended to alarm, since I don't believe that the known facts justify that either. The following text reviews what is known to date about EMR and its effects on biological systems. It separates fact from fiction and suggests a reasonable response by those who must swim in the sea of EMR that surrounds us.

history

Interest in the biological effects of EMR has waxed and waned considerably since the first true electromagnetic field was generated by Hertz in 1888. Some research was done in 1891 by D'Arsonval and Tesla, but this was essentially the only work done in this area until the 1930s. Before this, in 1891, when electric lights were installed in the White House for the first time, they were not placed in the rooms that the President used frequently, because they were considered potentially dangerous. During the 1930s interest in rf radiation began to grow, spurred by the development of high-power transmitting techniques.

But World War II caused this research to grind to a halt in the favor of work of a more certain and conventionally destructive nature. At the end of World War II we had a new toy to learn about and with which to experiment — nuclear energy. However, this knowledge didn't stop progress in the field of radar and communications. The existence of high-power rf-generating equipment began raising safety questions, principally within the military community.

Much has been published on the hazards of exposure to electromagnetic radiation, from low-power, low-frequency equipment to high-power microwave devices. The author of this article is a medical doctor and a Radio Amateur interested in this controversial subject. He has researched the available literature (domestic and foreign) on the subject. This article sums up the results of that research. The conclusions imply that Amateur Radio transmitting equipment probably does not impose health hazards on humans provided certain safety considerations are observed. **Editor.**

The Tri-Service study. In 1956 the Tri-Service program was established, coordinated by the Air Force. Its purpose was to conduct research to determine the biological effects of nonionizing radiation. This research effort lasted four years and four annual conferences were held. The outcome of the Tri-Service program was to suggest that there was no evidence implicating levels of electromagnetic radiation below 100 mW/cm^2 in damage to living tissues.

The implication that EMR could cause damage only in its capacity to heat was clear. It's of importance, however, that very little, if any, of this research was done at levels below 100 mW/cm^2 and, indeed, most of it was between the power levels of 300-400 mW/cm^2. The Tri-Service study, then, addressed only the problem of thermal effects and assumed this to be the only danger. The government accepted this opinion, and, partly as a result, little further research was done in this country from 1960 to 1970.

Federal EMR legislation. During the present decade, interest in the biological effects of EMR has escalated steadily, primarily because of technological advancements that have resulted in increased exposure to EMR by the general population. The skyrocketing popularity of CB radio and microwave cooking account for much of this increased exposure. The use of high-power communications and radar equipment has also heightened concern for personnel in the military. Additionally, several pieces of federal legislation have stimulated research in the field by calling for protection of the public from all sources of radiation, including ionizing, nonionizing, sonic, and ultrasonic devices. These federal acts are the Radiation Control for Health and Safety Act of 1968, the National Environmental Policy Act of 1969, and the Occupational Safety and Health Act of 1970. They require that users of EMR-generating equipment demonstrate the safety and effects on the environment of their equipment. Yet another factor that helped spur American researchers was the presence of a large body of Soviet-bloc literature that points to conclusions much different from those of the Tri-Service program and imply that very low levels of EMR (by our standards) could be dangerous. These studies, while for the most part poorly controlled, carried out, and reported, could not be totally ignored, since many of our own studies suffer from the same shortcomings.[5] The Russian studies are covered in more detail later.

physical characteristics of EMR

EMR must be distinguished from ionizing radiation (x-ray, nuclear), since its effect on molecular structure is much different. Nuclear radiation causes no significant heating of the irradiated object. Instead, its photon energy is sufficient to disrupt the atomic bonds, thereby causing ionization and damage to the molecular structure. If this molecule is part of a living cell, it may become damaged, die, or its genetic material may be changed. EMR doesn't cause these effects, because the photon energy of even microwaves is so small that it causes no ionization. The energy absorbed, however, can increase the speed of molecular vibration, thereby causing an increase in temperature. The more energy absorbed, the more heat produced. *To date, all known damage by EMR seems to be* the result of this heating. This effect is clearly seen through the window of a microwave oven.

Importance of EMR absorption. Depending on the EMR frequency, the size and character of the target, and the presence of other objects, a certain amount of energy will be absorbed and the rest will be reflected or refracted. Only the energy absorbed by the object affects it, and this has been one of the problems in microwave research. Most studies conducted to assess the effects of microwave radiation on biological subjects measure the field strength of the electromagnetic field, even though the actual amount of energy absorbed is unknown. Not only is the absorbed dose important, but so is the *size* of the subject and the EMR *frequency,* since the depth of penetration decreases as the frequency is increased.

An animal may handle a heat load more easily if the heat load is applied to its skin, where air cooling occurs, than if the heat load is developed internally within vital organs that are cooled only by blood circulation. In this way EMR research is years behind nuclear research, which has long recognized and used the *rad,* a unit of absorbed energy, when referring to exposure to radioactive materials. The *roentgen,* a unit of emitted energy, is fine for describing the generating equipment but says little about its effect on the person receiving the radiation. Recent studies in the field of EMR have used methods to calculate or measure the actual absorbed dose of radiation.

table 1. Properties of electromagnetic waves in biological media*

		muscle, skin, and tissues with high water content							
	wavelength	dielectric	conductivity	wavelength	depth of	reflection coefficient			
frequency	in air	constant	σ_H	λ_H	penetration	air-muscle interface		muscle-fat interface	
(MHz)	(cm)	ϵ_H	(mho/m)	(cm)	(cm)	τ	ϕ	τ	ϕ
1	30000	2000	0.400	436	91.3	0.982	+179		
10	3000	160	0.625	118	21.6	0.956	+178		
27.12	1106	113	0.612	68.1	14.3	0.925	+177	0.651	−11.13
40.68	738	97.3	0.693	51.3	11.2	0.913	+176	0.652	−10.21
100	300	71.7	0.889	27	6.66	0.881	+175	0.650	−7.96
200	150	56.5	1.28	16.6	4.79	0.844	+175	0.612	−8.06
300	100	54	1.37	11.9	3.89	0.825	+175	0.592	−8.14
433	69.3	53	1.43	8.76	3.57	0.803	+175	0.562	−7.06
750	40	52	1.54	5.34	3.18	0.779	+176	0.532	−5.69
915	32.8	51	1.60	4.46	3.04	0.772	+177	0.519	−4.32
1500	20	49	1.77	2.81	2.42	0.761	+177	0.506	−3.66
2450	12.2	47	2.21	1.76	1.70	0.754	+177	0.500	−3.88
3000	10	46	2.26	1.45	1.61	0.751	+178	0.495	−3.20
5000	6	44	3.92	0.89	0.788	0.749	+177	0.502	−4.95
5800	5.17	43.3	4.73	0.775	0.720	0.746	+177	0.502	−4.29
8000	3.75	40	7.65	0.578	0.413	0.744	+176	0.513	−6.65
10000	3	39.9	10.3	0.464	0.343	0.743	+176	0.518	−5.95

Near- and far-field considerations. Another factor affecting the field density is whether the object is in the near or far field. The far field is that distance (generally more than one wavelength) beyond which the electric and magnetic fields are coherent and in phase. The field impedance is constant in the far field, so that measurement of either the electric or magnetic component will be proportional to and will determine the power density. In the near field, however, the electric and magnetic fields are out of phase, and it becomes more difficult to measure power density. In this situation it's more convenient

*Tables reprinted from "Non-Ionizing Electromagnetic Wave Effects In Biological Materials And Systems" by C. C. Johnson and A. W. Guy, *Proceedings of the IEEE,* Vol. 60, No. 6, June, 1972.

to measure volts per meter (V/m).[6] This is of little consequence with microwaves, because the far-field situation exists at distances of only a few meters or so from the antenna. However, at most communications frequencies, the object in question is often in the near field. At a wavelength of 80 meters for instance, the entire dwelling and many of the neighbors may be in the near field, and this makes research into near-field phenomena quite important.

The so-called nonthermal effects of EMR are the more controversial aspects of the subject. The present research push is to discover if these effects exist and, if so, whether they offer significant health hazards. This research is necessary because most, if not all, past studies reporting to show nonthermal effects ignored regional temperature changes caused by concentration within objects. The studies were poorly or not controlled, entirely anecdotal in nature (and therefore impossible to evaluate), or were so incompletely described that they can't be duplicated.[7]

The scientific method demands that, as much as possible, all factors other than that being evaluated be accounted for and set aside to attribute a possible effect to a certain cause. This is impossible to do with certainty, so the importance of statistical analysis (to determine the possibility that chance alone caused a certain effect), and repetition by other experimenters, is of extreme importance if you don't want to be led astray. This is the essence of the scientific method, and while well understood for decades, it's often overlooked by scientists who may be in a hurry and feel secure that they can be objective.

The Soviets recognized the effects of EMR on human nervous tissue as far back as 1937, when Turlygin found that excitability of the central nervous system was increased when a spark oscillator was switched on near the subject's head (not a totally unexpected effect). Since then, with the exception of the war years, Soviet-bloc literature in this area has poured out in increasing quantities.

While many effects have been reported, those most frequently noted involve the central nervous system. These reports include frequencies from 30 to 30,000 MHz and power ranges of microvolts to tens of milliwatts/cm^2. Unfortunately, and as previously mentioned, most of these reports lack data without which intelligent evaluation is impossible; *i.e.*, frequency, power, waveform, orientation of the body with respect to the beam, and type of experimental animal used. Many of the reports involving people exposed to EMR quote a wide range of subjective complaints such as headache, weakness, depression, trembling, chest pains, inhibition of sex drive, inability to make decisions, general tension, and sense of anxiety. Other more objective findings reported are asthma, fast or slow pulse rate, high or low blood pressure, and EKG changes.

The belief that these ailments are being caused by EMR exposure is so strong in the Soviet Union that exposed workers can get the day off with pay if they complain of them. There is, however, a question regarding the willingness of Soviet plant managers to admit they've been exposing workers to higher-than-permitted levels of electromagnetic radiation for fear of losing their jobs. In the realm of parapsychology,

table 2. Properties of electromagnetic waves in biological media*

		fat, bone, and tissues with low water content							
	wavelength	dielectric	conductivity	wavelength	depth of	reflection coefficient			
frequency	in air	constant	σ_L	λ_L	penetration	air-muscle interface		muscle-fat interface	
(MHz)	(cm)	ϵ_L	(mho/m)	(cm)	(cm)	τ	ϕ	τ	ϕ
1	30000								
10	3000								
27.12	1106	20	10.9-43.2	241	159	0.660	+174	0.651	+169
40.68	738	14.6	12.6-52.8	187	118	0.617	+173	0.652	+170
100	300	7.45	19.1-75.9	106	60.4	0.511	+168	0.650	+172
200	150	5.95	25.8-94.2	59.7	39.2	0.458	+168	0.612	+172
300	100	5.7	31.6-107	41.0	32.1	0.438	+169	0.592	+172
433	69.3	5.6	37.9-118	28.8	26.2	0.427	+170	0.562	+173
750	40	5.6	49.8-138	16.8	23	0.415	+173	0.532	+174
915	32.8	5.6	55.6-147	13.7	17.7	0.417	+173	0.519	+176
1500	20	5.6	70.8-171	8.41	13.9	0.412	+174	0.506	+176
2450	12.2	5.5	96.4-213	5.21	11.2	0.406	+176	0.500	+176
3000	10	5.5	110-234	4.25	9.74	0.406	+176	0.495	+177
5000	6	5.5	162-309	2.63	6.67	0.393	+176	0.502	+175
5800	5.17	5.05	186-338	2.29	5.24	0.388	+176	0.502	+176
8000	3.75	4.7	255-431	1.73	4.61	0.371	+176	0.513	+173
10000	3	4.5	324-549	1.41	3.39	0.363	+175	0.518	+174

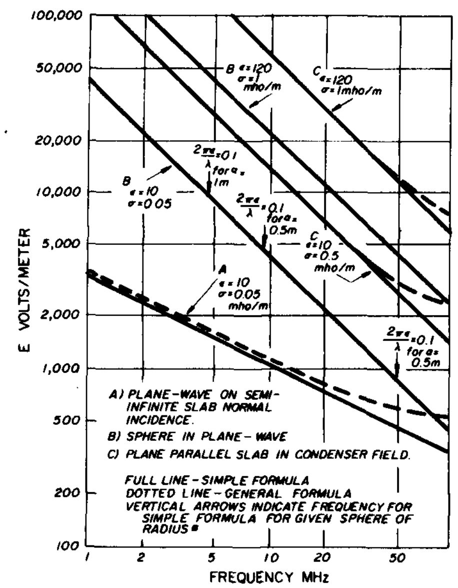

fig. 1. Variation with frequency of electric-field strength to produce rate of rise of temperature at 1°C/hr, in: (A) The skin depth of semi-infinite layer of lossy dielectric in plane-wave field, (B) Sphere of lossy dielectric in plane-wave field, (C) Slab of lossy dielectric in capacitor field.

many Russian experimenters believe not only in esp, but also that it is effected through microwave transmission and reception from brain to brain!

The Eastern European literature describes microwave effects on many constituents of the blood, on gland functions, eyes (cataracts), and reproduction (sterility, altered development of the fetus, altered sex ratio of births [with girls predominating]).[7] These results over many years have led the Soviet-bloc countries to adopt a maximum permissible dose (MPD) of EMR of 0.01 mW/cm^2 (one one-thousandth of the U.S. MPD). While this is looked upon with considerable disdain by Western scientists, the remarkable consistency of the many reports can't be discarded out of hand. This has led many U.S. investigators to run better controlled studies to try to prove or disprove these reports.

Recent studies, mostly from the West but also notably from Poland, have shown higher degrees of control and sophistication and are probably more reliable than older reports. More accurate generating and measuring devices are being developed and used, and we are becoming more aware of the physics of EMR and how the conditions of the experiments affect the rf fields produced and power absorbed. Grants from the National Institute of Health and Public Health Service, among other agencies, have spurred research in this field. Most of the work has been with microwaves, and I will outline some of it here.

cataracts

It's generally accepted that microwave radiation can cause cataract formation (opacities in the lens of the eye). Reports of this affliction occurring in relation to radar work or exposure to microwave ovens have been fairly well documented. Also, many animal studies have been made to determine the basis of this effect. While all is not known with respect to this matter, it seems quite certain that the formation of cataracts results from the heating of the lens in a strong microwave field (that is, a field not associated with correct use of properly operating equipment).

A fairly well documented case was that of a woman whose microwave oven leaked considerable radiation while the door was being opened. The level of radiation was 40-60 mW/cm^2 while opening and 1-2 mW/cm^2 while closed. She used it for years that way and developed cataracts in both eyes described as "typical microwave cataracts."

It must be mentioned that not all ophthalmologists agree that such cataracts are typical, and many people develop cataracts with age. Other reports of documented cataract formation are those of radar workers when abnormally over exposed (looking into the waveguide).

The optic lens is susceptible to selective heating because circulation is practically nonexistent. Therefore, it has limited capacity to dispose of heat loads. There is no evidence that nonheating levels of microwave radiation can cause damage, and the levels generated by a properly operating microwave oven are far below this point (1 mW/cm^2 when new; 5 mW/cm^2 when used — measured at 5 cm (2 inches) from the door).[8]

effect on the nervous system

Studies of the effects of microwave radiation on the brain have been spurred by the large number of reports in the Soviet literature of emotional and performance changes and the frequent reports of people who can "hear" microwaves — specifically radar. These studies conclude that many persons indeed can hear pulsed microwave energy. But they hear it at the frequency of modulation and cannot detect CW radiation. The actual cause of this is still not clear, but it may be due to selective heating and cooling of certain nerve cells in the ear or brain causing vibration that is detected as sound.[9] Actual brain damage has been demonstrated only with levels of radiation far in excess of the present U.S. safety limit.

Studies on the activity of mice subjected to low-power microwave radiation have been contradictory (some show no effect, some show decreased activity). These have been carried out at different frequencies and power densities so that further research is certainly needed to clear the air.

blood-forming system

Several studies have been done to assess the effects of microwaves on the blood and especially on the white blood cells, which are responsible for protection against infection. Most of these studies show little or no effect and the importance of this is questionable so far. More research is also needed in this area.[10]

reproduction

Reproduction research is of obvious importance because of the known sensitivity of the fetus to subtle changes in its environment. Caution in this matter is extreme. Witness the fact that there is not even one drug that is *known* to be safe for use during pregnancy. A potential risk is always assumed. Radio-frequency energy in the microwave range, however, is probably not a danger to pregnant women, because the energy is absorbed by more superficial tissues (penetration of EMR decreases as frequency increases). The effect on male fertility is real, however, since the testes are heat sensitive and must exist in an environment cooler even than body temperature to produce sperm. This is the reason for their location. Heating of the testes by microwaves or anything else will cause sterility, but this is temporary unless extreme heating occurs or exposure lasts over many months to years. So, while all the information is not in, there seems to be little or no danger to reproduction from current microwave exposure levels.

uhf, vhf, hf studies

Much less work has been done at these frequencies than at the microwave level. However, it's important to explore this area, since we're exposed more to EMR in this part of the spectrum and such energy can penetrate deeper into the body than microwaves do. As previously mentioned, the near field is more significant at lower frequencies because it occupies more space.

behaviorial effects

Little has been done in this area. One study of interest exposed rats to low intensity (0.5 mW/cm^2) 300-920 MHz radiation for 40 days. While the rats were probably in the near field (and this was really not accounted for), certain effects were noted:

1. Lower levels of activity

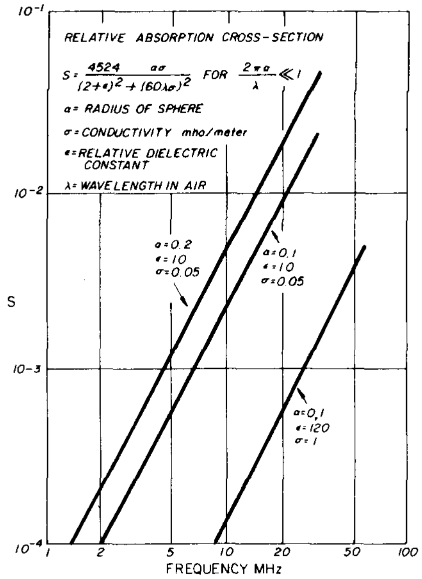

fig. 2. Variation with frequency of the radiation absorption cross section of a sphere of lossy dielectric.*

2. Greater emotionality

3. Longer period of time needed to recover from an electrically induced convulsion

4. Longer time to learn to swim in a water maze

5. Difference in weight of the adrenal gland[11]

This study raises more questions than it answers.

Another study measured egg production of hens exposed to various frequencies of emf from 260 MHz - 2.435 GHz at levels of power not thought to produce heating effects. All hens except those exposed to 915-MHz radiation layed fewer eggs during the experiment.[12] What this means in practical terms is far from clear.

The next study is perhaps more helpful to us. It was carried out by S.T. Rogers in England in an attempt to estimate the danger (if any) to shipboard personnel exposed for long periods of time to emissions of the shipboard radio. Since exposure under these conditions is mainly under the near field, it had to be determined how this exposure would compare to far-field exposure where the power density may be easily measured.

The present radiation power limit of 10 mW/cm^2 is based on the increase of body temperature of a person by 1-degree centigrade, while one-half the body surface area is exposed to the source of electromag-

*Graphs reprinted from "Radio Frequency Radiation Hazards To Personnel At Frequencies Below 30 MHz" by S. J. Rogers, *Biological Effects And Health Implications Of Microwave Radiation* (Symp. Proc. Med. College of VA, Richmond, VA, Rep. RRH/DBE 70-2) pages 222-232, September, 1969.

netic radiation. This is considered to be the highest safe exposure. When exposed to near-field radiation from a whip antenna, power density measurements become very complex and are difficult to relate to far-field density.

Rogers contends that "the electrical properties of human tissues show that they resemble lossy dielectrics and that any heating due to rf radiation would be a function of the electric component of the field." His theoretical and experimental approach to this subject is complex and elegant, and I refer those with more curiosity and a mathematical inclination to the original article.

Rogers concludes that, to cause a temperature increase of 1 degree centigrade per hour in a test liquid in a near field, a field strength of about 2840 V/m would be necessary. He further states that, to allow for a margin of error, a field strength of 1000 V/m would be a convenient and reasonable limit. To convert this to practical terms, the electric field was measured at various distances from a whip antenna radiating at an output power of 1000 watts at frequencies from 2.1225-21.480 MHz. At all frequencies above 4.455 MHz, the field strength at 1.5 meters (5 feet) from the antenna was less than 100 V/m, but this rose sharply as distance decreased. At 4.455 MHz the field strength at 1.5 meters (5 feet) was exactly 100 V/m, and at 2.1225 MHz it was about 500 V/m.[13]

On the basis of Rogers' findings, there would be relative safety in most situations an Amateur Radio operator may find himself in — the most caution to be observed at the lowest frequencies. It cannot be concluded, however, that distortion of the field by other objects would not focus rf energy to higher intensities than expected.

conclusion

Obviously, much work remains to be done in the field of EMR, its effects on biological systems, and on the safety of those exposed to it. I believe that it will be many years before anyone can say with adequate experimental support that our use of EMR is safe to us and future generations.

I think it safe to say that the lack of clear, nonthermal effects of EMR, despite many studies searching for it, supports the conclusion of the Tri Service program, which in 1960 said that "no data" was obtained to invalidate the safety level of 10 mW/cm^2."

We should remember, however, that distorted rf fields may focus power within objects, and that certain organs, and the fetus, are more susceptible to thermal damage.

I feel fairly secure in the use of Amateur Radio equipment in the way it's commonly employed, *i.e.*, high-power equipment radiating through antennas outside of the shack and some distance in the air, and low-power vhf and uhf transceivers used close to the body.

I urge avoidance of the following situations, due to knowledge of danger or insufficient studies:

1. Avoid high-frequency, high-power equipment with antennas in the shack within 3 meters (10 feet) of living areas.

2. Avoid direct radiation to the eye by a transmitter in the microwave region ("looking down the horn").

3. Avoid prolonged close contact with any antenna radiating more than minimal amounts of energy.

4. Women in the early months of pregnancy, or those who may become pregnant, should avoid contact with strong hf, vhf, and uhf fields.

I believe these are reasonable precautions that should cause no one much hardship, while allowing continued enjoyment of Amateur Radio equipment.

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uhf and microwave frequency counters

A discussion of frequency counters and counting techniques for use above 500 MHz

Today's frequency counters have upper frequency limits ranging from 1 MHz to 24 GHz. Much has been written in the various Amateur Radio journals describing counters that perform up to 500 or 600 MHz, but relatively little has appeared heretofore explaining the techniques by which higher frequencies are measured. This article is intended to supplement one previously published[1] and to explain the methods used today, and in the past, that permit uhf and microwave frequency measurements to be accomplished.

Although most lower-priced counters that can measure frequencies in the 500-MHz region use prescaling, state-of-the-art digital components in use today permit *direct* counting to well over 500 MHz. Frequency counters with ranges greater than this arbitrary, if not completely accurate, 500-MHz limit employ one of the following frequency-extension techniques:

1. Prescaling, which can extend the frequency range to about 1.5 GHz (although indications are that frequencies over 2 GHz will be practical within a year)

2. Transfer-oscillator down-conversion, which can extend the frequency range to over 40 GHz

3. Heterodyne down-conversion, which can extend the frequency range to about 18 GHz.

Prescaling is the simplest and most familiar technique used to extend the range of a direct counter. It entails scaling, or dividing, the input frequency down to one which is within the frequency range of the direct-counting logic in the counter. The dividing factor may be any integral number. If the prescaler is external to the counter, it will usually divide by ten or one hundred, so that the frequency can be read directly from the counter after you have mentally multiplied the counter reading by ten or one hundred, as applicable. If the prescaler is built into the counter, it may scale by any integral factor.

The advantage of using an external prescaler is obvious — it permits extending the frequency range of an existing counter at relatively low cost. Its disadvantages become equally obvious after it has been used. First, there is the necessity of mentally moving the decimal point, since the counter is actually displaying the divided input frequency. Second, one digit of resolution is lost for every decade of scaling. For example, a 900,000.208-kHz signal measured with a scale-by-ten prescaler will read 90,000.021 on a counter having a 1-second gate time (1-Hz resolution). Multiplying by ten yields a frequency of 900,000.21 kHz; the 1-Hz resolution is lost by scaling. It can be re-established only by increasing the gate time by a factor of ten, provided the counter has that capability.

If the prescaler is an integral part of the counter, mentally scaling the frequency and moving the decimal point is eliminated, since this will be accomplished in the counter when the mode is changed from direct count to prescaled count. Nevertheless, the loss of resolution remains. It can be reduced however, by scaling by a factor of less than ten and simultaneously increasing the gate time by the same factor, as shown in **fig. 1**.

Suppose that the internal prescaler divides the

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input frequency by four. If the time-base frequency is also divided by four, the gate time is increased by the same factor and there will be no change in the number of signal pulses gated through to the decade counters. Thus, prescaling is accomplished with only a fourfold increase in gate time and no loss in resolution.

Switching from direct to scaled operation may be carried out in one of three ways. If a single input connector is used, the counter mode is generally manually switched. If two separate input connectors are employed, one for low-frequency signals and the other for high-frequency inputs, the counter mode may be switched either manually or automatically when the input signal is present at the high-frequency port.

Fig. 2 shows the block diagram of a counter which employs automatic switching between separate direct and prescaled input connectors. The switches at the time-base output are actually logic circuits, but are shown as conventional switches to simplify the diagram. If there is no signal applied to the high-frequency input, or if the signal amplitude is below a pre-established level, the switching logic will connect the time-base oscillator directly to the frequency dividers. In that state, the counter will function in its direct-count mode.

When a signal of sufficient amplitude is applied to the high-frequency input, the threshold detector actuates the switching logic to connect the time-base output through the divide-by-N circuit before it reaches the frequency dividers. The prescaler output is fed to an appropriate point in the low-frequency signal conditioner. Thus, the counter is switched to its scaled mode automatically whenever a usable signal is connected to the high-frequency input.

A circuit of this type, which I have incorporated in my homebuilt nine-digit counter, is shown in **fig. 3**. Since the direct-count limitation of the counter is about 200 MHz, it was necessary to scale by eight in order to achieve the design objective of 1300 MHz

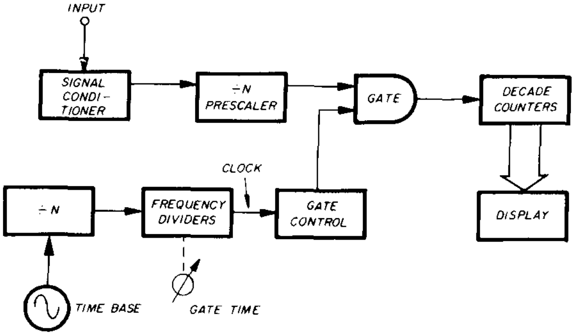


fig. 1. Block diagram of a frequency counter with an internal prescaler.

using available devices. The count limit is actually over 1450 MHz.

The prescaler input impedance is nominally 50 ohms. The use of a 3-dB pad between the input and the Amperex ATF417 amplifier keeps the input VSWR at less than 2.1:1 over its entire frequency range. The sensitivity of the prescaler is between 10 and 25 millivolts rms (depending on frequency) between 100 and 1000 MHz, rising to 100 millivolts at 1300 MHz. The decrease in sensitivity is attributable to two factors. First, the ATF417 is designed to cover the 40- to 860-MHz range, so that its gain drops off from 25 dB at 860 MHz to approximately 15 dB at 1300 MHz. Second, the Motorola MC1697 is guaranteed only to 1000 MHz, although it typically clocks to over 1500 MHz. However, as might be expected, the threshold level increases above 1000 MHz.

As is apparent from the schematic, the amplified signal from the ATF417 is divided by four in the MC1697 and then divided by two in a Fairchild 11C06. The scaled output is coupled through a small capacitor to a suitable point in the low-frequency signal conditioner. Because I did not want to add a negative supply for the ECL integrated circuits in the prescaler, I chose to power the devices from the +5 volt and +24 volt supplies already in the counter.

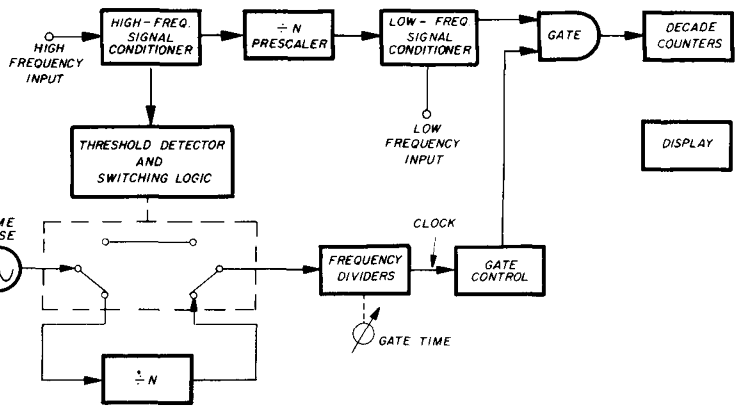


fig. 2. Block diagram of a frequency counter which employs automatic switching from direct to scaled count. The switches shown schematically are actually logic circuits rather than mechanical switches.

This necessitated capacitive coupling between the MC1697 and the 11C06, since the former requires a supply of 6 to 7 volts. Bias at the clock input of the 11C06 is optimized by means of the 2.5-kilohm pot.

Automatic gate-time switching is accomplished by dividing the clock frequency by eight when an input signal of sufficient amplitude is applied to the prescaler input. An LM311 comparator is configured so that when there is no prescaler input, the positive dc at the comparator's noninverting input exceeds that at the inverting input and keeps the output high. This inhibits both the 11C06 (via pin 9) and the 7493 (via pin 3), and also enables a path from the clock input terminal to the clock output terminal through two gate sections of a 7402, which has no effect on the clock frequency.

When an input signal is present at the prescaler input, a portion of the amplified signal from the output of the ATF417 is sampled and applied to a Hewlett-Packard 5082-2835 hot-carrier diode for rectification. The resultant negative dc, applied to the noninverting input of the comparator, causes the comparator output to go low. This enables the 11C06 and the 7493, and inhibits the direct path between the clock input and clock output terminals. The clock frequency is scaled in the divide-by-eight section of the 7493 and applied to the counter logic from the clock output terminal.

The MC1697 is prone to false counting below 100 MHz and when the input signal amplitude is too low. To prevent false readings, the comparator voltage reference is set by a 1-kilohm pot at the inverting input to establish a threshold level below which the prescaler is inhibited and above which erroneous readings will not occur.

transfer-oscillator down-conversion

One of the earliest methods of measuring frequencies in the uhf and microwave regions was by means of the manual transfer oscillator. The transfer oscillator was completely separate from the counter. It consisted of a stable VFO (typically 100 to 200 MHz), a harmonic mixer, and a zero-beat indicator, usually a cathode-ray tube.

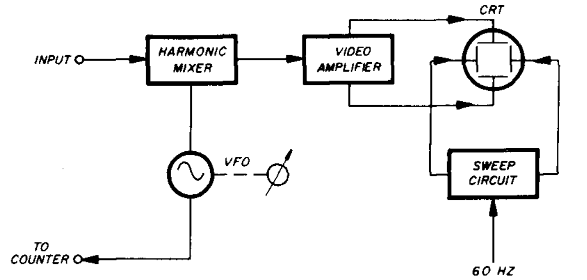


fig. 4. Simplified block diagram of a manual transfer oscillator. The frequency range of the VFO must be within the frequency-measuring range of the counter used in conjunction with the transfer oscillator.

A simplified block diagram of a transfer oscillator is shown in **fig. 4**. The input signal is connected to one input of a harmonic mixer, and one output of the VFO is routed to the other input of the mixer. A second VFO output is connected to the counter, which obviously must be capable of measuring the VFO frequency. The harmonic mixer serves both as a mixer and a harmonic generator, mixing the input signal with the fundamental VFO frequency and with harmonics of the VFO generated within the mixer. The VFO is tuned to the lowest frequency to produce an output from the mixer that is within the passband of the video amplifier. This produces a display on the cathode-ray tube, whose horizontal sweep is usually derived from the ac line frequency. The VFO is then carefully tuned for a zero-beat indication on the CRT, and the fundamental VFO frequency is read on the counter.

If the approximate frequency of the input signal is

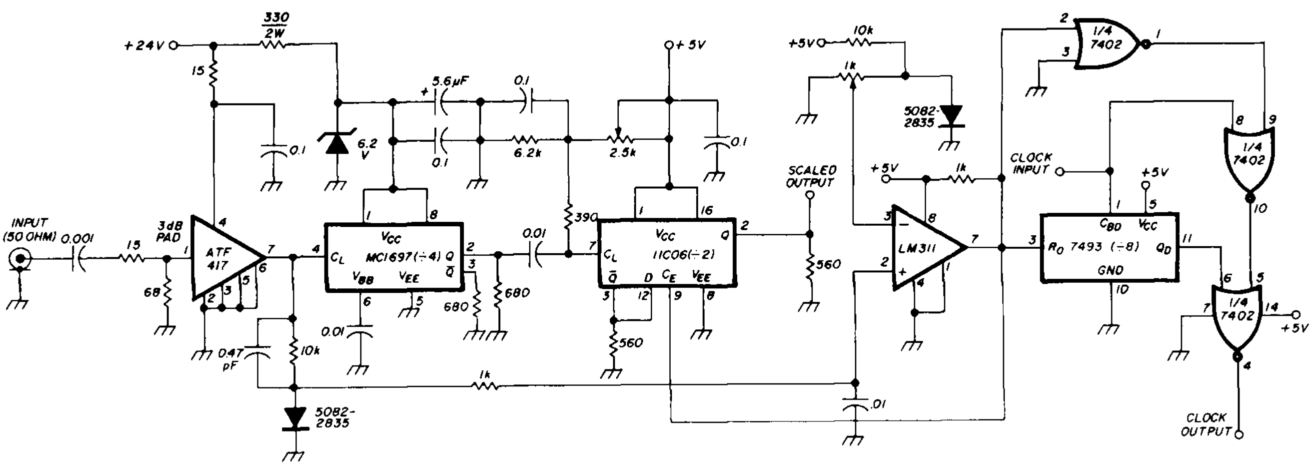


fig. 3. Schematic diagram of a 1300-MHz prescaler which incorporates circuits for automatically switching the gate time when an input signal is applied.

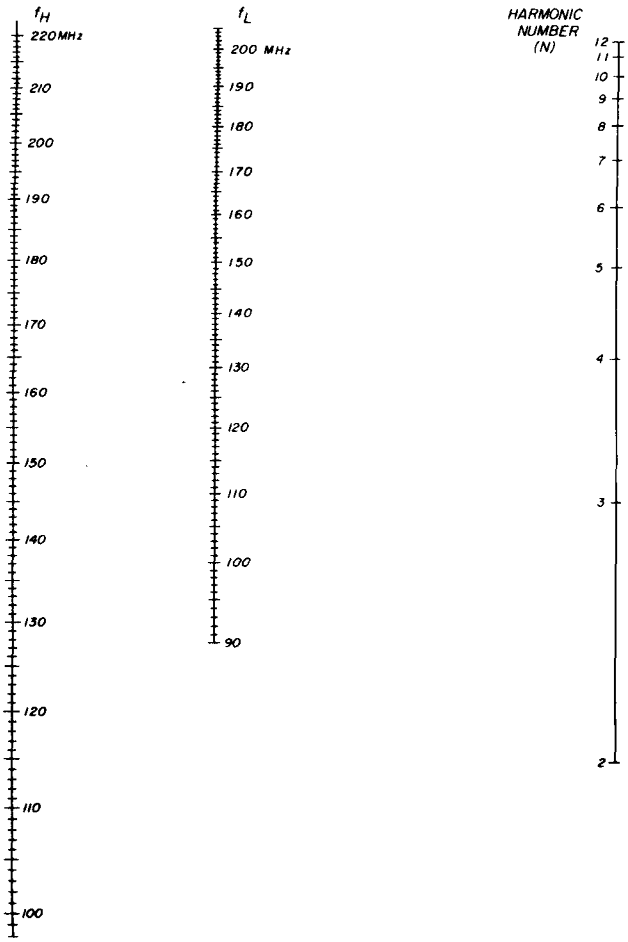


fig. 5. Nomograph for determining the harmonic number of an unknown frequency between 400 MHz and 2 GHz when measured using a manual transfer oscillator whose VFO tunes from 90 to 220 MHz (*courtesy Hewlett-Packard Company*).

known, and it is a relatively low multiple of the VFO frequency so that there is no ambiguity in determining the harmonic number, the input frequency is calculated by multiplying the counter frequency reading by the harmonic number. However, if the unknown frequency is much higher than the VFO frequency, it becomes necessary to determine the VFO harmonic with which the input signal has been mixed. This entails an even more time-consuming procedure of measuring two adjacent fundamental VFO frequencies whose harmonics produce a zero beat, and then determining the input frequency or harmonic number, as follows.

If f_X is the input frequency, f_L is the lower of the two adjacent VFO frequencies, and f_H is the higher VFO frequency, then

$$f_X = \frac{f_H \times f_L}{f_H - f_L}$$

The harmonic number may be determined from the following equations, where N_L is the harmonic number of f_L, and N_H is the harmonic number of f_H

$$N_L = \frac{f_H}{f_H - f_L}$$

$$N_H = \frac{f_L}{f_H - f_L}$$

The harmonic number may also be determined from the nomographs of **figs. 5** and **6** (extracted from reference 2) for the two preceding equations.

The modern transfer-oscillator frequency counter performs essentially the same procedures, but does so automatically. **Fig. 7** is a much simplified block diagram of such a counter. The automatic transfer oscillator consists of two channels, a lock channel and an N-computing channel. The input signal is split in a power divider and applied to one input of the lock harmonic mixer and to one input of the N harmonic mixer. A low-frequency, voltage-controlled oscillator (VCO 1) is swept from its minimum to maximum frequency, typically 100 to 200 MHz, until an output is obtained from the lock harmonic mixer which will pass through the lock video amplifier. The

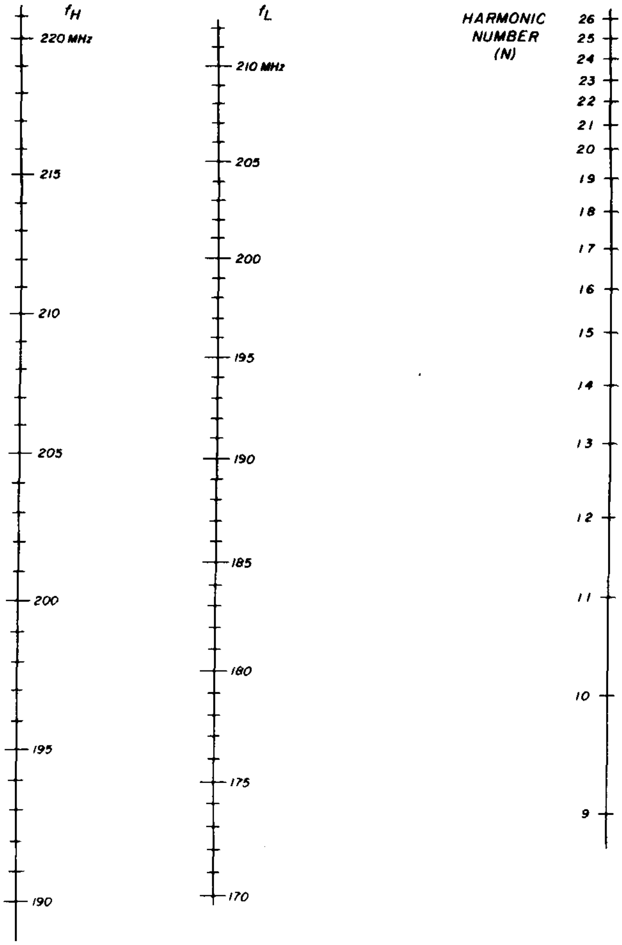


fig. 6. Nomograph for determining the harmonic number of an unknown frequency between 2 and 5 GHz when measured using a manual transfer oscillator whose VFO tunes from 170 to 220 MHz (*courtesy Hewlett-Packard Company*).

signal from the video amplifier is applied to one input of a phase detector, and a reference signal derived from the time base is fed to the other input of the phase detector. Since the output of the phase detector controls the VCO sweep generator, VCO 1 will be phase-locked to the input signal. The output of VCO 1, when so locked, will be 1/N times the input frequency.

A second voltage-controlled oscillator (VCO 2) also provides a signal, via the N harmonic mixer and the N video amplifier, which feeds one input of the N mixer. This signal, when mixed with the reference signal derived from the time-base oscillator, results in an output from the N mixer which is proportional to the harmonic number, N. The N counter then increases the gate time by a factor equal to N. (Note the similarity to the prescaling counter in this respect.) Thus, the counter will provide a direct readout of the input frequency in terms of N times the frequency of VCO 1, whose output is fed to the direct-counting circuits after being converted to the appropriate logic level by the Schmitt trigger.

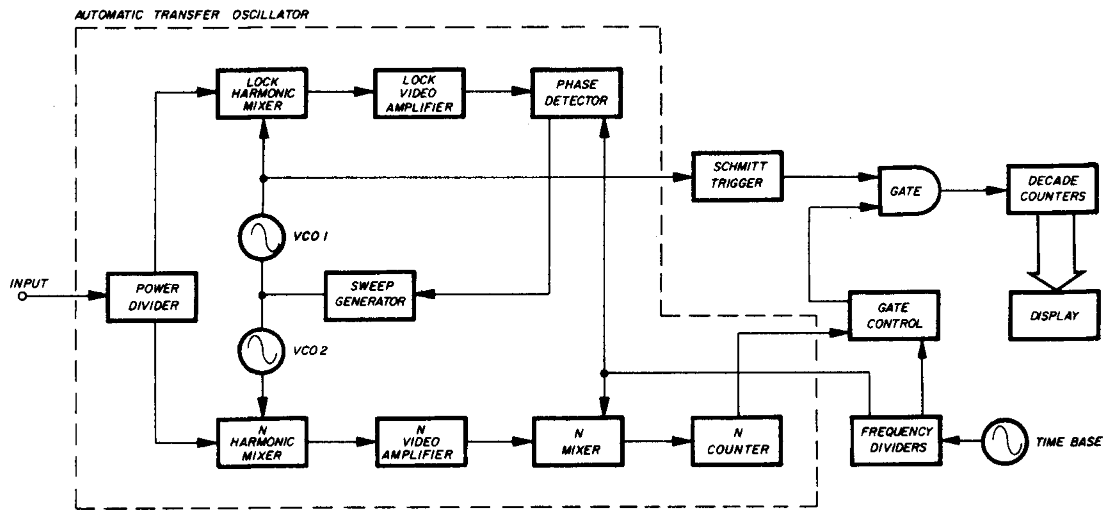


fig. 7. Simplified block diagram of an automatic transfer-oscillator frequency counter.

heterodyne down-conversion

The concept of heterodyning a high input frequency down to one within the range of a low-frequency counter is one that should be completely familiar to anyone with a basic knowledge of electronics. Implementing this concept, however, requires that the heterodyne oscillator frequency be known to the same degree of accuracy as the counter time base if accurate frequency measurements are to result. This is accomplished both in manual heterodyne down-converters and in automatic heterodyne counters by generating the heterodyne frequency from the counter time base.

A simplified block diagram of a typical manual heterodyne converter, which may be either a separate instrument or a plug-in unit, is shown in **fig. 8**. It will accept any frequency between 1.1 and 10.1 GHz and down-convert it to one within the 100-MHz range of the counter connected to its output. Down-conversion is realized by applying the unknown frequency to one input of a microwave mixer, with a known frequency fed to the other input of the mixer.

The known frequency is derived from the time-base oscillator in the counter through frequency multiplier and harmonic generator circuits. The output of the harmonic generator is a comb of frequencies which are multiples of 200 MHz and are fed to the harmonic selector. This circuit is a tunable cavity whose *Q* is high enough to select only a single frequency from the comb input and whose dial is calibrated in terms of the 200-MHz harmonics between 1 and 10 GHz. Obviously, it is possible for the input frequency to heterodyne with either of two adjacent 200-MHz harmonics to produce a beat frequency of less than 100 MHz. However, the lower of the two adjacent harmonics will produce a heterodyne frequency equal to the input frequency minus the harmonic frequency, while the higher harmonic will result in a heterodyne frequency equal to the harmonic frequency minus the input frequency. Since the former will result in a counter reading, which, when added to the selected harmonic frequency is the input frequency, it is the desirable one to use. This is accomplished by always tuning the cavity from the low-frequency end until the indicator shows the first output. The indicator responds to any output from the amplifier which is in the counter's frequency range, so that the lowest harmonic can be selected and harmonic ambiguities eliminated.

Because the tunable frequency is a harmonic of the counter time base, determining the unknown frequency is dependent on only the tuning dial calibration for the selected harmonic; this calibration need only be sufficiently accurate to discriminate between adjacent harmonics. Therefore, as long as you are certain that the lowest harmonic has been selected, operation and frequency determination using a manual heterodyne down-converter is somewhat simpler than the same process involving a manual transfer oscillator.

To iterate a point made previously, the measurement accuracy of the heterodyne conversion process is essentially the same as that of the basic counter because the harmonic frequencies are derived from or phase-locked to the time-base oscillator. Because of the problem of sweeping and selecting the appropriate harmonic, an automatic heterodyne converter became realizable only with the advent of the electrically tuned YIG (*Y*ttrium-*I*ron-*G*arnet) filter.

The YIG filter consists of a single-crystal sphere of yttrium-iron-garnet in a controllable magnetic environment. The ferromagnetic resonance of such a sphere in an rf field can be varied by changing the magnetic field, and therefore can be controlled electrically. An rf signal can pass through the filter when the signal frequency is the same as the ferromagnetic resonant frequency; all other frequencies will be greatly attenuated. Thus the YIG filter is actually the heart of an automatic heterodyne counter, a block diagram of which appears in **fig. 9**.

The counter time base is multiplied and drives a harmonic generator, much the same as in the manual heterodyne converter. The comb output of the harmonic generator feeds the input of the YIG filter, with the filter output applied to one input of a microwave mixer. The unknown input frequency is fed to the second input of the mixer. The filter control circuit drives the control magnet coils in the YIG filter so that

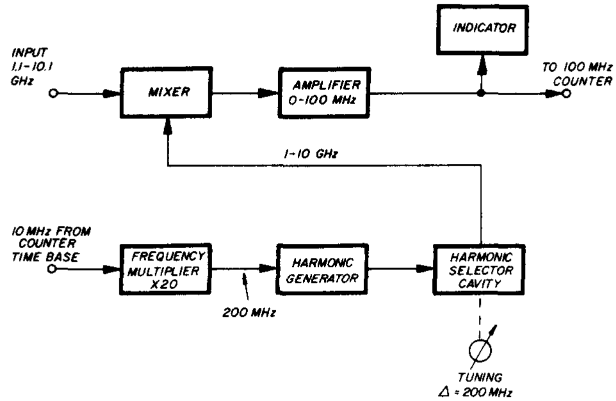


fig. 8. Simplified block diagram of a manual heterodyne down-converter used with a 100-MHz counter to measure frequencies between 1.1 and 10.1 GHz.

the resonant frequency of the filter is swept from its lowest to its highest frequency. The mixer output, generated from the lowest harmonic frequency to pass through the filter, is amplified, converted to an appropriate logic level by the Schmitt trigger, and passed through the gate to the decade counters.

The output of the video amplifier is also applied to an in-band signal detector, whose output inhibits the filter control sweep and keeps the YIG filter at the acquisition frequency. The signal detector also controls the signal acquisition logic, which further controls the filter tuning as required for successive measurements. The signal acquisition logic also controls the gate and other logic circuits (not shown on the block diagram) in the counter so that a direct reading of the input frequency is displayed on the counter readout.

a look at today's technology

Although there are a considerable number of uhf

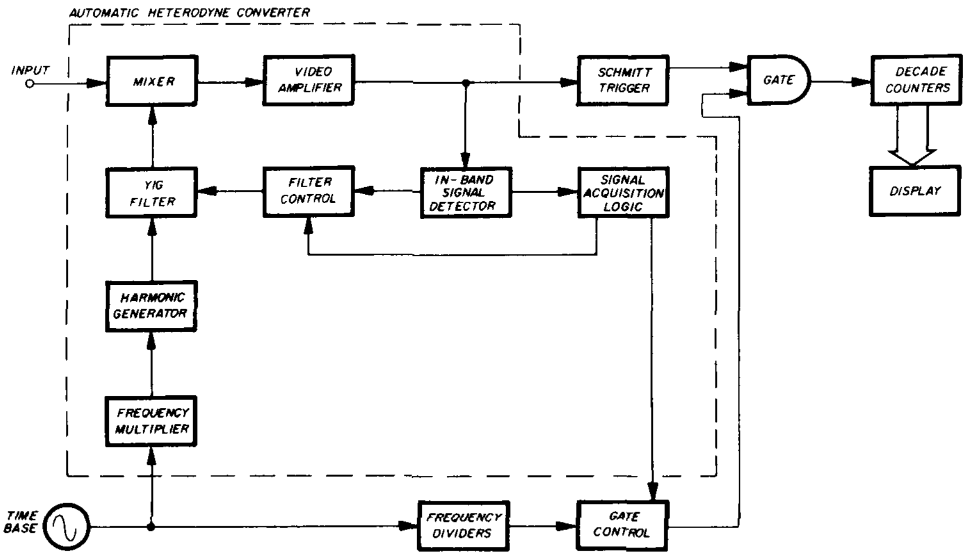
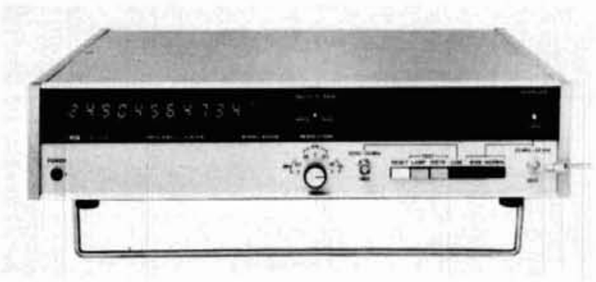


fig. 9. Simplified block diagram of an automatic heterodyne-converter frequency counter. The heart of the system is the electrically-tunable YIG filter.



The Systron-Donner model 6054B employs a FLACTO™ (Frequency Locked Automatic Computing Transfer Oscillator) to permit measurements, with 1-Hz resolution, of frequencies as high as 24 GHz (*courtesy Systron-Donner Corporation*).

counters available in today's market, the over-10-GHz microwave counter field is dominated by three manufacturers: EIP, Hewlett-Packard, and Systron-Donner. Since microwave counters employ the latest technology, a brief look at some typical instruments should be of interest to readers who are not employed in the microwave electronics industry.

The Systron-Donner model 6054B covers a frequency range of 20 Hz to 24 GHz. It is an automatic transfer oscillator type of counter which employs a circuit designated by Systron-Donner as an ACTO™ (Automatic Computing Transfer Oscillator). A simplified block diagram of the ACTO circuit is shown in **fig. 10**. The similarity to **fig. 7** is apparent, with the input signal split and applied to the lock and N channels. However, only a single VCO is used for both frequency and harmonic (N) determination.

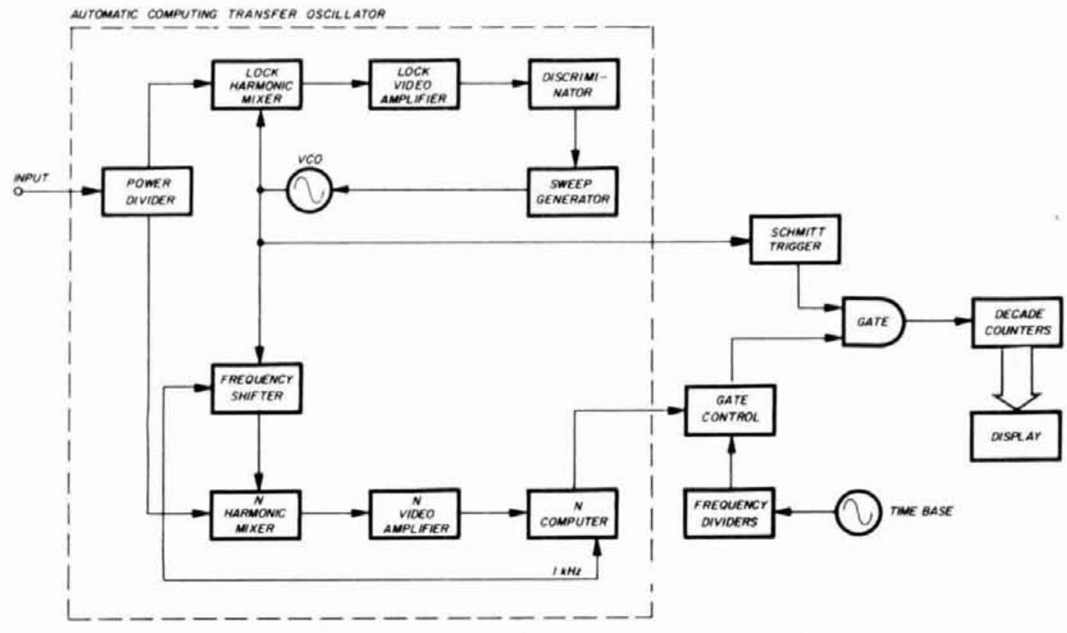


fig. 10. Simplified block diagram of the Systron-Donner ACTO™ (Automatic Computing Transfer Oscillator) down-converter. This is an automatic transfer oscillator which requires only one VCO.

The N-computing channel is used to determine the value of N in the following manner. As can be seen from the block diagram, the VCO output is fed to the frequency shifter, as is a 1-kHz signal derived from the counter time base. The frequency shifter is a single-sideband generator that produces one sideband which is 1 kHz higher than the phase-locked VCO frequency. This frequency-shifted signal is routed to the N harmonic mixer and heterodyned with the unknown input frequency. The resultant mixer output, which will pass through the N video amplifier, has a frequency of N times 1 kHz; this signal is applied to the N computer. The N computer digitally compares the video amplifier output frequency with the 1-kHz reference and generates a signal which corresponds to the harmonic number, N. The signal is further processed and applied to the gate-control circuit to increase the gate time by a factor equal to N. Thus, the counter will provide a direct readout of the input frequency, in terms of N times the phase-locked VCO frequency, which is fed to the direct-counting circuits after being converted to digital levels by the Schmitt trigger.

The counter also employs a Frequency Locked Automatic Transfer Oscillator (FLACTO™), which is a modification of the ACTO technique. The frequency-lock feature permits the counter to tolerate very high levels of frequency modulation and makes the measurement virtually immune to the rate of modulation. Additional details may be found in reference 3.

The model 6054B has two signal inputs. One is a high-impedance, direct-counting input for 20 Hz to

20 MHz; the other is a 50-ohm input for signals between 20 MHz and 24 GHz. When the latter is used, one of two operational modes may be selected. In the normal mode, the local oscillator is locked to an internal reference, which results in a high resolution reading in the shortest period of time (1-Hz resolution for 1-second sampling). In the wide mode, the local oscillator will harmonically track the input frequency, which enables it to track swept or frequency-modulated signals.

A new technique, known as harmonic heterodyne conversion, is used in the Hewlett-Packard model 5342A microwave frequency counter. This conversion scheme is a hybrid of the heterodyne and transfer-oscillator down-conversion circuits in that the counter acquires the input frequency in the manner of a transfer oscillator, but measures the frequency as does a heterodyne converter.

A block diagram of the harmonic heterodyne down-converter appears in **fig. 11**. In this arrangement, the conversion oscillator is a programmable frequency synthesizer locked to the counter time base. The synthesizer output is applied to a sampler, as is the input signal. The microprocessor increments the synthesizer until one of the inputs from the sampler is in the counting range of the direct counter. At that time, the signal detector generates a signal that causes the microprocessor to cease incrementing the synthesizer, and the amplified sampler output frequency is counted; this frequency is the input frequency divided by a harmonic number, N.

To determine N, the microprocessor increments the synthesizer to cause a small frequency change.

Hewlett-Packard's 5342A Microwave Frequency Counter uses a microprocessor-controlled harmonic heterodyne down-converter. When the counter is equipped with its amplitude-measurement option, both frequency and amplitude can be displayed simultaneously. The five left-hand digits are used to display frequency with 1-MHz resolution, and the four right-hand digits display amplitude with 0.1-dBm resolution and a polarity sign (*courtesy Hewlett-Packard Company*).

Hz to 18 MHz, with a recently announced option which extends its upper limit to 24 GHz. Also available as an option is amplitude measurement. This feature allows simultaneous measurement of both the frequency and amplitude of an incoming sine wave. Amplitudes are displayed with a resolution of 0.1 dBm over a dynamic range of −22 to +20 dBm. The amplitude-measuring scheme employs a diode detector circuit in conjunction with an internal reference oscillator for level comparison. The amplitude

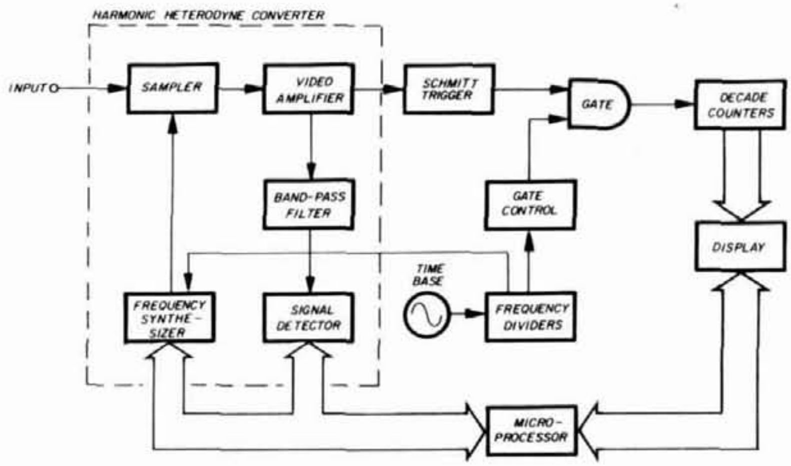


fig. 11. Simplified block diagram of the harmonic heterodyne converter used in the Hewlett-Packard model 5342A Microwave Frequency Counter.

Since there are now two outputs of known frequencies from the sampler, which result from beating the input signal with N times two known frequencies, the microprocessor is able to perform the simple algebraic computation required to determine N.

The 5342A counter covers a frequency range of 10

measurement circuit is calibrated during production and, for signals over 500 MHz, error correction values, as a function of frequency, and input level are stored in an amplitude PROM (programmable read-only memory) for use by the microprocessor. This technique ensures an accuracy of ±1.5 dBm for

The keyboard controls the source-locking circuitry in the EIP model 371 Source Locking Microwave Counter. A separate LED display, located just to the left of the keyboard, displays the desired frequency entered via the keyboard (*courtesy EIP, Inc.*).

sine-wave input signals within the operational dynamic range.

As can be seen from the photograph of the Hewlett-Packard model 5342A, operation of the counter is controlled by means of a front-panel keyboard. The keyboard provides control of resolution, self-check, automatic or manual modes, amplitude and/or frequency measurements (with the amplitude option installed), frequency and amplitude offset, etc. Such is the power of the microprocessor-controlled instrument. A detailed discussion of the model 5342A appears in reference 4.

A unique instrument manufactured by EIP is their model 371 source-locking microwave counter. This counter is an automatic heterodyning type that covers a range of 20 Hz to 18 GHz, and, in addition, has the ability of locking any signal source between 10 MHz and 18 GHz to the same long-term accuracy and stability as the time-base oscillator in the counter. The only requirements for the signal source are that it have an fm input and that it can be set manually to within 20 MHz of the desired output frequency. A block diagram of the source-locking circuits is shown in **fig. 12**.

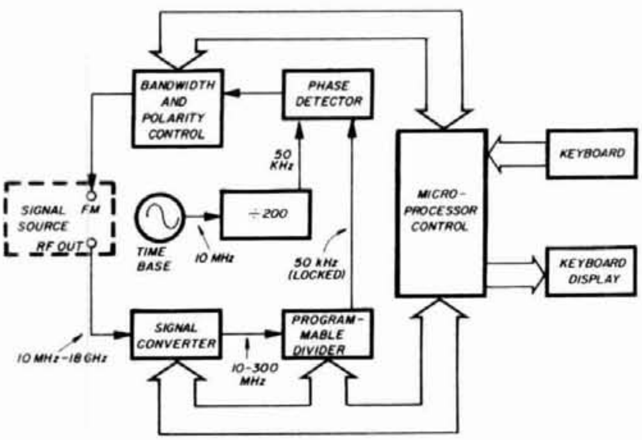


fig. 12. Block diagram of the source-locking section of the EIP model 371 Source Locking Microwave Counter. The frequency of an external signal source can be locked to a preset frequency between 10 MHz and 18 GHz with the same long-term stability and accuracy as the time-base oscillator in the counter.

Source locking is accomplished by converting the input signal to one that is in the 10- 300-MHz range, using heterodyne down-conversion. The microprocessor control then calculates the proper division ratio to produce a 50-kHz output from the programmable divider when the input signal is equal to the desired frequency, which has been entered via the keyboard. (An auxiliary keyboard display on the counter records the frequency which has been keyboarded in). The dc component of the phase detector output, applied to the fm input of the signal source via the bandwidth and polarity control circuit, alters the frequency of the signal source until it is equal to the desired frequency.

The microprocessor controls the overall loop response by systematically varying the bandwidth and polarity parameters until a phase lock is achieved at a nominal bandwidth of 2 kHz. If the loop cannot be locked at this bandwidth, because of inherently low bandwidth in the signal source, the microprocessor repeats the process at a nominal bandwidth of 500 kHz. The automatic bandwidth and polarity control permits the use of the source-locking counter with signal generators and sweepers of different modulation sensitivities and polarities.

summary

This has been a necessarily brief overview of uhf and microwave counters. I have intentionally omitted a comparison of the several down-conversion systems being used today, since such comparisons are often a matter of specsmanship. Readers who are interested in such comparisons, or who want more detailed information on the conversion techniques used by the manufacturers dominant in the field, should consult references 3 through 7.

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3. Dick Voyles and Dick Hall, "Microwave Frequency Measurement — A New Approach," Technical Article No. 14/2, 1975 IEEE Intercon Conference Record, *Trends in Electronic Measuring Instruments*.
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ham radio

vertical antenna

for 40 and 75 meters

Design and construction of a two-band vertical antenna that fits into a modest city lot

Many Amateurs live on small lots in crowded neighborhoods and don't have the space for full-size horizontal antennas for the low-frequency bands. For example, my lot is small and cut into the side of a hill. The house and a swimming pool occupy the only flat area. A few years ago, during the sunspot doldrums, I became interested in working some low-band DX with emphasis on 75 meter SSB. The two-band vertical antenna described here is the result of my experiments.

First I tried a two-band inverted V hung from my beam antenna tower. I had little success competing on 75 SSB, although the antenna worked fairly well on 40 meters. It was also a bother whenever I had to lower the tower because of weather. The next antenna considered was a ground-mounted vertical cut to one-quarter wavelength on 75 meters. This design became very unattractive for a number of reasons. Very little free ground area was available for a good radial system, and the ground is exceedingly hard with low-conductivity soil. The only available site locations were either difficult for running coax or were in locations where a considerable amount of the radiation would be into my house and those of neighbors. About the only place left to consider was the top of the house, which is about 9 × 12 meters (30 × 40 feet).

This led to the design of an inexpensive two-band groundplane vertical antenna for 40 and 75 meters mounted on the top of the gable roof. The antenna was made from a three-section push-up TV mast, about 8 meters (25 feet) of RG-8/U coax, some galvanized TV guy wire, a TV-type ceramic pot capacitor, a short piece of stair railing dowel, a few insulators, and 61 meters (200 feet) of almost any kind of copper wire for two sets of radials. The antenna is shown in **fig. 1**. A simple fixed-tuned L network was

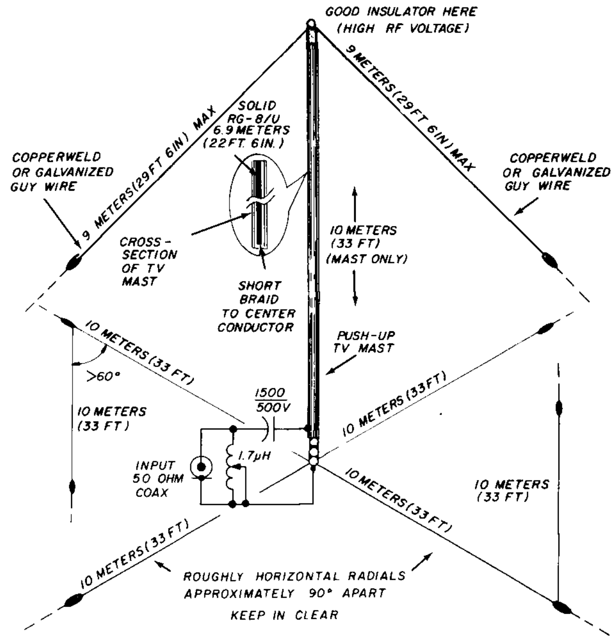

fig. 1. Two-band vertical antenna for 40 and 75 meters.

mounted at the base to obtain a good impedance match on both bands. Two sets of two radials were used, one straight set for 40 meters and a Z configuration for 75 meters.

operating principles

The TV mast (**fig. 1**) is one-quarter wavelength long on 40 meters. From the top a length of RG-8/U

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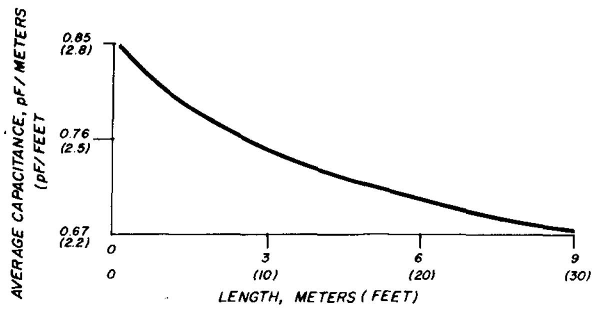

fig. 2. Capacitance of wire in space as a function of wire length. To find total capacitance, multiply length by capacitance per meter (foot). Curve assumes length-to-diameter ratio greater than 50. Curve was used to determine capacitance of the antenna top-hat radials.

coax is dropped, which is shorted at the bottom end. The top outer conductor (shield) is connected to the top of the TV mast. The top center conductor is connected to two slanted radials, which act as guy wires and the capacitive-loading element, or "top hat."

The coax on 40 meters appears as a parallel-resonant circuit and isolates the mast from the top-hat radials. On 75 meters the coax is one-eighth wavelength long and acts as a series inductance of Z_0, or 50 ohms. The base section has a characteristic impedance* of about 350 ohms. Accordingly it appears as an inductive reactance on the order of 350 ohms. The top radials act as capacitance loading and have an effective capacitance of about 100 pF with a reactance of about 400 ohms (**fig. 2**). The mast, coax, and capacitive top hat form a series-resonant circuit on 75 meters, allowing the mast to be an effective one-eighth-wavelength radiator with a fairly flat current distribution. Note from **fig. 3** that the current distribution along the mast on 75 meters is fairly uniform.

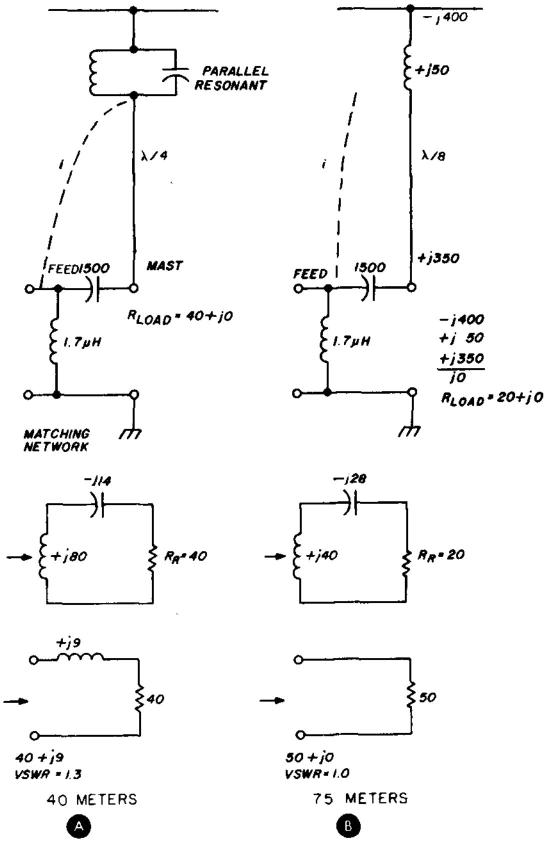

fig. 3. Equivalent circuit of the two-band antenna. The 40-meter version is shown in (A), 75-meter version in (B).

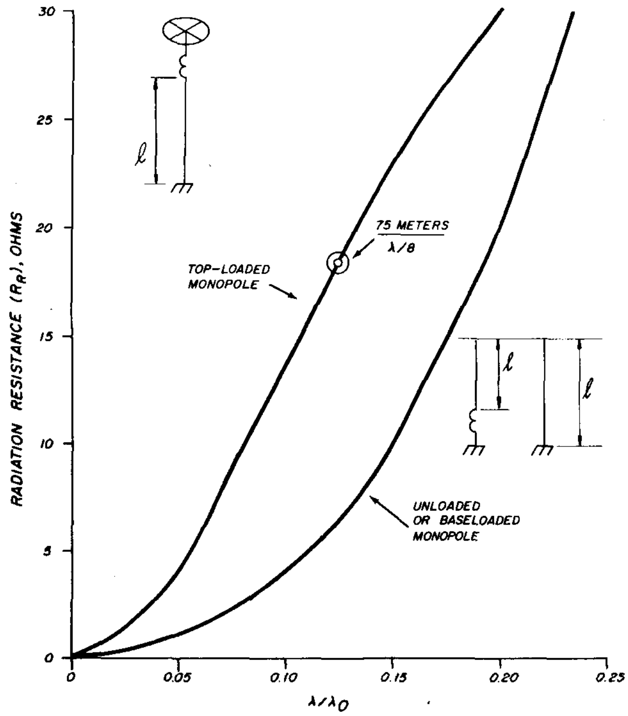

fig. 4. Radiation resistance of short monopole antennas, which was used to derive base resistance of the antenna L matching network.

Fig. 4 is used to derive the base resistance for the design of the L matching network. On 40 meters the antenna is one-quarter wavelength long and has an input impedance in the order of 40 ohms. On 75 meters, because of top loading, the base impedance is about 18 ohms. If base loading were used, one-eighth wavelength on 75 meters would have a base impedance of only 7 ohms and would be inefficient.

*Characteristic impedance is

$$\sqrt{\frac{\text{inductance per unit length}}{\text{capacitance per unit length}}}$$ and

is approximately $60(\ln \frac{2h}{d} - 1)$, where h is height and d is diameter.

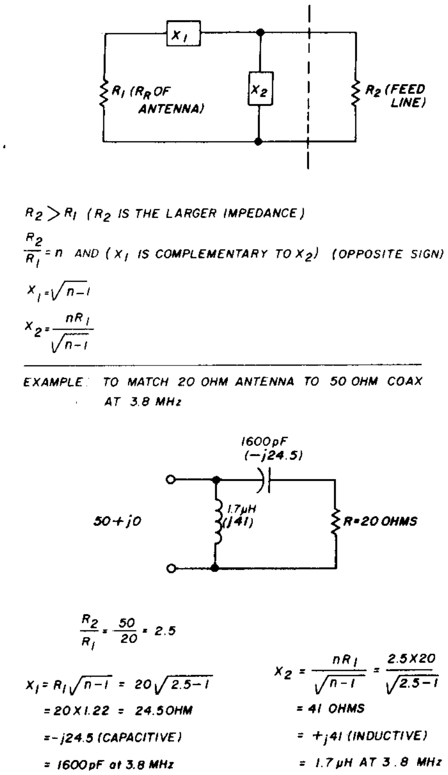

fig. 5. L network matching circuit development.

The L network design is developed in **fig. 5**. The resultant circuit is derived in **fig. 3**. The measured input matching characteristics are shown in **fig. 6**.

adjustment

Adjustment is straightforward. Little interaction occurs between 40- and 75-meter adjustments. First adjust the 40-meter radials, the two straight 10-meter (33-foot) lengths equally until a minimum VSWR is obtained at the desired operating frequency. Next, for 75-meter operation, adjust the Z-configuration radials equally. If this doesn't quite hit the desired frequency a slight adjustment of the top-hat radials may be necessary. These radials don't affect 40-meter operation to any extent. Recheck 40 meter operation. The VSWR on 40 meters will not be unity, because the base impedance is on the order of 40 ohms. The matching network may need a slight inductance change for the best match on 75 meters. This adjustment will have negligible effect on 40 meters.

construction

Details are shown in **figs. 1** and **7**. The top of the mast on 40 meters is at a very high voltage, so a good-quality top insulator is needed. I used low-sap wood (maple) boiled in wax. The insulator has been reliable over the past four years. Plastic materials such as acrylic are suitable. See **fig. 7** for top insulator assembly. Each end of the coax cable mounted inside the mast should be dipped in wax or otherwise sealed.

The two top radials are made from galvanized or copperweld guy wire. Soft copper wire was used originally but broke at the top end as a result of wind stress. The top radials are at high voltage on 75 meters but have low current. They are not used on 40 meters. Accordingly, it's not necessary to use high-conductivity wire. The lower end of the radials terminate in small, corrugated antenna insulators. The wire is cut longer than shown, passed through the insulator, and twisted back on itself so that easy adjustment may be made. The vertical angle of the radials isn't critical. An anticorona noise loop is formed at the top of the mast either by extending the coax or one of the top radials.

The base-mounting insulator can also be made from wood boiled in wax. This point is at low voltage and insulation isn't critical. I used a tilt-over, U-channel TV base for convenience in mounting. The mast should be supported by one set of four insulated guys at the top of the base section of the three-section TV push-up mast. Guy rings are usually supplied with the mast.

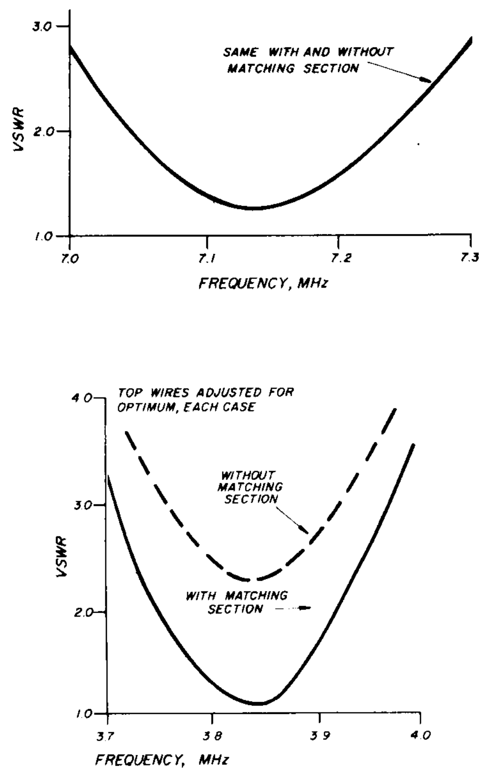

fig. 6. Measured input matching characteristics for the two-band antenna, 40 meters (top) and 75 meters (bottom).

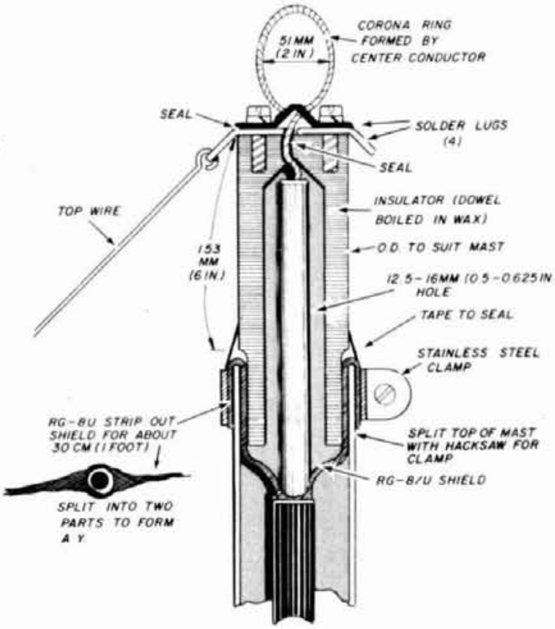

fig. 7. Construction details of the top insulator.

The L matching network is mounted on the base of the mast above the mounting insulator. A heavy, flexible, stranded wire is run from the bottom side of the L network to the center of the groundplane radials. This makes it convenient to tilt the mast without disassembly of the network or feed coax. The 1.7-μH inductor is ten turns of 1.6-mm (no. 14) bare wire, 38 mm (1 1/2 inches) diameter and 35 mm (1 3/8 inches) long. Spacing is six turns per 150 mm (six turns per inch).

The groundplane radials form resonant elements and should be separated from surrounding surfaces except at their center. The vertical angle of the radials is not critical. The ends are supported and terminated by insulators similar to the top radials for convenience of adjustment. Bend the radials to suit the shape of roof. The bend at the 10-meter (33-foot) point can vary slightly. The wire type isn't critical. Anything larger than 1 mm (no. 18) either enameled or covered will suffice.

performance

The antenna has been in use since 1974. Operation has been satisfactory. The only mode used on 75 meters was SSB, with some CW on 40 meters. All continents except Europe have been worked several times on 75 meters, with good reports. If you live in a high noise location, this may not be the antenna for you. If you live in a place where lightning is active, make sure an adequate ground is provided. This antenna makes a dandy lightning rod.

ham radio

updating the Collins KWM-2

Important modifications are described for modernizing the KWM-2 high-frequency transceiver

Introduced to the Amateur world in the fall of 1959, the Collins high-frequency KWM-2 transceiver quickly became the classic, with over 40,000 units in use worldwide by Amateurs, commercial services, and the military of numerous nations. The latest version of this popular rig is the KWM-2A. Time-proven by its robust construction and its long life in these days when circuit-boarded, solid-state gear quickly eliminates obsolete designs, this fine transceiver has more than held its own. Over the years revisions have been made to the original design. This article covers some important modifications to the KWM-2 family and describes how you can incorporate them into your unit to help bring it up to date.

the KWM-2

The Collins KWM-2 high-frequency transceiver is widely recognized as a superior piece of Amateur gear and is continuing a long and useful life. A decade ago a military overview of communications equipment in governmental service praised the KWM-2 for reliability, ruggedness, and ease of repair. Countless thousands of Amateurs agree with this conclusion.*

*Gus Browning, W4BPD, tells the story of a KWM-2 he took along on a DXpedition in the Indian Ocean. It was dropped overboard by a crew member during an attempt to land on an obscure island. Native divers finally fished up the KWM-2 and brought it ashore. After flushing with fresh water and drying out for a few hours, the rig was hooked up — and it worked! The only casualty was the meter movement, which had opened up. A local artisan repaired the meter winding and Gus was back on the air.

While the newest KWM-2s retain the original classic appearance, numerous revisions and modifications have been incorporated over the years which make the modern version easier and better to operate than the older sets. Some of the important modifications that can be made by the advanced Amateur with adequate test equipment are described here. For those who don't want to dig into their transceiver, information is furnished on getting the more sophisticated and difficult modifications made by a professional. In any event, before undertaking any revision or modification to your KWM-2, make sure the change has not already been incorporated into your equipment. Many hams own second-hand units, so it's wise to make sure your manual agrees with the particular transceiver you own, at least as far as the schematic and voltage charts are concerned. Be suspicious of an older model KWM-2 that has a new manual. The two may not be in exact agreement.

All modifications should be made with a 40-watt (or smaller) soldering iron, so as to protect the insulation on wires next to the soldering iron. A magnifying glass is helpful, as are needle-nose pliers. You'll be working in an area with a high parts density and you don't want to damage some circuits while you modify others!

the "wing" versus the "meatball"

Around mid 1968, Collins changed their old winged emblem and adopted a new, round escutcheon known as the "meatball." This cosmetic change allows you to determine the approximate age of your

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KWM-2, as the random serial numbers on the KWM-2 after 1968 no longer date the equipment for the layman. On the used-equipment market, the "meatball" KWM-2 commands a somewhat higher price than the older "wing" model. It's best to buy the KWM-2 on performance and appearance, however, and forget about the emblem. Sometimes you can realize a tidy savings by buying a "wing" model in good condition rather than the newer "meatball" model.

If you do buy a used KWM-2, check it carefully in both transmit and receive modes on all bands before you part with your money. Look under the chassis to make sure the previous owner hasn't made his own unique (and often unworkable) modifications. Many good KWM-2s for sale are showing up in the classified ads, as bedazzled hams trade in their units for the latest solid-state transceiver complete with bells, whistles, and a six-month wait for replacement parts. Good! Their loss is your gain if you want to own a rugged and reliable transceiver that you can service and repair yourself.

minor bugs you may not have observed the first time around

Transmitter instability? Signs of oscillation? Before you tear things apart or attempt reneutralization of the amplifier stage, remove the amplifier-compartment lid and make sure the tube shield of the 6CL6 driver stage (V8), is firmly in place. A loose tube shield can play havoc with transmitter operation!

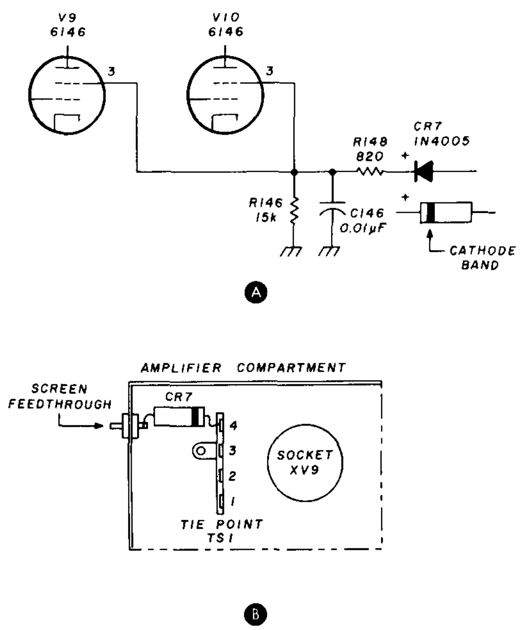

fig. 1. Diode CR7 placed in amplifier screen circuit protects receiver from blocking caused by screen emission of 6146 tubes. (A) circuit modification, (B) diode placement.

Receiver blocking on switch-over? Sometimes you'll notice a delay of up to 30 seconds during which time the receiver is blocked and no signals are heard after switch-over from the transmit mode. This problem is caused by screen emission from the 6146 amplifier tubes (V9), (V10), which paralyzes the receiver agc (automatic gain control) circuit. New 6146* tubes will sometimes cure this annoying problem, but a permanent fix is easily achieved by placing a diode in the amplifier screen power lead, which blocks negative current (**fig. 1**). This mod is easily and quickly made in the bottom of the amplifier compartment. The diode is substituted for the wire lead between the screen feedthrough terminal in the compartment wall and nearby socket tie-point strip (TS1). The diode anode is connected to the feedthrough terminal. Put insulated sleeving on the diode leads. This mod has no effect on transmitter performance.

ALC meter instability? Does the zero reading of the alc meter float around during transmit, or does it gradually drift up-scale as the KWM-2 warms up? This annoying fault can usually be cured by replacing capacitor C157 (0.01 µF, 200 volts) with a new low-leakage mylar or polypropylene unit. You'll find the old unit attached to pins 1 and 3 of socket XV17A (6BN8).

Equipment runs hot? Short tube life? The 6U8/6U8A and 6AZ8 tubes in the KWM-2 are said to have short lives. The grapevine suggests replacing the 6U8/6U8A with a 6EA8 for longer life. This can be done in most sockets, with no change in performance, except for the 6U8/6U8A used as the audio tone oscillator (V1). Some 6EA8s will not work in this circuit, and others will distort the audio tone signal, which then bleeds into the receiver audio system during CW operation. Stick with the 6U8/6U8A in this socket and look for short-life tube problems elsewhere.

In some KWM-2s the low-voltage dc supply (supposed to be a nominal 275 volts) runs from 300 to over 340 volts when the standard Collins 516-F2 power supply is used.† No wonder some of the small tubes are cooked! Measure your low-voltage supply. It should not run much over 290 volts on receive and

*Folklore has it that either 6146B tubes won't perform properly in the KWM-2, or that 6146Bs are the only tubes to use in the KWM-2. Forget both of these fairy tales. The differences between the 6146, 6146A, and 6146B are minimal (mostly being one-upmanship in advertising policy). All do the job equally well. It's not necessary to match 6146-type tubes, either, although it's suggested that a 6146A not be used with a 6146B.

†Overvoltage is presumed due to various manufacturers having supplied the power transformer and filter chokes. Design and windings of these components seem to vary, especially in the dc resistance of the transformer or choke coils. This could account for the voltage variance.

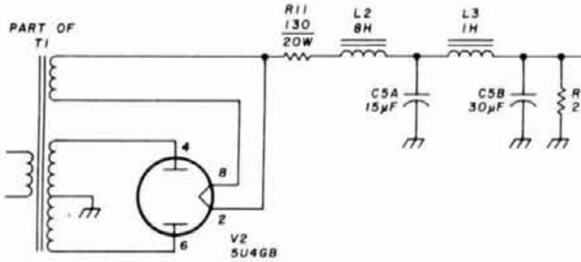

fig. 2. Voltage-dropping resistor R11 added to reduce low voltage to small tubes in KWM-2. A heat-dissipating resistor in a finned housing is recommended. This is bolted to the chassis sidewall near the 5R4GY socket. A Dale type RH-25 resistor or equivalent is suggested.

260 volts on transmit. If the voltage is much higher than these values, add a 75-150 ohm, 25-watt wirewound dropping resistor, R11, as shown in **fig. 2**. The small tubes in the KWM-2 will run much cooler if you do this. The 5U4GB rectifier in the power supply should be replaced with a solid-state plug-in rectifier, and the resistance value of R11 should be chosen to deliver the correct voltage. Substitution of the rectifier improves regulation and removes 15 watts of filament power from the supply transformer.

You should also replace the 5R4GY high-voltage rectifier with a suitable solid-state plug-in device. This action will remove an additional 10 watts of filament power from the transformer and will increase the B-plus voltage by about 40 volts, providing a few more watts of power output and a cooler-running transformer. This simple substitution also boosts the 6.3-volt filament supply, which is marginal at best. You'll probably have to readjust the amplifier bias control, R9, in the supply for the correct resting plate current of the amplifier tubes after these mods have been made (40 mA for general use or 50 mA when driving a linear amplifier).

Old filter caps in the power supply? It's a good idea to replace the high-voltage filter capacitors and the bias filter capacitor in the power supply if the KWM-2 is an older model. The capacitors become leaky with age and the capacitance value drops off at the same time. You can put more microfarads in the same space occupied by the old units and this improves the supply's dynamic stability. When you put the new capacitors into the circuit be sure to observe polarity, for the bias capacitor, which is hooked up "backwards," with the positive terminal grounded. Capacitors C2, C3, and C4 can be replaced with equivalent 80-μF, 450-volt units, and C5A-B can be replaced with a dual 30-μF, 250-volt unit. Capacitors C6, C7 can be replaced with 40-μF, 250 volt units. Unless the shunt capacitor, C1, is defective (a rare occurence), don't bother to replace it.

Dial chatter or backlash? Underneath the VOX plate atop the main tuning dial assembly is a small idler pulley mounted to the front panel to the left of the dial mechanism (as viewed from the front). This pulley holds the two dial plates in alignment as the dial is rotated. Unbolt and lift up the VOX plate; this requires removal of one screw at the top left of the plate and two screws above the panel escutcheon. Now you can see the dial pulley. If it's loose it will rattle, and the dial will show backlash to a greater or lesser degree. The amount of mesh with the dial mechanism is determined by the center screw holding the gear. For a quick fix, loosen the screw and slide the gear into the dial mechanism a very small amount and retighten. Caution! The gear-retaining screw is very short. Don't loosen it too much or it will

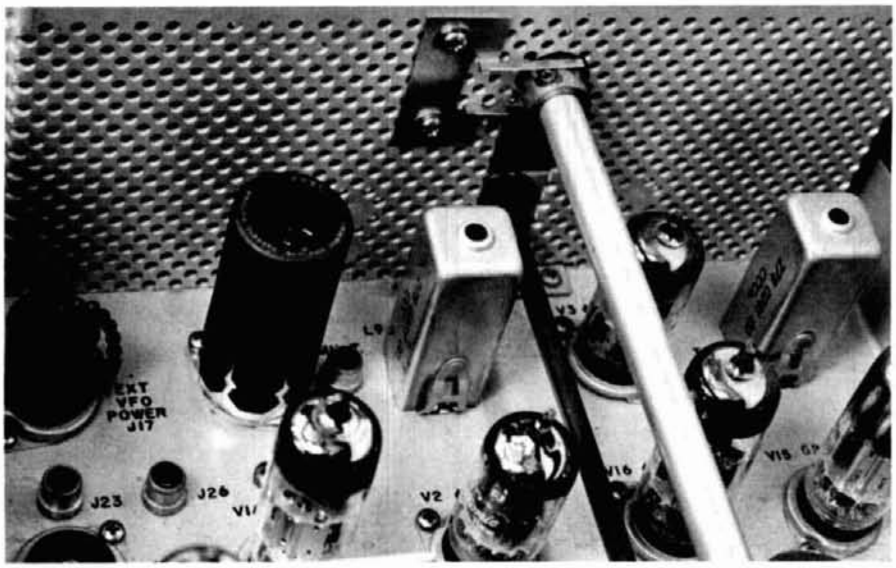

fig. 3. Grounding clip on amplifier loading shaft reduces harmonics escaping from amplifier compartment. Tube shield on V1 (6AZ8) decreases tube temperature. Heat sink shield is used.

fall out and you'll lose dial alignment. However, if you hold the two dial disks together to keep them from losing alignment, you can completely remove the idler gear and coat the gear shaft with silicone grease, which will *eliminate dial rattle*. Maintain the position of the dial plates so that you don't lose calibration.

TVI on 10 meters? Why do some KWM-2s show bad TVI on 10 meters while others don't? And why does the TVI often worsen when you bring your hand near the final amplifier tuning/loading panel controls? The answer is that these concentric shafts come out of the amplifier compartment and are insulated from the front panel of the KWM-2 by an almost invisible panel bushing. In effect, the shafts *act like a radiating antenna* for amplifier harmonics that would otherwise remain bottled up in the amplifier compartment. A shaft grounding clip* bolted to the outside of the amplifier enclosure (as shown in **fig. 3**) grounds the outer shaft and reduces the harmonic signal at this escape point to near zero. The grounding clip is held in position with (4-40) hardware.

If your KWM-2 doesn't incorporate a vhf choke (L128) in the power amplifier B-plus lead immediately following plate choke L17, a 120-μH choke should be added to prevent harmonic currents from passing into the power supply (**fig. 4**).

Receiver i-f tube V1B run hot? Place your hand on V1, the 6AZ8 i-f amplifier tube after the KWM-2 has been running for a few hours. Wow! Hot! No wonder this tube is said to have a very short operating life. And no wonder the S-meter zero-signal reading shifts about on the scale. The latest versions of the KWM-2 have incorporated a protective resistor (R75) in series with pin 3 (cathode) of tube V1 to ground to limit plate current. If you don't have this resistor in the circuit, a 10-ohm, 1/2-watt resistor placed in series with the ungrounded terminal of the receiver GAIN ADJUST potentiometer, R132, mounted on the VOX plate, will help reduce the tube temperature. In addition a heatsink-style tube shield† is placed over V1.

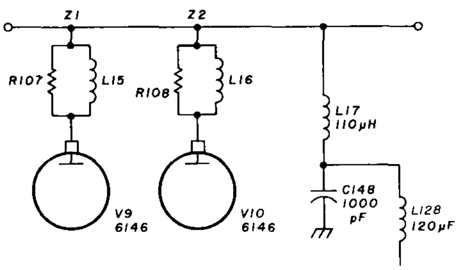

fig. 4. Vhf choke (L128) in B-plus lead to final amplifier helps suppress TVI-causing harmonics. A J.W. Miller 9360-13 choke rated at 400 mA is suggested.

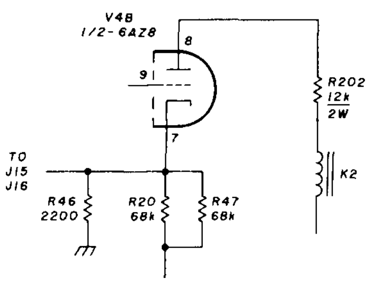

fig. 5. Modified VOX relay control circuit. Resistor R202 is added to reduce current through the relay coil. It may be necessary to reduce the value of the resistor in some cases to provide proper pull-in current. In some KWM-2s resistor R202 is 330 ohms and is located in the cathode circuit between pin 7 of socket XV4 and the circuit to J15 and J16. In this case, no plate resistor is required. In some units resistor R46 is 3.3k. It should be replaced with 2.2k for this modification.

Heat-sink shields are hard to come by, but perhaps your friendly electronics store (or the local flea market) has some. A retainer mounting shell is also required. The shell is mounted to socket XV1 using the existing mounting bolts. You'll probably find (as I did) that the mounting shell has a negative clearance with respect to the socket. The solution is to cut tiny slots around the bottom edge of the shell with metal snips. Cut to a depth of about 1.5 mm (1/16 inch) then bend out the tabs you've made with a pair of long-nose pliers. The shell will then fit snugly over the socket rim. Snap the heat-sink shield over the tube, and longer tube life will be your reward.

Relay problems? Some KWM-2 owners have found to their sorrow that the coil of VOX relay K2 burns out after prolonged use. The popular and expensive solution is to get a "meatball" KWM-2 with plug-in relays. However, a circuit modification somewhere along the long production history of the KWM-2 has solved this vexing problem, even in some of the older models. A 12k, 2-watt safety resistor (R202) is placed in series with the plate of the VOX relay amplifier tube, V4B, **fig. 5**. If your KWM-2 doesn't have this modification it's a good idea to incorporate it, as it *might save you a destroyed relay coil*. The resistor can be mounted between pin 8 of socket XV4 and a tie-point epoxied to the chassis near the socket.

Lack of receiver sensitivity on some bands? Even after repeated alignment some KWM-2s show

*The Collins part number of the grounding clip is 553-2555-002. You may be able to obtain a clip from Dennis Brothers, WAØCBK, Route 1, Box 1, Potter, Nebraska 69156.

†Suitable heat-dissipating tube shields are manufactured by, and available from, International Electronic Research Corporation, 135 West Magnolia Boulevard, Burbank, California 91502. The shield cools tube bulb temperature to below that of the bare bulb. A type TR6-6020B shield is used for the 6AZ8 or 6U8/6U8A. A TR6-6025B is recommended for use with the 6CL6 driver tube.

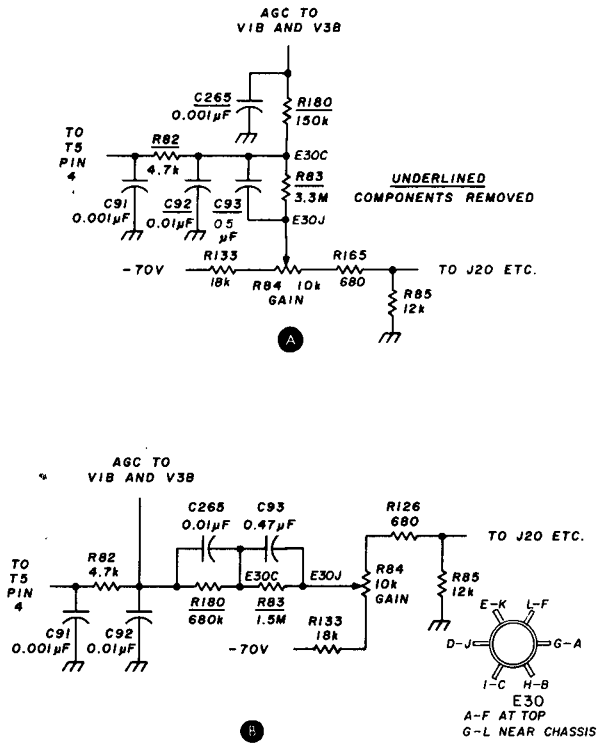

fig. 6. KWM-2 mods to improve agc response. Original circuit is shown in (A). Circuit in (B) is the modified setup as described in the text. Terminal support E30 is adjacent to the audio output transformer T6. Terminal TS8 is equidistant from E30 and T6.

poor sensitivity on some bands, or sensitivity seems to change when sidebands are switched. This problem can be caused by signal overload from low-frequency sideband oscillator V11A, whose signal is coupled into the receiver section wiring harness.

The culprit is rf choke L22 in the crystal-oscillator plate circuit. The choke is part of a tuned circuit and can radiate energy furiously. Radiation from this inductor can get into circuits where it doesn't belong and reduce overall receiver sensitivity. The cure is to remove choke L22 and replace it with a shielded rf choke. A parasitic suppressor (R195, 47 ohms, 1/2 watt) should also be placed in series with the grid, pin 2, of socket XV11A.

Receiver agc pumping and overshoot on noise pulses? Some of the older model KWM-2s use the agc time constant circuit shown in **fig. 6A**. A newer circuit is also shown in this illustration. The components to be changed are on terminal support E30, shown in the technical manual. These are R82 (4.7k), R83 (3.3 meg), C92 (0.01 μF) and C93 (0.05 μF). In addition, R180 (150k) from terminal TS8-1 to E302, and C265 (0.001 μF) from TS8-1 to the power-connector grounding ring are removed. (Note that R180 and C265 are not incorporated in some early models.)

The following components are now added:

1. Connect new R83 (1.5 meg) from E30C to E30J.
2. Connect new C93 (0.47 μF) from E30C to E30J.
3. Connect new R82 (4.7k) from T-5 terminal 4 to TS8-1. Use sleeving on leads and route around E30.
4. Connect new R180 (680k) from TS8-1 to E30C.
5. Connect new C265 (0.01 μF) from TS8-1 to E30C.
6. Connect new C92 (0.01 μF) from TS8-1 to ground ring on power connector J13. Check wiring against **fig. 6B**. Mark the modification in your manual for reference.

Agc overload and audio distortion on strong SSB signals? It's recommended that this useful modification be performed along with the previous one in cases where both arrangements are missing from the transceiver. This modification adds hang agc to the receiver rf amplifier (**fig. 7**) and greatly improves strong-signal reception. Refer to the underchassis layout of **fig. 2** for placement of parts:

1. Remove screw and lockwasher nearest front panel used to secure audio transformer T6.
2. Install a two-terminal, lug-type strip on T6 using screw and lockwasher.
3. Disconnect the white-green-blue wire at TS8-1, pull it back through the cabling and reconnect it to terminal 1 of the newly installed lug-type strip. Call this new strip TS11.
4. Connect R213 (2.2 meg) from TS11-2 to TS11-1. Use sleeve resistor leads as necessary.
5. Connect diode CR11 (1N458) from TS11-2 (cathode) to TS11-1 (anode). Use sleeve diode leads as necessary.
6. Connect C276 (0.05 μF) from TS11-1 to E30B.
7. Of the two white-green-blue wires connected to E40-I, disconnect and tape the end of the one showing continuity to TS11-1. You'll have to disconnect both wires to make this check, then resolder the wanted wire.
8. Connect an insulated wire from E40-I to TS11-2, routing it along the cabling. Check wiring against **fig. 7**. Mark the modification in your manual for reference.

Audio distortion on strong signals? Aside from the above modification, another cause exists in some KWM-2s for fuzzy audio. Place a 0.01 μF, 600-volt capacitor from the screen of audio output tube (V16B, pin 8) to ground. Also place a 56-ohm, 1-watt

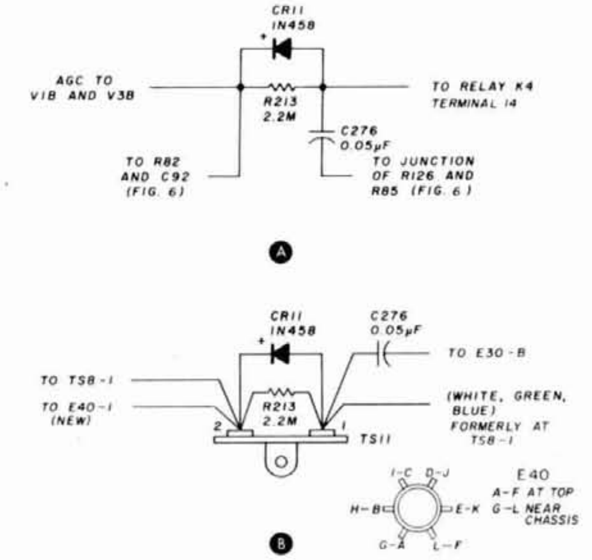

fig. 7. Adding hang agc to your KWM-2, (A). Parts placement is shown in (B). (See also fig. 2.) Terminal support E40 is between socket XV16 and inductor L9. New resistors are ½ watt.

resistor from the yellow (4-ohm) lead of output transformer T6 to ground. These mods will eliminate a weak audio parasitic oscillation sometimes encountered in some receivers.

general modification notes

Modification of the KWM-2 is not recommended for those who have no experience working with small components in cramped spaces. Many KWM-2s are wired with PVC wiring insulation, which melts quickly at the inadvertent touch of a soldering iron. Always check transceiver operation before and after each modification. After your modification, check for wiring errors or shorts and make sure that small specks of solder and wire are blown out of the chassis before power is applied. Also be aware that I've not seen *all* existing KWM-2s and that these mods may not work as shown with some transceiver variations. If you don't understand your present circuit wiring or if it doesn't match the schematics, don't attempt the modification!

where to get help

This material has been prepared with the help of Dennis Brothers, WAØCBK, formerly an engineering technician of KWM-2 production at Collins-Rockwell Company. For those not wishing to make these (and other more sophisticated modifications) themselves, I suggest they contact Dennis at Western Nebraska Electronics, Route 1, Box 1, Potter, Nebraska 69156. A self-addressed, stamped envelope for rapid reply is requested.

ham radio

More Details? CHECK—OFF Page 126

commutating filters

Discussion of the commutating filter — the application of analog and digital techniques to implement a bandpass filter

The world is becoming increasingly digital. In fact, many engineers and electronic technicians are worried about their positions in a technical scenario wherein the linear art is shrinking as digital techniques take over. The real truth, as I see it, is that digital is *not* going to take over at all, but will provide additional techniques creating circuit solutions where the linear techniques they replace are shaky. In fact, in applying these new digital techniques, the linear circuit area will be even further expanded.

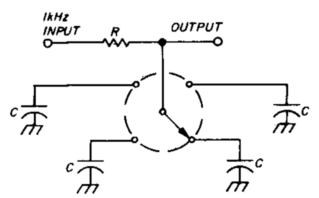

fig. 1. Diagram of a simplified commutating filter. The switch makes one revolution each period of the desired signal frequency to be filtered.

The commutating filter is a good example of how digital techniques provide a simple solution to an analog problem, but which would not work without the addition of some circuitry that is strictly analog. The commutating filter, as presented in **fig. 1**, is designed for 1 kHz; it is a bandpass filter, and its center frequency is dependent only on the frequency with which it is driven. The bandwidth is dependent only on R and C; in fact, this bandwidth is exactly twice the cutoff frequency of the single RC lowpass filter of **fig. 2**. If you look at the voltage on any one capacitor of the filter of **fig. 1**, you'll see a near-dc signal which is the difference-frequency between the drive frequency and the signal frequency. Like the simple RC lowpass of **fig. 2**, it will drop to -3 dB when the signal frequency drops to ½πRC above the drive frequency. It is easily seen why the bandwidth is double the cutoff frequency of a simple RC lowpass filter. Since the output is being commutated sequentially through the four capacitors, it is modulated back up in frequency to that of the input, and phase is preserved.

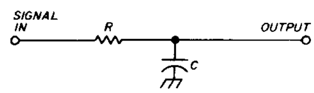

fig. 2. Simple RC lowpass filter from which the commutating filter is derived.

It might sound as if the perfect filter has been

By Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

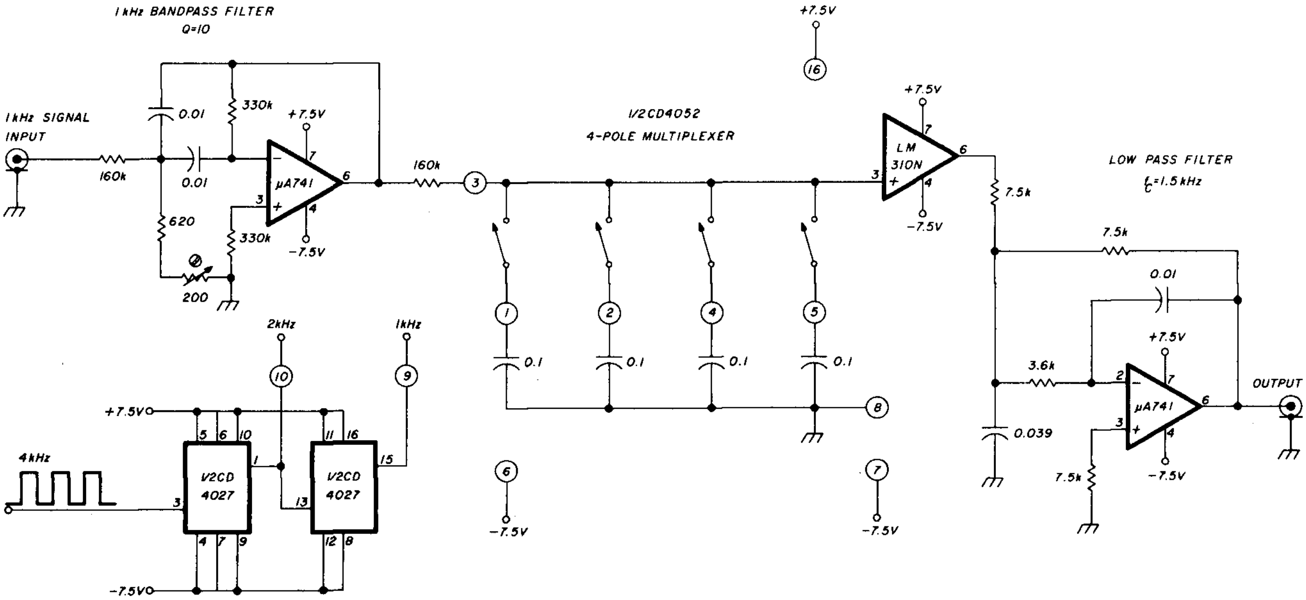

fig. 3. Example of a four-sample-per-cycle commutating filter; circled numbers represent pins of the CD4052 4-pole multiplexer.

achieved with no drawbacks. As usual, there's no free lunch, and you'll find that the commutating filter has some inherent problems. One of these problems is the phenomenon known as "aliasing." Aliasing is the disagreeable habit of filters of this sort to pass not only the same frequency as the drive frequency, but also harmonics of the drive frequency. This can be alleviated by preceding the commutating filter with a simple conventional bandpass filter that attenuates signal frequencies that correspond to harmonics of the drive frequency. This "pre-filter" can, of course, be much broader than the ultimate system bandwidth that our commutating filter provides.

Another drawback of the commutating filter is that

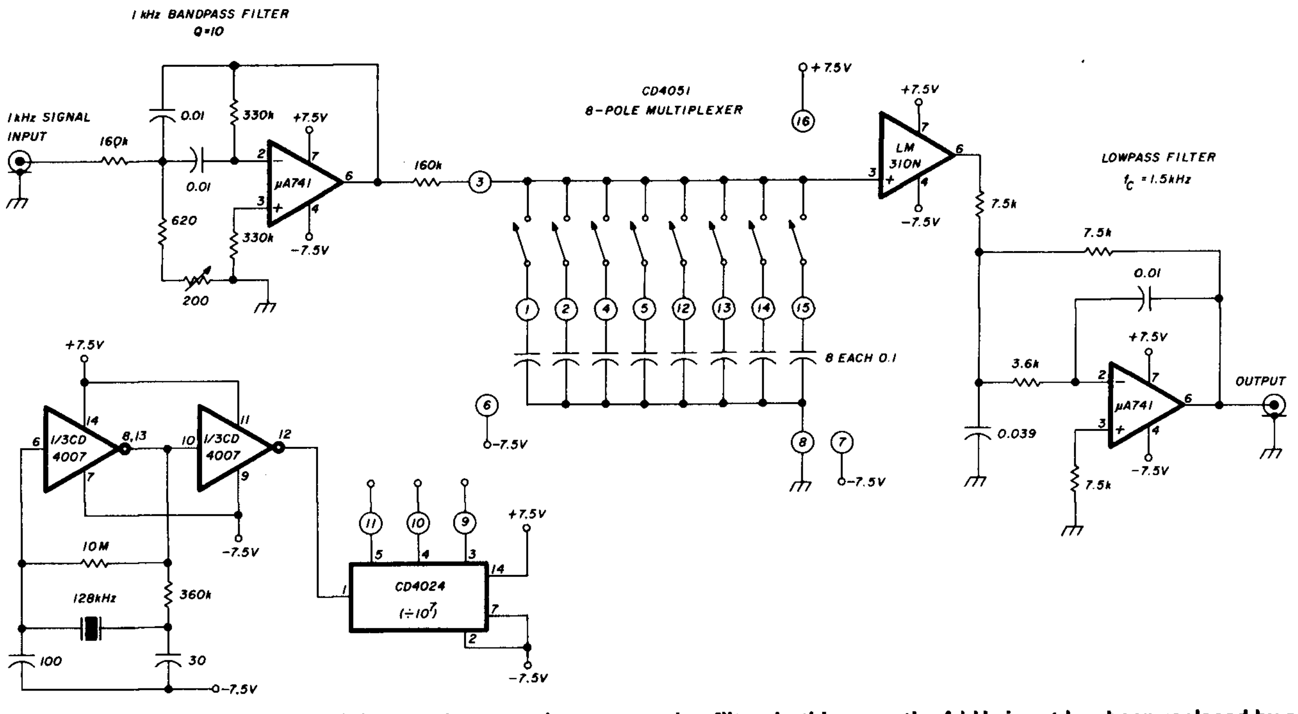

fig. 4. Schematic diagram of an eight-sample-per-cycle commutating filter. In this case, the 4-kHz input has been replaced by a crystal oscillator and a CD4024.

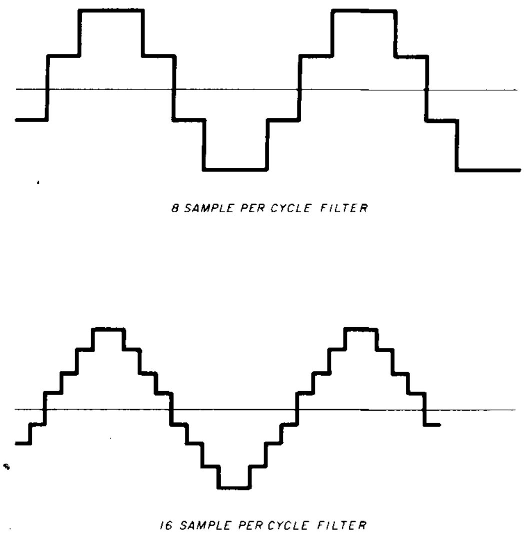

fig. 5. Examples of the idealized waveforms from the 8 and 16 samples per cycle filters. These waveforms are before the post-filter.

it has essentially an infinite output impedance (very much like any simple RC lowpass filter). Practically, however, if the load into which the filter operates is a couple of orders of magnitude higher than R, all is well. The simple solution is to terminate the filter in a high-impedance, noninverting follower.

The last problem of the commutating filter is that it contains harmonics of the drive frequency in its output. These can generally be removed by a lowpass filter, but it is an extra little requirement that must be met if you are to take advantage of the performance of a commutating filter.

As disheartening as all the above restrictions may seem, modern ICs (both linear and digital) come to the rescue to make the commutating filter a fairly simple one. In **fig. 3** is shown a complete four-sample-per-cycle commutating filter using an operational amplifier as a pre-filter bandpass, a CD4052 (CMOS multiplexer/demultiplexer) as the switching (and steering) element, an operational amplifier as a noninverting follower, and an operational amplifier as a lowpass post-filter. Since the CD4052 (half of it) has built-in decoding (steering), it requires 2 kHz and 1 kHz (two-bit) input. These inputs are derived from a CD4027 dual flip-flop wired as a ripple counter and having a 4-kHz input.

The filter of **fig. 3** samples the input signal four times per cycle, and thus the "steps," or discontinuities, in the output (before post-filtering) are relatively large. By going to a filter that takes eight samples per cycle, you decrease the "step" size and ease the post-filter requirements. In **fig. 4**, a CD4051 and eight capacitors replace one half of the CD4052 and the four capacitors. This multiplexer is another member of the same CMOS family as the CD4052, but it requires three-bit drive: 4 kHz, 2 kHz, and 1 kHz. To accomplish this drive requirement, another CD4027 flip-flop could be used with an 8-kHz input. Or you could use a single CD4024, which is a seven-stage ripple counter (divide-by-128), and use any convenient three adjacent outputs for the drive. In **fig. 4**, the last three outputs (pins 5, 4, and 3) of the CD4024 are used to drive the CD4051, thus requiring a 128-kHz input to pin 1. This 128 kHz is provided by a 128-kHz crystal oscillator made from two-thirds of a CD4007.

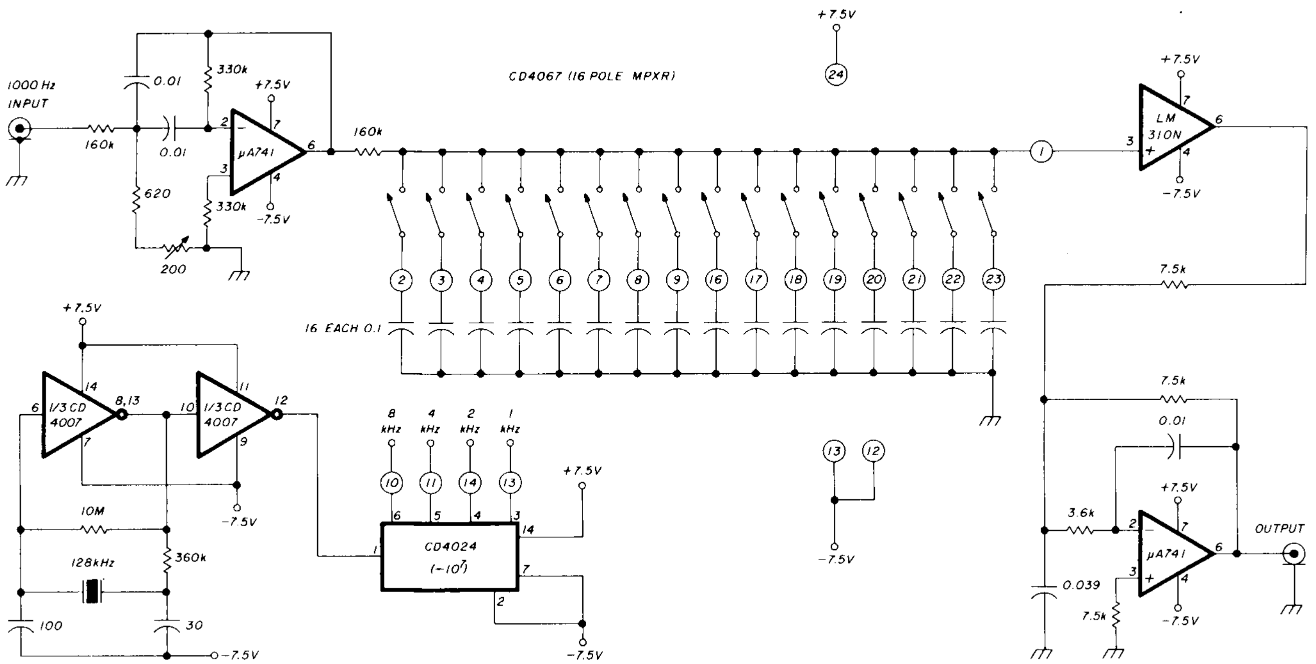

fig. 6. Schematic diagram of a 16-sample-per-cycle commutating filter.

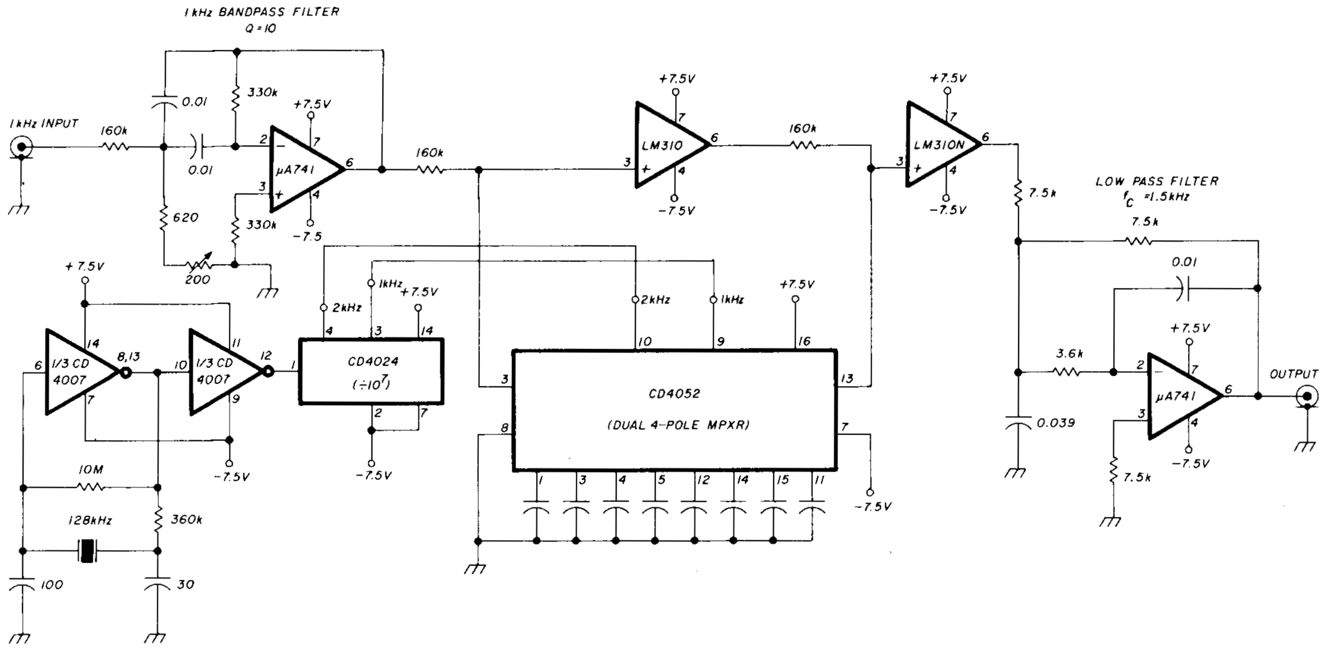

fig. 7. In this case, two stages of four-sample-per-cycle filters have been cascaded. Only one set of pre- and post-filters are necessary, with an impedance follower between the sections to lower the driving impedance to the second section.

Fig. 5 shows the unfiltered outputs of the eight-sample and the sixteen-sample filters. Note how the eight-sample filter has more (and smaller) "steps" in its output, and is thus easier to post-filter. It is even fairly simple to expand the filters of figs. 3 and 4 to sixteen samples per cycle, which really cuts down the quantization ripple in the output. Such a circuit is shown in fig. 6, using a CD4067 and sixteen capacitors.

It is even possible to cascade commutating filters, and the pre-filter and post-filter need not be replicated. A follower between sections is all that is required for lowering the driving impedance to the second section. An example of a two-section, four-sample-

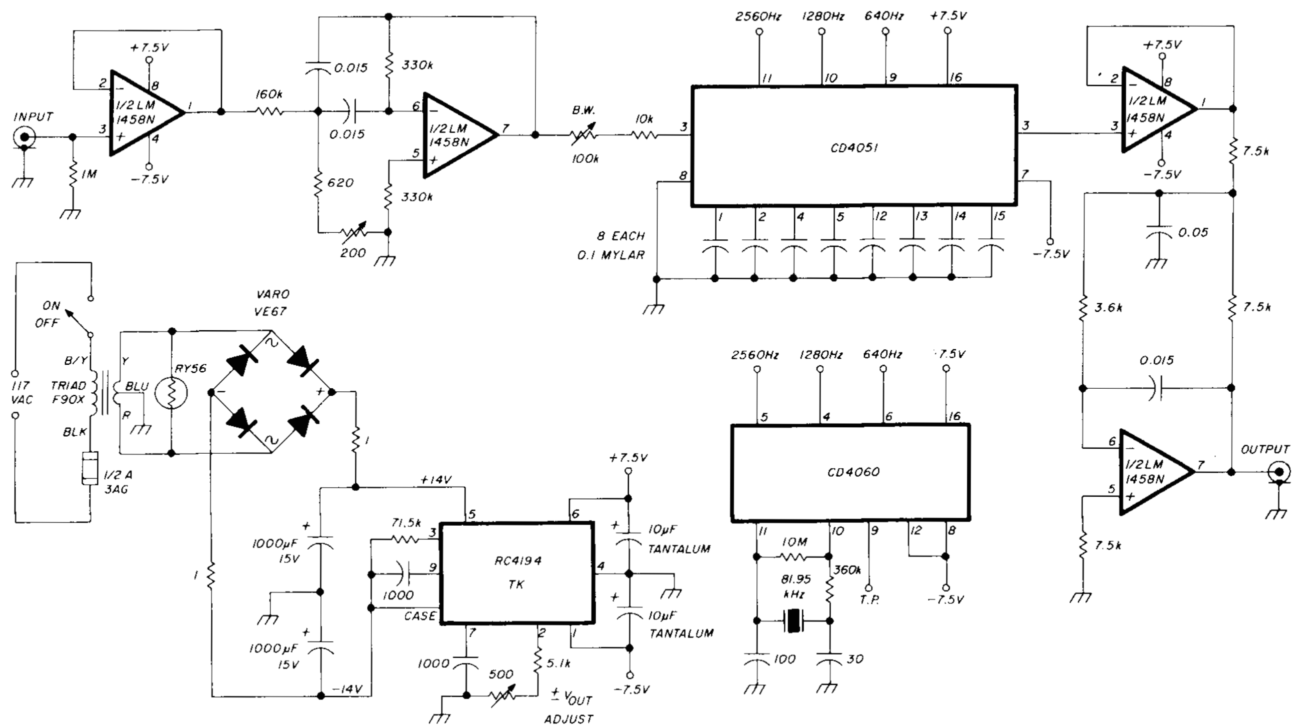

fig. 8. Example of a practical commutating filter. This filter has been set up for an operating frequency of 640 Hz.

per-cycle filter is shown in **fig. 7**. The advantage of cascading is the sharpness of rolloff outside the passband. The rate of rolloff of the equivalent RC lowpass is then 12 dB/octave instead of 6 dB/octave.

Finally, a construction project using a commutating filter in a useful piece of ham equipment is presented in **fig. 8**. The design process was as follows:

1. The operator tuned in a CW signal off-the-air and adjusted the BFO until the tone was of an agreeable pitch. This pitch was measured using an oscilloscope to find out what frequency the operator likes to copy. This subjective determination of the operator pitch preference may seem like wasted motion, but many people have "holes" in their hearing response (especially older CW operators).

2. Once the desired pitch frequency is determined, multiply it by 128 and get the oscillator frequency to the CD4060. As an example, say that the operator preference turned out to be 640 Hz; then, the input frequency would be 81.92 kHz. If one has a crystal of about that frequency, it can be used directly in **fig. 8**. Otherwise, higher frequency crystals could be used, with taps at positions further down the divide-by-two chain. For instance, a 328-kHz crystal could be used, the outputs taken from pins 6, 14, and 13 (still yielding 2560, 1280, and 640 Hz respectively).

3. The pre-filter center frequency is then adjusted to the chosen operating frequency, in this case 640 Hz. The C values scale with frequency so that C is 0.01 μF for 1 kHz and 0.015 μF for 640 Hz.

4. The post-filter cutoff frequency is adjusted to be 1.5 times the bandpass filter center frequency. Again, the capacitor values scale with frequency. Capacitor values of 0.01 and 0.033 μF give a 1.5-kHz cutoff frequency, and 0.015 μF and 0.05 μF give a 960-Hz cutoff frequency.

In **fig. 8**, an input noninverting follower has been added so that the unit may be driven from almost any impedance. An LM1458 dual op amp is used for both the input follower and the pre-filter. Addition of a variable resistance in place of R allows the passband to be varied from 3.0 Hz to 30.0 Hz, continuously. Another LM1458 dual op amp is used in the output section as the noninverting follower and lowpass filter. By using two dual op amps, and using the CD4060 (which combines the crystal oscillator and divider in one IC package) I've reduced the circuit down to four ICs, plus the one IC used as the power supply regulator. The 81.92-kHz crystal was actually an 81.95-kHz unit that is quite common on the surplus market, 81.95 kHz being a standard time base frequency for a variety of distance-measuring devices.

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accu-keyer speed readout

Another addition to the feature-packed WB4VVF Accu-Keyer — a readout system for code speed

There are thousands of Accu-Keyers[1] already in use, and the appearance of articles[2-8] to add message memories to the basic keyer has undoubtedly resulted in another flurry of Accu-Keyer construction. It is an excellent and highly versatile keyer, and deserves the fine reputation that it has. It might seem that there is little else that one could want from this, or any other, keyer.

There is one useful addition, however. Most of us vaguely know our sending speed. It is true that a speed scale could be put on the front panel behind the control, but the speed vs rotation dependence of most controls is highly nonlinear, especially at the high-speed end of the range where the scale becomes compressed. Any semblance of accuracy is lost in the compressed scale.

A desirable feature, which I have incorporated into the Accu-Keyer system, is a direct words-per-minute speed readout. This is useful for many purposes, and at the least is an interesting conversation piece in the hamshack.

The readout and keyer clock, which I will describe, may be easily used in any Accu-Keyer design, and possibly in other types of keyers as well. The main precaution to be observed with the Accu-Keyer family is to be sure the 5-volt power supply in your keyer is capable of handling the extra current drain, about 370 mA.

I do not consider it feasible to use my readout with a battery-operated keyer,[9,10] but it should be possible to make relatively simple modifications to the circuit and use CMOS integrated circuits. It would be necessary to choose some other type of display, and I would recommend a liquid-crystal type.

principles of operation

A continuous speed readout in wpm *requires* a free-running clock. The Accu-Keyer clock, however, is not free running. It starts when either side of the paddle is closed, and is stopped by an inhibit signal from the logic when all characters have been completed. This method has a considerable advantage over a free-running clock, since the operator initiates a character at the time he chooses rather than at the time the clock is finally ready.

This dilemma is easily overcome, and the unit I have developed gives an accurate, continuous readout of the speed without sacrificing the advantages of the operator-started clock. A fringe benefit of the unit is that it does not have the problem, common to some keyer clocks, of a first clock pulse different in duration from the rest of the pulses in the sequence. Because of these features, it may be worthwhile to use the clock portion of this unit, even without the readout.

The speed is variable from five to around fifty wpm, an adequate range for almost anyone from Novice to Extra. The speed display is updated approximately six times each second, whether or not any sending is being done. I incorporated it into the WA9LUD memory version[2] of the Accu-Keyer, but of course it can be used with any similar keyer

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design. Different speeds may be selected as you build.

Recent editions of the ARRL *Radio Amateur's Handbook* give the relationship between code speed and keyer clock frequency as:

$$speed\ (wpm) = 1.2 \times f\ (clock\ frequency)$$

Twenty pulses per second of the clock result in a keying speed of 24 wpm. A scheme for reading out this relationship has been described previously,[11] but that system has several disadvantages which are overcome by my circuit.

Suppose you have a high-frequency pulse generator running at 2420 pulses per second. Three decade counters hooked in series would count to 242 if they are allowed to count for exactly 0.1 second. If the least-significant digit (2 in this example) is ignored, it is then possible to display 24 in the readout connected to the digital counters. The reason for this approach will be discussed more fully later.

The high-frequency pulse generator can also be divided down by a decade and a duodecimal (divide by twelve) divider, a total division of 120, to give twenty pulses per second for the keyer clock. If you gate the divide-by-120 divider on and off with the original inhibit line in the Accu-Keyer, the resulting keyer clock line acts much like the operator-started clock, which is the key to the success of the Accu-Keyer design. This scheme allows us to have a free-running clock that can be accessed at the operator's

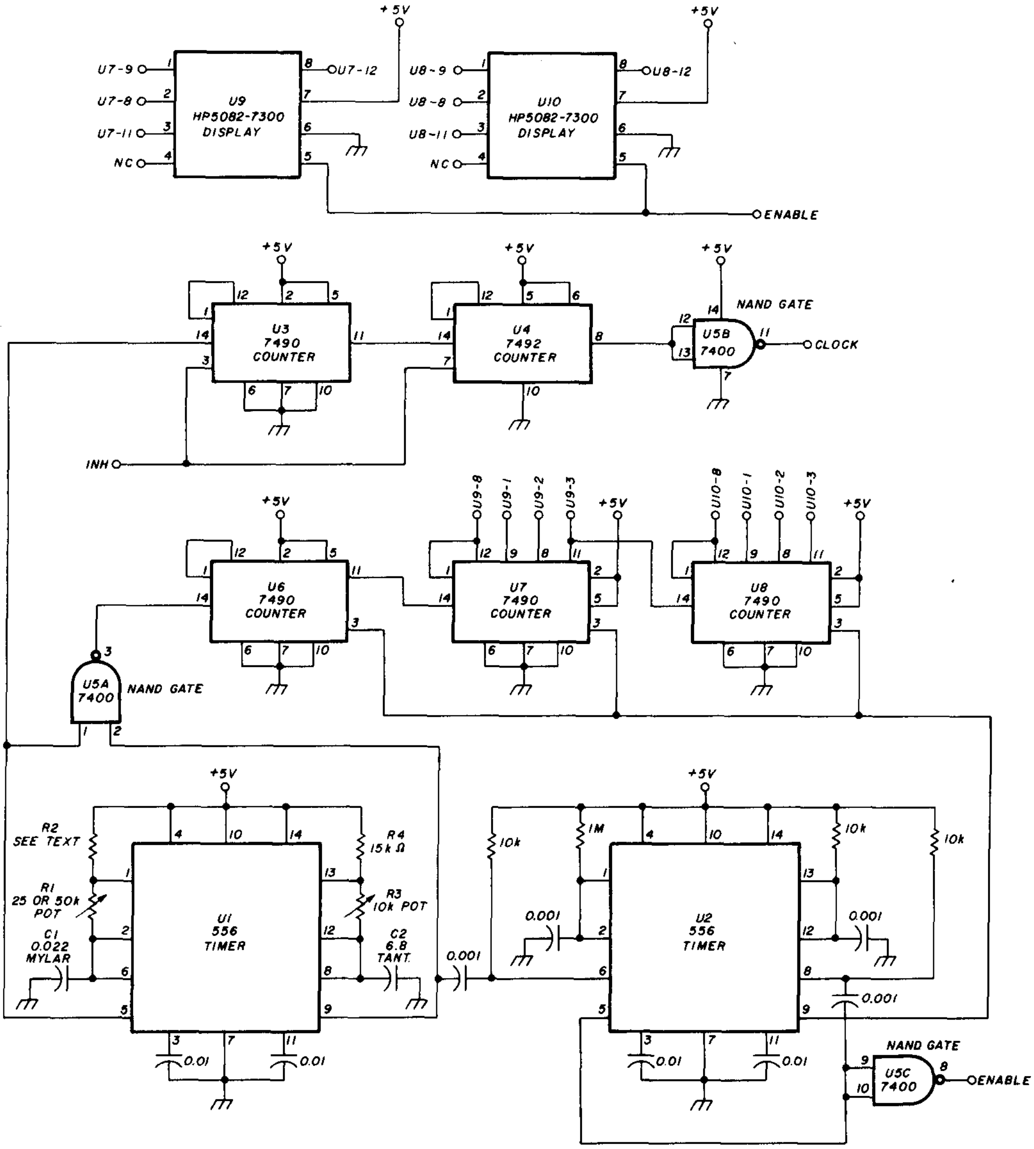


fig. 1. Schematic diagram for the Accu-Keyer speed readout. This circuit incorporates a free-running clock which can be accessed at will by the operator. The frequency of the clock is high enough that the delay between accessing and the first clock pulse is negligible. U9 and U10 are HP 5082-7300 displays that have the latches and display drivers incorporated within the display. C1 and C2 should be of the type indicated to ensure adequate stability.

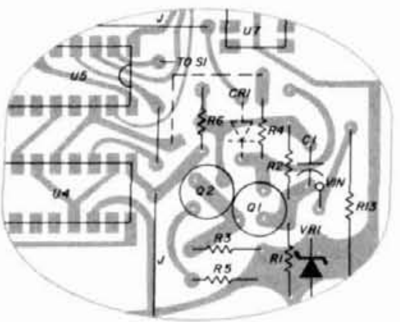


fig. 2. Blowup of the portion of the keyer board which is changed to incorporate the speed readout. CR1 and the original speed control wires must be removed. The foil is cut and new wires attached at the indicated spots.

convenience. I have never been able to detect any delay because of the free-running pulse generator, even at the slowest keying speed.

circuit description

The logic diagram for the clock/readout is given in **fig. 1**. U1 and U2 are 556 dual timers. One half of U1 generates the high-frequency pulses, available from pin 5, that form the basis of the clock/readout. The other half of U1 is the time base for the display counter, with the output on pin 9.

R1 is the speed control and is mounted on the front panel of the keyer. C1 must be a reasonably stable capacitor, *not* one of the ceramic bypass types. C1 and R2 determine the maximum keying speed, and the value of R1 determines the range. The value of R2 will probably be between 6,000 and 22,000 ohms for a 50 wpm maximum, and may be selected for this purpose. If C1 is changed for any reason at some later time, it may be necessary to change R2 to bring the maximum speed back to the one desired.

R3 is mounted on the printed circuit board and is used to adjust the 100-ms time base for the display counter. If it is not possible to adjust the "on" time at pin 9 of U1 to 100 ms, it may be necessary to change the value of R4 to bring the pot within the proper range. C2 is the most critical component in this entire circuit.

U2 is simply a sequential timer. The trailing edge of the 100-ms counter gate triggers a pulse of short duration at pin 5 of U2. This pulse, after inversion by U5C, strobes the count in the decade counters into the display. It also triggers another short pulse, at pin 9 of U2, which is used to reset the counters to zero, preparing them for the next update.

U3, a 7490 decade counter, and U4, a 7492 duodecimal counter, form the divide-by-120 divider that generates the clock pulses for the keyer logic. This divider is gated on and off by the inhibit line from the keyer, with the inhibit signal resetting the divider to zero and holding it there when all keyer action is complete. Inverter U5B ensures that the clock pulses have the right polarity for the Accu-Keyer, and might not be necessary in other keyer designs. This combination forms a keyer clock which is always within 1/120th of a dit of starting, a negligible delay at any speed.

U5A controls the display counting. The pulse generator pulses are fed to the counter only when pin 9 of U1 is high. When it is high for precisely 100 ms, exactly one tenth of the pulse generator frequency is counted. U6, another 7490 counter, is for the least significant digit and, by including it without display, the jitter inherent in this digit is eliminated. This results in a stable display considerably superior to using only two decade counters with a 10-ms time base. U7 and U8, both 7490 counters, are the actual display counters, with U8 serving as the most-significant-digit counter.

The displays themselves, U9 and U10, are easy to use, with an attractive, bright display, although they are a bit expensive. Other displays may be substituted, but it might be necessary to incorporate data-storing latches, which are built into the 5082-7300 displays. A nonblinking display is a necessity, so be sure to add latches if they are not in the displays you choose.

Connection to the Accu-Keyer is really quite simple. CR1 in the original Accu-Keyer clock *must* be

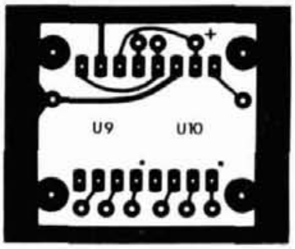
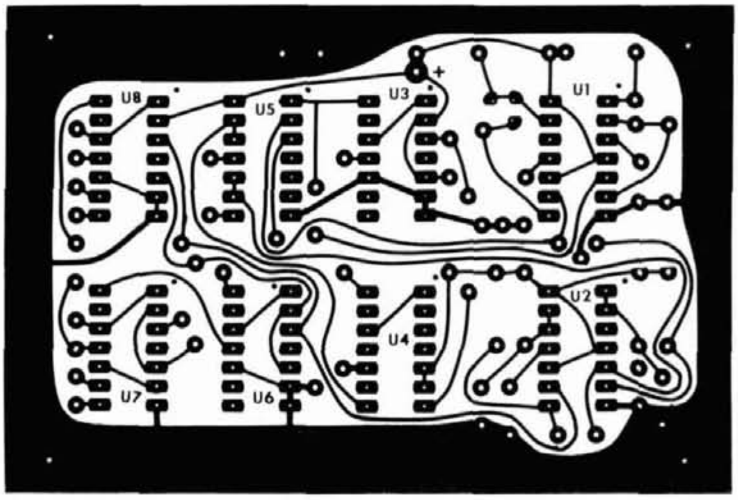


fig. 3. Full-size printed-circuit layout for the Accu-Keyer speed readout. Parts layout is shown in fig. 4.



removed from the circuit. A wire is connected to the vacated hole at the anode end for connection to the inhibit line in the new clock. The foil should be cut as indicated in **fig. 2** and the old speed control wires should be removed. The clock line may then be connected to the vacant hole near the cut in the foil. Connect V_{cc} and ground both the readout and the clock board, and you're in business. You may wish to remove the old clock components from the Accu-Keyer board, but that is not really necessary.

I have not included a power supply, since most will be able to use the supply in the Accu-Keyer. It might be necessary to increase the size of the input capacitor ahead of the regulator to keep the voltage high enough to maintain regulation. If your supply is incapable of providing the necessary current, any standard 5-volt power supply design will be satisfactory.

Full-size board layouts and the component placement diagram are shown in **figs. 3** and **4**. They are single-sided boards, and should be easy to duplicate by those who wish to roll their own. There is no reason why point-to-point wiring cannot be used, since the layout is not critical.

accuracy and calibration

The key to the accuracy of this unit is how carefully the 100-ms time base for the display counter is calibrated, and how stable it is. It would have been possible, of course, to use a crystal-controlled clock to control this counter, but that seemed quite unnecessary. One half of a 556 timer, with a high-quality, stable capacitor, results in quite adequate performance for this purpose. It saves considerably on circuit complication and expense.

There are three methods of calibration, and they will be described in order of increasing accuracy.

1. Set the keyer to match as closely as possible W1AW's 18-wpm bulletin broadcasts (or better yet their 35-wpm code practice), and adjust R3 until the readout indicates 18 (35).

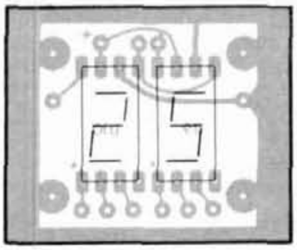
2. Use a calibrated scope to set the "on" time (output high), as seen at pin 9 of U1, while adjusting R3.

3. Connect a counter with a 1-second time base to U1, pin 5, to measure the pulse generator frequency, and adjust the keyer speed control until the counter reads about 4000. Adjust R3 until the display reads 40. This is the method I prefer, and should be used if a counter is available.

My own keyer has been in use for almost three years and seems to be accurate within one wpm at all speeds throughout its range at all temperatures encountered so far in my shack. Accuracy is not a problem if a sufficiently stable capacitor is used for C2.

I'll be happy to answer any correspondence regarding this readout or any modifications people may wish to make. I'll try to furnish circuit board availability information, provided that a self-addressed, stamped envelope is supplied.

It has been a pleasure to use this keyer with its readout. Now, when someone says QRQ by 5 wpm, I can do it quite accurately, depending on my skill of sending, of course!



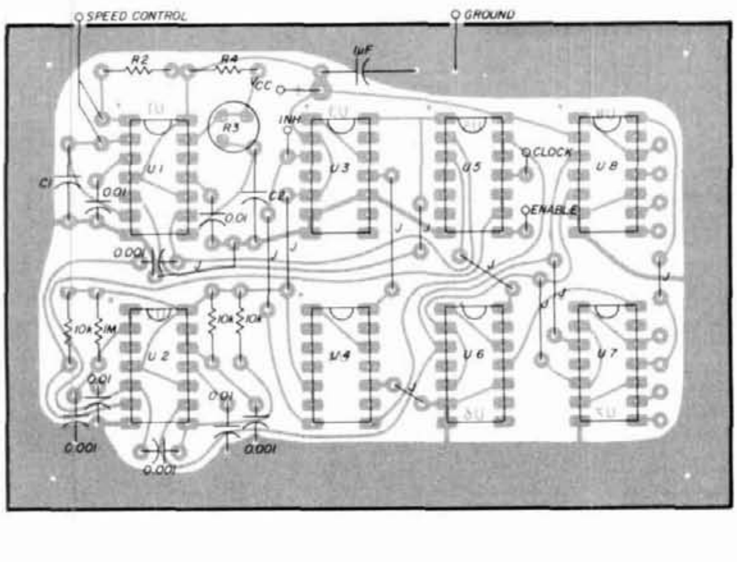


fig. 4. Component placement for the Accu-Keyer speed readout.

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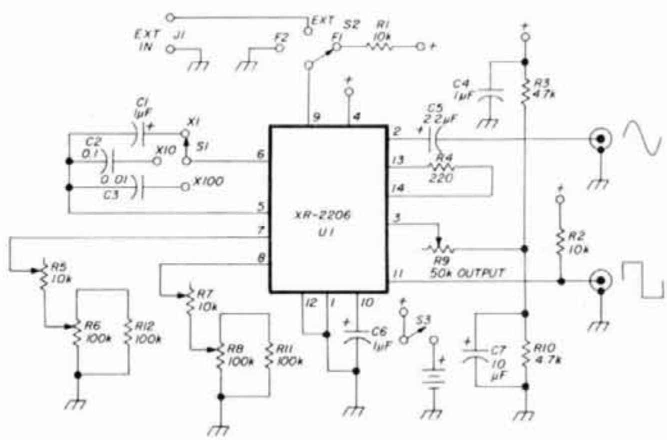
The duplex audio generator is built into a Mod-U-Box available from Quement Electronics (see text). Controls are f1 (upper left), f2 (upper right), and output level (center).

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



B1	battery, 12.6 V, Mallory 304116 Smoke Detector Battery
C1, C4, C6	1 μf 35 vdc*
C2	.1 μf 35 vdc*
C3	.01 μf 35 vdc*
C5	2.2 μf 35 vdc*
C7	10 μf 16 vdc*
J1, J2, J3	jacks or binding posts of your choice
R1, R2	10K ¼ w*
R3, R10	4.7K ¼ w*
R4	220 ¼ w*
R5, R7	10k trimmer, Radio Shack 271-218*
R6, R8	100k pot, audio taper, Radio Shack 271-1722
R9	50k pot, linear taper, Radio Shack 271-1716
R11, R12	100k ¼ w*
S1, S2	switch, single-pole, three-position
S3	switch, mounts on R9, Radio Shack 271-1740
U1	XR2206, James Electronics XR2206*
Case	Mod-U-Box 3-7-6, Quement Electronics
PC Board	J. Oswald 1006J*

PC board and board-mounted components kit available from J. Oswald, part 1006K (includes parts marked*)

J. Oswald, 1436 Gerhartd Avenue, San Jose, California 95125.
1006J $4.75 PPD. 1006K $17.75 PPD.

James Electronics, 1021 Howard Street, San Carlos, California 94070.

California residents add sales tax.

fig. 1. Schematic of the duplex audio-frequency generator. Design is built around the James Electronics XR-2206 function-generator IC and includes an AFSK signal.

and, because component count is small, cost of the project is minimal.

description

The generator is built around the XR-2206 monolithic function generator IC. This little IC can perform many functions, and in this particular application we're using only a couple of its many features. As seen from the schematic, **fig. 1**, the XR-2206 and a handful of passive components form the entire generator.

The audio-frequency spectrum is covered in three ranges; 20-200 Hz, 200-2000 Hz, and 2 kHz-20 kHz, as selected by the range switch, S1. The range switch is labeled X1, X10, and X100, allowing the use of a single scale on the frequency dials. The specific frequencies desired within these three ranges are selected by the frequency controls, F1, (R6), and F2 (R8). Switch S2 selects the generator output frequency as F1, F2, or EXT. In the F1 position R6 determines the output frequency, while the F2 position allows R8 to control the output frequency. With switch S2 in the EXT position, the output frequency is selected by the signal applied to the external input jacks. Frequency F1 is selected by a high level or open contact, and F2 is selected by a low level, or closed contact. In this manner an AFSK signal of adjustable frequency, shift, and amplitude is generated.

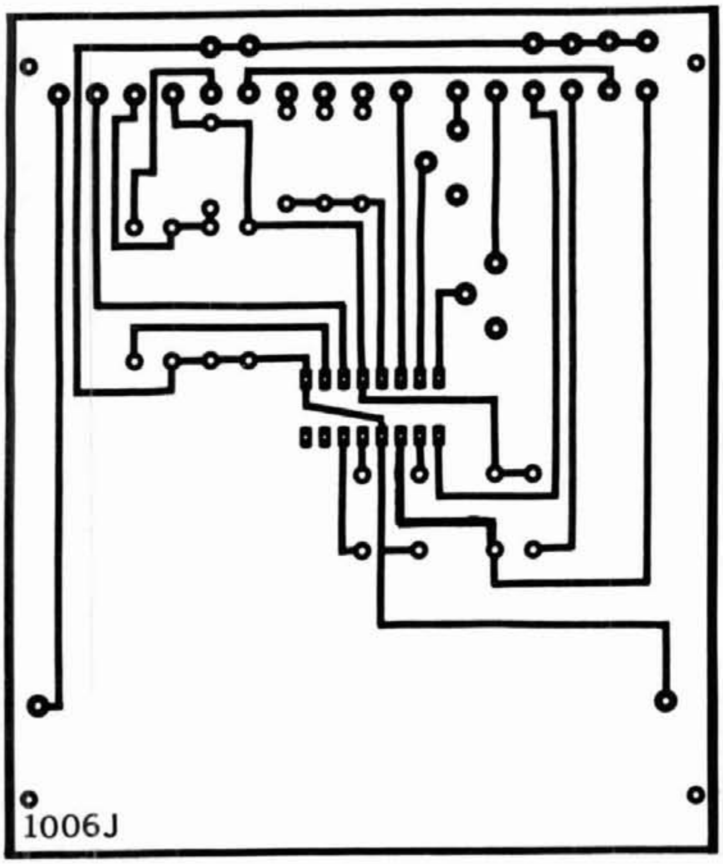
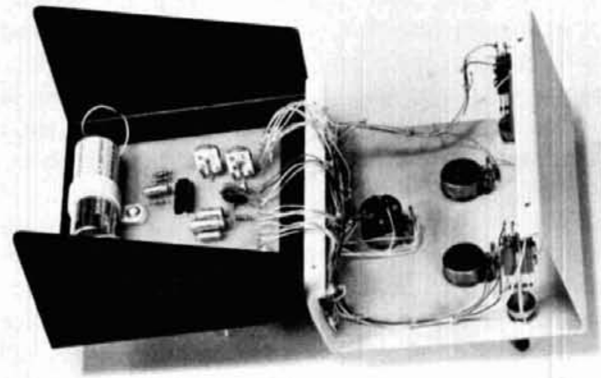


fig. 2. Full-size layout of the PC board (foil side).



Construction of the duplex audio generator showing the two switches and three variable controls on the removable front panel (right), and the printed-circuit board in the base (left). Power is provided by a 12.6-volt battery designed for smoke detectors.

The sine-wave output of the generator is variable to a maximum of 3 volts peak-to-peak through the output level control, R9. The square-wave output is fixed at a TTL level; and because both outputs are available simultaneously, the square-wave output provides a very handy sync point for scope triggering.

The generator is powered by a 12.6-volt battery. Current consumption is low, so extended battery life can be expected. The basic circuit is not overly critical to voltage changes, and the first indication of battery failure will be flat topping of the sine-wave output at high-amplitude levels. Trimmer resistors, R5 and R7, are used to calibrate the frequency controls, and capacitors C1, C2, and C3 determine the range-multiplier accuracy. Generator output impedance is a nominal 600 ohms and provides a good match to most standard audio equipment.

construction

Virtually any type of construction practice could be used for the audio generator, because the circuit isn't critical the way rf circuits are. PC board construction was chosen for ease of assembly and predictable results. A full-size layout of the foil side of the PC board is shown in **fig. 2**. The component layout as viewed from the top, or component, side of the board is illustrated in **fig. 3**.

A practical approach to construction is to mount and solder all board-mounted components and then add the interconnecting wires to the front edge of the board. Leave these wires about 30 cm (12 inches) long for connection to the front panel after doing the sheet metal work. Drill and deburr all holes for the front panel controls and jacks, as well as the PC-board mounting holes in the case. Next, mount all front-panel components and the PC board, using small standoff spacers to elevate the PC board above the case.

Place the front panel next to the case and wire the interconnecting leads from the PC board to the panel controls, jacks, and switches. As the case goes together, the interconnecting wires should fold over neatly. After you're sure that the wiring doesn't interfere with the case assembly, the wires can be spot-tied to form neat groups and retain their positions. Assemble the case again to insure that everything fits well.

test and calibration

A very simple test of the generator can be made with the aid of a pair of headphones. Set trimmers R5 and R7 to midpoint, connect the phones to the sine-wave output, set S2 to the F1 position, range switch to X1, the frequency controls fully counterclockwise. Advance OUTPUT LEVEL control, R9, until an audio tone is heard in the phones. The tone should be about 250 Hz. Flip the function switch to the F2 position and adjust trimmer R7 until you can flip from F1 to F2 without detecting a change in the tone. Change the range switch to the X10 position and rotate the frequency controls fully clockwise. Again this should yield a tone about 200 Hz.

Flip the range switch to the X100 position. The tone should be approximately 2 kHz. Switch back down to the X10 range, listen to the 200-Hz signal for a few seconds, then move the phones to the square-wave output jack, J3. The signal should be a bit raspy because of the square waveform. Now set the function switch to the EXT position and the F2 control to its midpoint. Short the external input jacks with a jumper, and the 200-Hz signal should shift to approximately 750 Hz, indicating that the AFSK circuitry is functioning. This evaluates all the functions of the audio generator and should make you feel pretty good.

Calibration of the little generator is subject to the equipment you might have available, such as a scope or frequency counter, and also to the degree of accuracy you're trying to attain. The generator is a small package, so there really wasn't much room for large dials that would provide high resolution. With the small dials, accuracy is adequate for general audio work, and if you are going to do anything critical you'd probably use a counter with the generator.

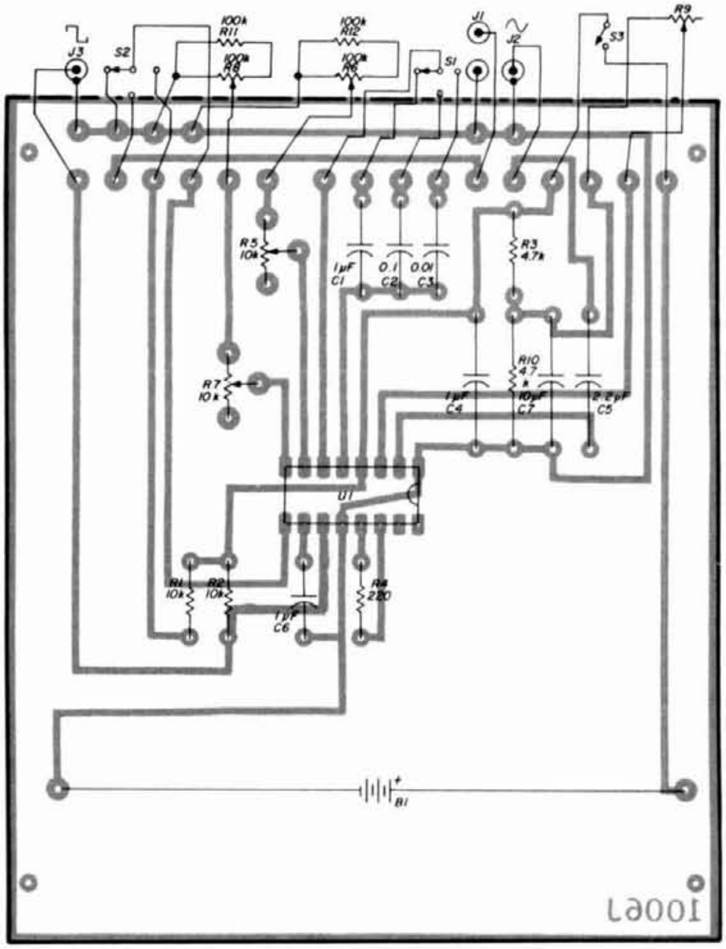


fig. 3. PC-board component layout viewed from the top, or component, side.

The dual outputs make this a very easy thing to do. I used a scope to measure the output pulsewidth and put small pencil marks around the dial area of the frequency controls. Then I removed the controls, switches, and jacks from the front panel and applied the lettering using Datak rub-ons, followed by a light coat of clear Krylon to protect the lettering. This gives the instrument a professional look and provides adequate calibration. Use care in putting the front panel components back in place so that the lettering isn't damaged.

The duplex audio-frequency generator is a very flexible piece of test equipment that can be built at a low cost and is worthy of a place in every experimenter's shop. It can be built by a beginner and will remain useful to him as his interests change to the more complex phases of Amateur Radio. Also, the basic circuitry can be lifted for many other applications, limited only by the creative ability and interests of amateurs and experimenters.

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The 100 HH and 500 HH hand held frequency counters represent a significant new advancement, utilizing the latest LSI design . . . and because it's a DSI innovation, you know it obsoletes any competitive makes, both in price and performance. No longer do you have to sacrifice accuracy, ultra small readouts and poor resolution to get a calculator size instrument. Both the 100 HH and 500 HH have eight .4 inch LED digits — 1 Hz resolution — direct in only 1 sec. or 10 Hz in .1 sec. — 1 PPM TCXO time base. These counters are perfect for all applications be it mobile, hilltop, marine or bench work. **CALL TODAY TOLL FREE:** (800—854-2049) **Cal. Res. CALL** (800—542-6253) **TO ORDER OR RECEIVE MORE INFORMATION ON DSI'S FULL PRODUCT LINE OF FREQUENCY COUNTERS RANGING FROM 10 Hz TO 1.3 GHz.**

FREQUENCY COUNTER CONSUMER DATA COMPARISON CHART

MANUFACTURER	MODEL	SUG'STD. LIST PRICE	FREQUENCY RANGE	TYPE OF TIME BASE	ACCURACY OVER TEMPERATURE		SENSITIVITY			DIGITS		PRE-SCALE INPUT RESOLUTION	
					17° - 40°C	0° - 40°C	100 Hz - 25 MHz	50 MHz - 250 MHz	250 MHz - 450 MHz	No.	SIZE IN INCHES	.1 SEC	1 SEC
DSI INSTRUMENTS	100 HH	$ 99.95	50Hz-100MHz	TCXO	1 PPM	2 PPM	25 MV	NA	NA	8	.4	100 Hz	10 Hz
DSI INSTRUMENTS	500 HH	$149.95	50Hz-550MHz	TCXO	1 PPM	2 PPM	25 MV	20 MV	30 MV	8	.4	100 Hz	10 Hz
CSC‡	MAX-550	$149.95	1kHz-550MHz	Non-Compensated	3 PPM @ 25°C	8 PPM	500 MV*	250 MV	250 MV	6	.1	NA	1 kHz
OPTOELECTRONICS	OPT-7000	$139.95	10Hz-600MHz	TCXO	1.8 PPM	3.2 PPM	NS	NS	NS	7	.4	1 kHz	100 Hz

* 1 KHz - 50 MHz ‡ Continental Specialties Corp.

The specifications and prices included in the above chart are as published in manufacturer's literature and advertisements appearing in early 1979. DSI INSTRUMENTS only assumes responsibility for their own specifications.

100 HH . . . $ 99.95 W/Battery Pack . . . $119.95
500 HH . . . $149.95 W/Battery Pack . . . $169.95

Prices and/or specifications subject to change without notice or obligation.

These prices include factory installed rechargeable NiCad battery packs.

DSI INSTRUMENTS, INC.
7924 Ronson Road, Dept. G
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T-500 Ant. $ 7.95
AC-9 Battery Eliminator . . . . $ 7.95

TERMS: MC - VISA - AE - Check - M.O. - COD in U.S. Funds. Please add 10% to a maximum of $10.00 for shipping, handling and insurance. Orders outside of USA & Canada, please add $20.00 addition to cover air shipment. California residents add 6% Sales Tax.

DIGITAL TECHNIQUES

down counters

Most counters can be considered *up* counters. *Their binary states usually increase in bit weight.* This is sometimes a problem when applying them to a phase-locked loop as the programmable divider. In such an application, the divider is first preset to a particular binary state by the front panel control. Counting then proceeds to an all-ones or all-zeroes state. This state outputs a pulse to the phase detector and presets the counters again. Division ratio is the difference between preset state and end-of-count state. It *can be a problem with up counters.*

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

Suppose you want to divide by 888. An up counter must be preset to the nine's complement of each decimal digit (nine minus the desired digit), or decimal 111 in this case. The up counter will then increase through 888 states until an all-ones condition is reached for end-of-count. Confusion arises because the decimal preset is in reverse of the desired decimal division.

A solution is to use a down counter, one whose states decrease with the number of input clocks. Preset and division are now the same number. Motorola makes such a device with the designation MC4016 *and it is designed for PLL applications.**

a BCD down counter

The counter portion of the MC4016 is shown in **fig. 1** with waveforms. D flip-flops are used in place of the usual JKs, and all gates are ANDs. G3 is an open-collector AND to sense all-zeroes from the $\overline{Q}$ outputs.

*The designation was formerly changed to MC74416 but is back to the original number. MC4316 is the military temperature version.

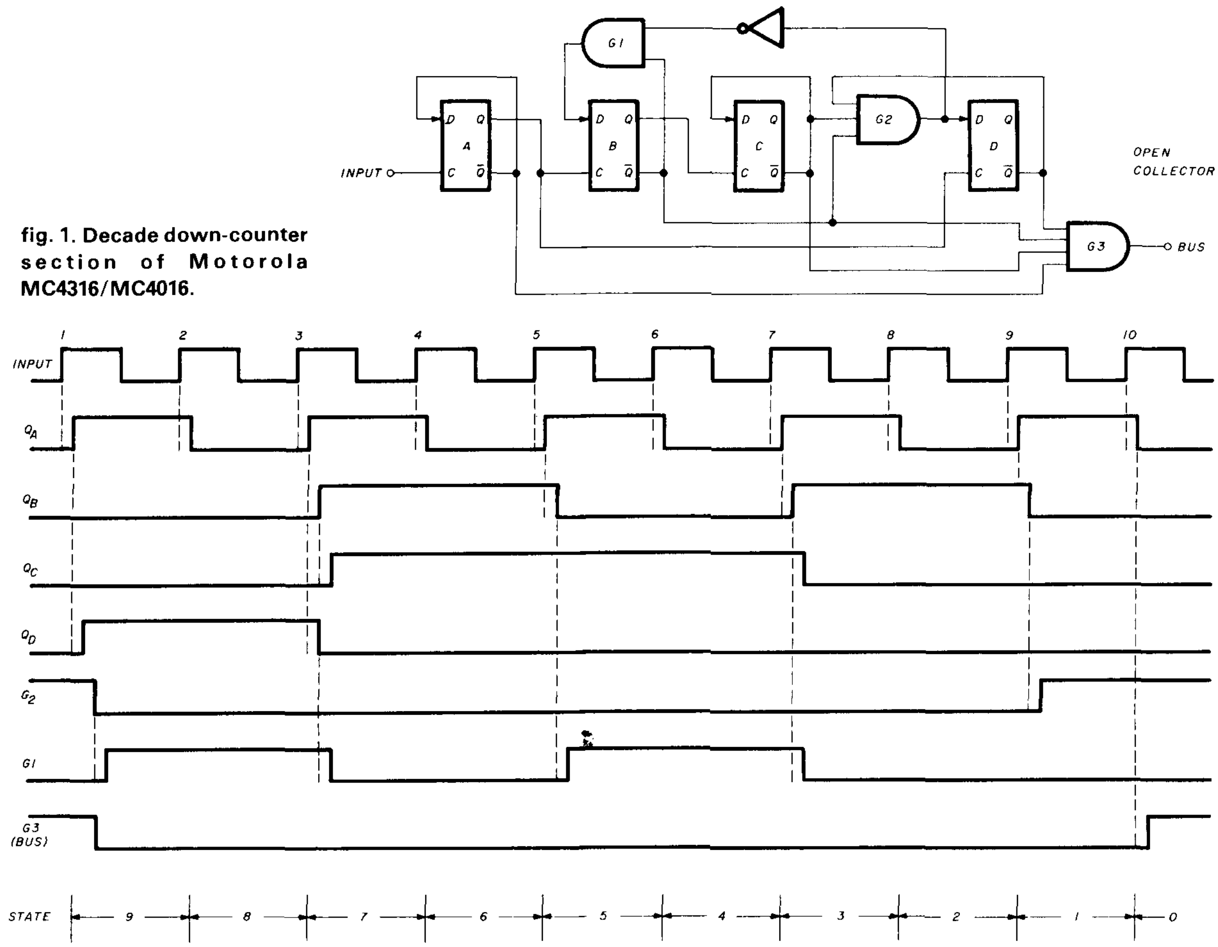
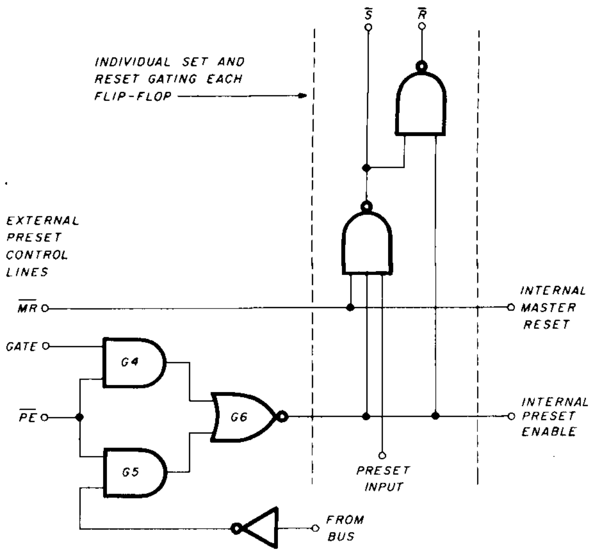


fig. 1. Decade down-counter section of Motorola MC4316/MC4016.



$\overline{MR}$	GATE	$\overline{PE}$	BUS	$\overline{S}$	$\overline{R}$
0	X	0	X	1	0
0	0	X	1	1	0
1	0	0	X	$\overline{P}$	P
1	0	1	1	$\overline{P}$	P
1	1	0	X	$\overline{P}$	P

TRUTH TABLE

fig. 2. Preset control section of the Motorola MC4316/MC4016.

Decreasing BCD states can be seen from the waveforms and the effect of gate output states. As in its up-count version, it's a divide-by-two in cascade with a divide-by-five. Carry out is from Q_D to the next input. A chain of three will go from an initial decimal 000 to a decimal 999 on the first input. Subsequent inputs will change decimal states to 998, 997, 996.

preset control

Each stage has direct set and reset inputs active low. Counting will be overridden when either is low. Two internal buses and three gates provide versatility in preset control. **Fig. 2** shows the preset control section with a truth table for external control inputs. P indicates the state of the external preset input for each stage. An X is a don't-care state; it may be 1 or 0 without changing a particular state combination.

BUS gate 3 was stated as being open-collector. AND gates with open collectors may be wired-AND just as NAND gates may be wired-OR.[1] The internal connection to the inverter doesn't change the open-collector condition. An internal, separate, pull-up resistor is provided on each package.

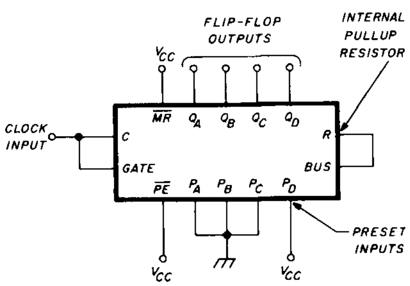
Fig. 3 indicates a single package connected for division by eight. Waveforms are expanded to show automatic presetting. Preset inputs are wired for binary 1000 (decimal 8). Control lines $\overline{MR}$ and $\overline{PE}$ are tied high. A preset can occur only when GATE is low and BUS high — the all-zeroes condition.

The waveforms assume that six clock inputs have occurred. Counter state is then binary 0010. The next clock (seventh) will make it binary 0001. The eighth clock will cause several actions.

The counter goes first to binary 0000 and the BUS goes high. External control GATE is connected to the clock. It is still high after binary 0000 has been reached, so a preset doesn't begin until the clock goes low. At that time, the internal D stage active low set changes Q_D from 0 to 1. The other three stages reset; it doesn't change anything since they are already 0.

The counter is now set to binary 1000 and BUS is low, but the next positive clock edge will change the counter state to binary 0111. It counts down again until all zeroes are present. The carry out is only the width of the clock low state.

Maximum clock input frequency is limited by three propagation delays: clock positive edge to BUS going



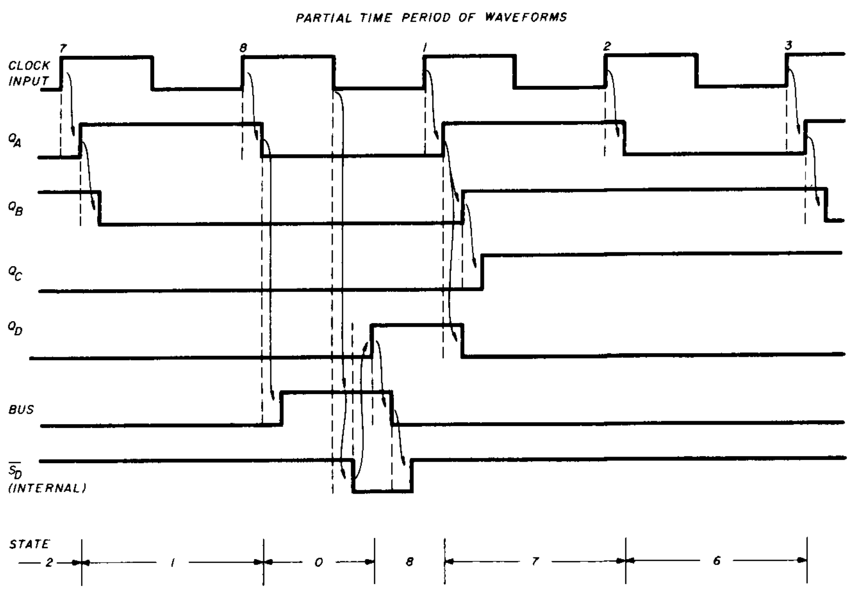


fig. 3. Single MC4016 connected as divide-by-eight.

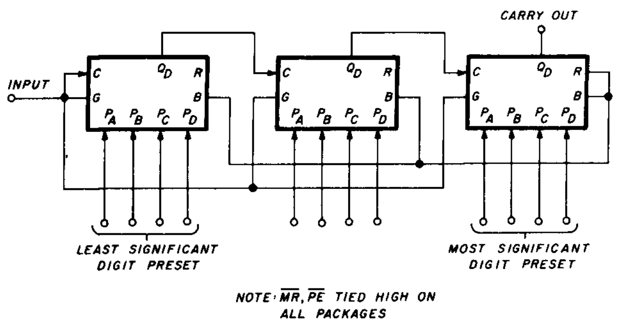


fig. 4. Three MC4016 counters in cascade for variable division.

high (65 ns maximum), clock negative edge to any flip-flop set (35 ns), and next-positive clock to a flip-flop toggle (78 ns). Inverse total is 5.6 MHz, but the nominal maximum frequency is 8 MHz.

cascading

Fig. 4 shows the circuit for three packages. It can divide by any number from 1 to 999 depending on the BCD input to each counter. An MC4018 can be substituted. It's a hexadecimal (divide-by-sixteen) version, and three packages would yield a maximum count of 2^{12}, or 4096; four-bit binary preset inputs would be required.

All $\overline{MR}$ and $\overline{PE}$ control lines are tied high. All GATE inputs are connected to the input clock. All BUS pins are tied together, but only one R or pullup resistor connection is required.

Preset action is the same as in **fig. 3** and depends on the first, or left-hand, counter for speed. The last, or right-hand, counter will reach all-zeroes first, then the middle. The BUS is almost ready to go high, but the wired-AND connection makes it dependent on the first counter. When the first counter goes zero, preset is enabled to all; the BUS goes high, then the common GATE goes low.

Carry out may be from the third Q_D, but fast inputs should use the BUS line since it's slightly wider. Speed is limited, but a few extra devices will increase this.

increasing speed

Input frequency can be increased to at least 25 MHz by adding a D flip-flop, 5-input gate, and three inverters as in **fig. 5.** Schottky TTL devices are recommended. Note that this version has the first counter's BUS pin grounded and all $\overline{MR}$ and GATE pins tied high.

Previous connections initiated a preset on the input clock low state. **Fig. 5** allows nearly a full clock period for preset. This is possible by arming the preset when countdown has reached decimal 002 (binary 0010 in the first counter). Presets have been hard wired for 888 division for illustration.

G7 goes low on the 886th input (representing decimal 002). The external flip-flop will toggle on the 887th input. This action initiates a preset by making the common $\overline{PE}$ control line low. Preset completion will make G7 high but won't change the external flip-flop, because its clock, the 888th, has not yet arrived.

Once the 888th clock arrives, the external flip-flop will toggle, but the counters will not change; $\overline{PE}$ is

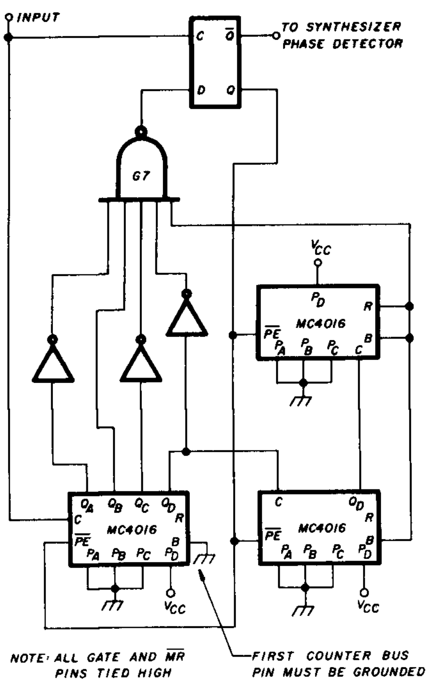
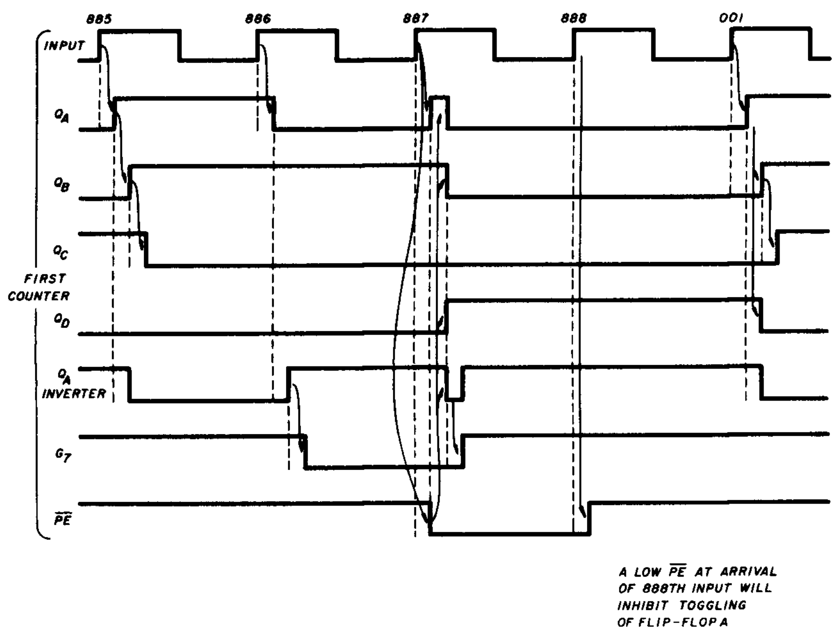


fig. 5. High-speed divide-by-888 for use with phase-locked loop.



still held low at clock edge, and all counters remain at preset at that time. (See the third state of the truth table in **fig. 2**.) The first counter essentially ignores the 888th clock.

The external circuitry permits a substantial increase in speed even though the counters are not synchronous. The only disadvantage is slight: division by less than three is not possible.

other packages

Presettable up/down counters are available. These can be connected for down counting only with external circuitry added for similar preset-enable control. The use of synchronous counters and Schottky TTL programmable dividers is possible up to 60 MHz. Great attention must be paid to propagation delay at high speed.

The Motorola device was selected for this example because it contains the essential ingredients of a counter with preset control ability.

reference

1. Leonard H. Anderson, "Digital Techniques: Gate Structure and Logic Families," *ham radio*, February, 1979, page 66.

ham radio

"I think you'd better unplug it, Stan."

More Details? CHECK—OFF Page 126

Full Features and Superior Performance

ST-6000 RTTY DEMODULATOR

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Accurately Tuned Rx Filters
Crystal Controlled Tx Tones
True Transceive Operation

Invert Both Rx Demod. and Tx Tones

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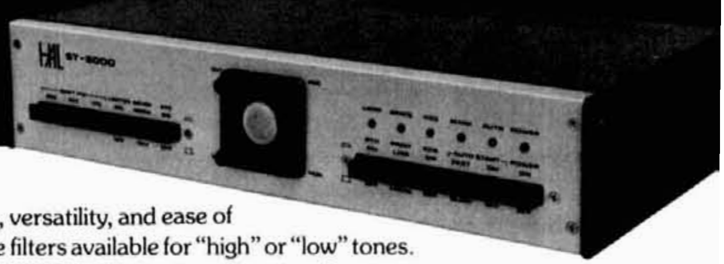

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The HAL ST-6000 Demodulator offers outstanding performance, versatility, and ease of operation. The Receive Demodulator features multiple-pole active filters available for "high" or "low" tones. These filters are frequency-matched to the transmit tone crystals for true transceive operation. Input bandpass filters, discriminator filters, and post-detection filters are carefully designed and tested for optimum weak-signal recovery. The ST-6000 has an internal loop power supply, 2 loop keyers, RS-232, MIL-188C, and CMOS data I/O, and rear panel connections to data and control circuits for connection to UART and computer devices. Use it with the HAL DS-3000 KSR for the best in RTTY performance. **$595.00**

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Model
1514

Drake WH-7 Directional RF Wattmeter

1.8-30 MHz

Specifications

Frequency Coverage	1.8-30 MHz
Line Impedance	50 ohm resistive
Power Capability	2000 W continuous
Jacks, Removable Coupler	Two S0239 input and output connectors
Semiconductors	Two power meter rectifiers
Accuracy	± (5% of reading + 1% of full scale)
VSWR Insertion	Insertion of wattmeter in line changes VSWR no more than 1.05:1
Shipping Weight	3 lbs (1.4 kg)
Dimensions	5.3"H x 6.9"W x 7.5"D (13.5 x 17.5 x 19 cm)

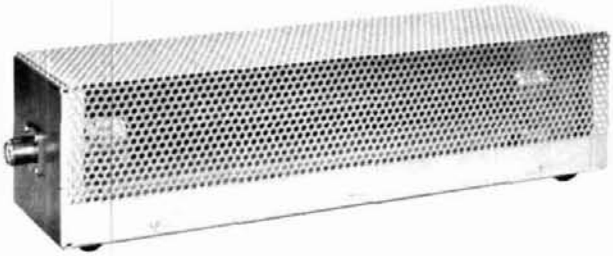
Drake directional, through line wattmeters, using printed circuits, toroids, and state of the art techniques, permit versatile performance and laboratory accuracy, yet at a lower cost.

Removable coupler provides remote metering, and allows convenient positioning of coaxial cable.

WH-7 wattmeter makes possible quick, accurate adjustments of antenna resonance and impedance match, when placed between transmitter and matching network.

Drake WH-7: Designed for user convenience and high accuracy. This instrument includes three calibrated scales for rf power to satisfy applications from QRP to high power (0-20, 0-200 and 0-2000 watts full scale) A fourth calibrated scale provides direct reading VSWR information, and is switch selected from front panel. The WH-7 is styled to match the 7-line.

Drake "Dry" Dummy Loads—no oil required



Model 1551 Drake DL-1000

- **1000 watts** for 30 seconds, with derating curve to 5 minutes. Designed to accept Drake FA-7 cooling fan for extended high power operation.
- **VSWR of 1.5:1 max.** 0-30 MHz.
- Provided with SO-239 coax connector, and rubber feet for desk or bench use.
- **Size** 14" x 3.6" (35.6 x 9.1 cm). Wt. 2 lbs (910 g)

Model 1550 Drake DL-300

- **300 watts** for 30 seconds, with derating curve to 5 minutes.
- **Built-in PL-259** coax connector for direct connection to rear of transceiver or transmitter-no jumper coax necessary.
- **VSWR of 1.1:1 max.** 0-30 MHz 1.5 max. 30-160 MHz.
- Ideal as bench test device for amateur or commercial hf and vhf gear.
- **Small size** fits conveniently in any field service tool box. 6.7" x 2.08" (17.0 x 5.3 cm). Wt. 11 oz (310 g)

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High Pass Filters for TV Sets

provide more than 40 dB attenuation at 52 MHz and lower. Protect the TV set from amateur transmitters 6-160 meters.

Model No. 1603
Drake TV-300-HP
For 300 ohm twin lead. New terminals for easy installation.

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For 75 ohm TV coaxial cable; TV type "F" connectors installed.

Low Pass Filters for Transmitters

have four pi sections for sharp cut off above the hf amateur bands and to attenuate transmitter harmonics falling in any TV channel and fm band. 52 ohm. SO-239 connectors built in.

Model No. 1608 **Drake TV-3300-LP**

1000 watts max. below 30 MHz. Attenuation better than 80 dB abo 41 MHz. Helps TV i-f interference, as well as harmonic interference.

Model No. 1605 **Drake TV-42-LP**

is a four section filter designed with 43.2 MHz cut-off and extremely high attenuation in all TV channels for transmitters operating at 30 MHz and lower. Rated 100 watts input.

Drake TVI Filters help you keep peace with your neighbors

solid state continuous coverage synthesized hf system

Model 1337

DRAKE ®

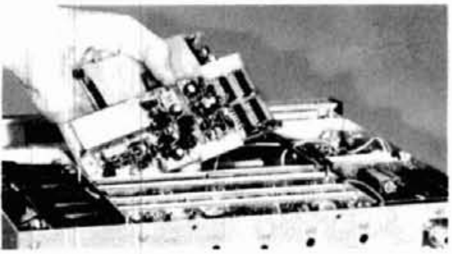
The design philosophy behind the new Drake "7 system" has created a most sophisticated system concept, extending from engineering to the visual appearance of the system and each of its parts.

The TR-7 System is the result of one of the most extensive engineering and development programs in the history of the R. L. Drake Company, and provides the user with many innovative design features.

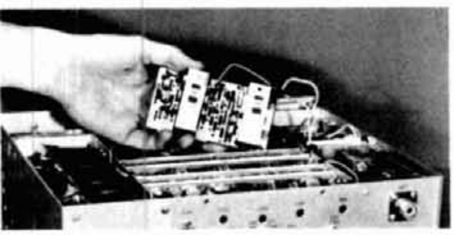
With the excellent design of its front panel and controls, the system is simple and straightforward to operate–makes state of the art performance a pleasure.

Broadband, Solid State Design—100% solid state throughout. All circuits are broadbanded so there is no need for preselection tuning or transmitter adjustments of any kind.

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



Continuous, Wide Range Frequency Coverage—The TR-7/DR-7 provides reception from 1.5 thru 30 MHz—continuously, and **zero** thru **30 MHz** continuously with the optional Aux-7 Range Program Board. No gaps or range crystals required. The highly advanced Drake Synthesizer makes this possible, and is an industry first. The TR-7/DR-7 provides transmit coverage for all Amateur Bands 160 thru 10 meters. With the optional Aux-7 Range Program Board, diode-programmable out-of-band transmit coverage is available for MARS, Embassy, Government, and future band expansions in the range 1.8 thru 30 MHz.* The Aux-7 Board also provides 0 thru 1.5 MHz receive coverage and crystal-controlled fixed channel operation for Government, Amateur, or semi-commercial applications anywhere in the hf range. The TR-7 w/o DR-7 provides coverage of the Amateur Bands 160 thru 15 meters and the 28.5-29.0 MHz range of 10 meters. The Aux-7 Range Program Board is also useable in the standard TR-7 for extra range coverage as noted.

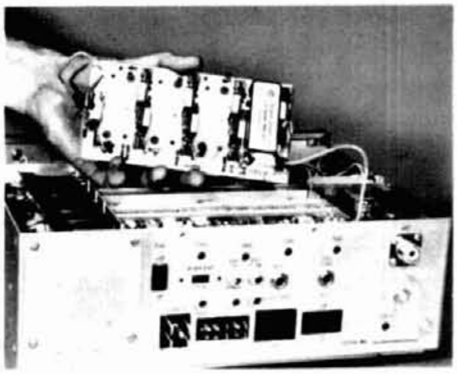


State of the Art Receiver Design—The Drake TR-7 introduces another industry first for amateur transceivers: "Up-Conversion," in combination with a special high level double balanced mixer for superior strong signal handling, spurious and image response performance. The first i-f of 48.05 MHz places images well outside the receiver passband, and provides for true general coverage operation without i-f gaps.

True Passband Tuning—The TR-7 employs the famous Drake Full Passband Tuning instead of the limited range "i-f shift" found in some other units. The Drake System tunes from the top edge of one sideband, through center, to the bottom edge of the other sideband. In fact, the range is even wider to accommodate RTTY. Full passband tuning greatly improves receiving performance in heavy QRM.

Unique Independent Receive Selectivity—Optional receiving selectivity filters can be installed internally and pushbutton-selected from the front panel. These may be selected independently of transmit mode and provide optimum response for various conditions of ssb, cw, RTTY, and a-m. You may also transmit cw while receiving ssb, or vice versa, or even transmit one sideband while receiving the other. The standard filter is 2.3 kHz for ssb. You may choose from optional 300 Hz, 500 Hz, a special 1.8 kHz for crowded ssb, or 6 kHz filter for a-m. Any three may be installed in addition to the ssb filter.

Effective Noise Blanker—This accessory is custom engineered to provide true impulse-type noise blanking performance.



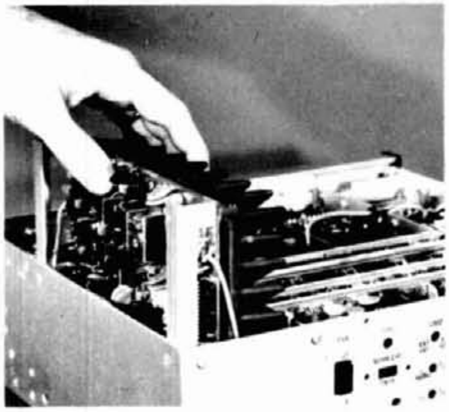
Special High Power Solid State PA—A Drake custom-designed diagonal heat sink provides for an internally mounted power amplifier with nothing mounted outboard subject to physical damage. The unique air ducting effect of this amplifier allows an optional rear-mounted fan to provide continuous duty on SSTV/RTTY. Continuous ssb/cw

(TR-7 features continued on next page)

ote: Out-of-band transmitter coverage for MARS, Government, etc. is available nly in ranges authorized by the FCC, Military, or other government agency for a pecific service. Proof of license for that service must be submitted to the R. L. rake Company, including the 500 kHz range to be covered. Upon approval, and at the discretion of the R. L. Drake Company, a special range IC will be supplied for use with the Aux-7 Range Program Board. Prices quoted from the factory. See operator's manual for details. (Not available for services requiring type acceptance.)

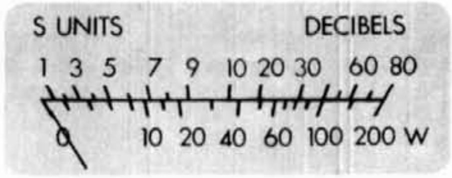
(Continued from preceding page)

DRAKE TR-7 solid state continuous coverage synthesized hf system



operation is available without the fan, due to the excellent heat sink design. The optional Drake PS-7 Ac Supply is rugged, rated for continuous duty, and will easily handle power requirements. The System is rated 250 watts input—in any of its modes. Fully VSWR protected.

TR-7 Internal Test Facilities—As well as the standard "S" meter function, the TR-7 metering includes a built-in rf Wattmeter/VSWR Bridge. Also, the DR-7 digital counter reads frequencies to 150 MHz for test purposes. Access to the counter is from the rear panel.



Receiver Incremental Tuning (RIT)—Complete RIT Flexibility is provided for both the TR-7 and RV-7 remote VFO for maximum convenience. The RV-7 also includes a special "spot" function for easy zero beating.

Model 1337 Drake TR-7 Transceiver

Model 1530 Drake DR-7 General Coverage/ Digital Readout Board

Model 1336 Drake TR-7/DR-7 General Coverage Digital R/O Transceiver

Model 1338 Drake RV-7 Remote VFO

Model 1502 Drake PS-7 120/240V Ac Supply includes special wide range voltage and frequency capability. Operates from any nominal line voltage (90-132 V/ 180-264 V; 50-60 Hz) ideal for overseas

Model 1536 Drake Aux-7 Range Program Board†

Model 1531 Drake MS-7 Matching Speaker

Model 1537 Drake NB-7 Noise Blanker

Model 1529 Drake FA-7 Fan

Model 7021 Drake SL-300 Cw Filter, 300 Hz

Model 7022 Drake SL-500 Cw Filter, 500 Hz

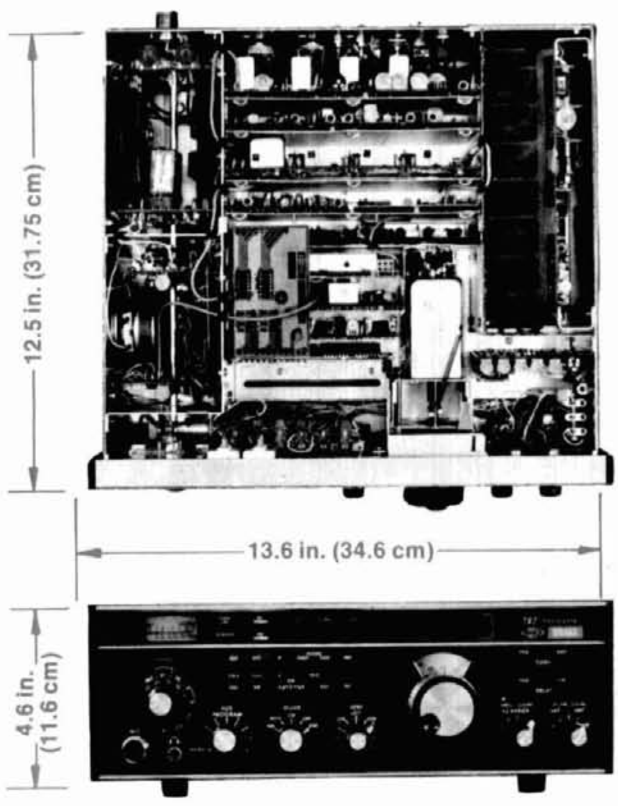
Model 7023 Drake SL-1800 Ssb/RTTY Filter, 1.8 kHz

Model 7024 Drake SL-6000 A-m Filter, 6.0 kHz

Model 1335 Drake MMK-7 Mobile Mounting Kit

Model 7037 Drake TR-7 Service Kit/Extender Board Set

Model 385-0004 Drake TR-7 Service/Schematic Book



DRAKE TR-7 SPECIFICATIONS

GENERAL

Frequency Coverage
(with DR-7 Digital R/O Gen. Cov. Board)

Receive
Without Aux-7 ... 1.5 to 30 MHz, continuous, no gaps
With Aux-7† ...... Same, plus 0 to 1.5 MHz at reduced performance in this range

Transmit
Without Aux-7 ... 1.8-2.0, 3.5-4.0, 7.0-7.5, 14.0-14.5, 21.0-21.5, 28.0-30.0 MHz
With Aux-7† ..... Above ranges, plus any eight 500 kHz segments from 1.8 to 30 MHz

Frequency Coverage
(without DR-7 Digital R/O Gen. Cov. Board)

Receive/Transmit (Transmit above 1.8 MHz only)
Without Aux-7 ... 1.5-2.0, 3.5-4.0, 7.0-7.5, 14.0-14.5, 21.0-21.5, 28.5-29.0 MHz, plus Receive only on 2.5-3.0 MHz and 5.0-5.5 MHz
With Aux-7† ..... Above ranges, plus any eight 500 kHz segments from 0 to 30 MHz, (0 to 1.8 MHz Receive only)

Modes of Operation ... Usb, Lsb, Cw, RTTY, A-m equiv. (A-3H)

Frequency Stability ... *Total drift is less than 100 Hz after* warm up. Total frequency change is less than 100 Hz over the 11-16 V-dc input supply range

Frequency Readout Accuracy
Analog .... Better than ± 1 kHz when calibrated at the nearest marker point
Digital .... 15 ppm ± 100 Hz

External Counter Mode
Maximum
Input Frequency ... 150 MHz
Input Level Range ... 50 mV to 2 V, rms

Power Supply
Requirements .... 11-16 V-dc (13.6 V-dc nominal), 3A receive, 25A transmit

Dimensions
Depth .... 12.5 in. (31.75 cm), excluding knobs and connectors.
Width .... 13.6 in. (34.6 cm)
Height .... 4.6 in. (11.6 cm), excluding feet
Weight .... 17.1 lb. (7.75 kg)

RECEIVER

(1.8-30 MHz, reduced specs 0-1.8 MHz)

Sensitivity
Ssb, Cw .... Less than 0.5 μV for 10 dB (S+N)÷N
A-m (30% Mod.) ..... Less than 2.0 μV for 10 dB (S+N)÷N

Selectivity .... 2.3 kHz at −6 dB and 4.1 kHz at −60 dB (1.8:1 shape factor)

Ultimate Selectivity ... Greater than 100 dB

Agc .... Less than 4 dB output variation for 100 dB input signal change, referenced to agc threshold

Intermodulation .... Intercept Point, +20 dBm
Two-tone Dynamic Range, 99 dB
(at tone spacings of 50 kHz and greater)

I-f Frequency .... First I-f ....48.05 MHz
Second I-f ....5.645 MHz

Image and I-f Rejection Greater than 80 dB

Spurious Response ... Greater than 60 dB down

Internally Generated
Spurious .... Less than 1 μV equivalent, except 3 μV equivalent from 5 to 6 MHz. (Reduced specs on internal osc frequencies)

Audio Output .... 2.0 watts @ less than 10% THD (4 ohm load)

TRANSMITTER

Power Input (Nominal)
Ssb .... 250 watts PEP
Cw .... 250 watts
A-m equiv. .... 80 watts (carrier), plus upper sideband

Load Impedance ...... 50 ohms, nominal

Spurious Output ...... Greater than 50 dB down

Harmonic Output ...... Greater than 45 dB down

Intermodulation
Distortion .... 30 dB below PEP (24 dB below one of two tones)

Undesired Sideband
Suppression ...... Greater than 60 dB @ 1 kHz

Duty Cycle
Ssb, Cw 100%
Tune, SSTV, RTTY, A-m w/o 1529 FA-7 Fan: 33%, 5 min. transmit, max.
with 1529 FA-7 Fan: 100%

Wattmeter Accuracy ... ±5% @ 100 watts (50 ohm load)

Carrier Suppression ... Greater than 50 dB

Microphone Input ..... High impedance

VSWR Turndown (Nominal) (Percent rf power turndown)
@ 1:1 .... 0%
@ 2:1 .... 10%
@ 3:1 .... 25%
@ 4:1 .... 50%
@ 5:1 and above 90%

† Aux-7 must be used with either Model 1546 RRM-7 Range Receive Module, or Model 1547 RTM-7 Range Transceive Module. Use one module per 500 kHz range. Modules plug directly into Aux-7.

R. L. DRAKE COMPANY DRAKE®
540 Richard St., Miamisburg, Ohio 45342
Phone: (513) 866-2421 • Telex: 288-017

DRAKE® UV-3 uhf-vhf fn

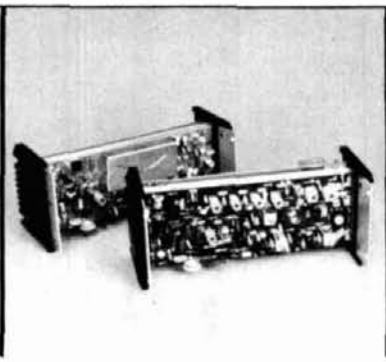
Optional
Drake 1525EM
Encoding Mike

Designed and manufactured in U.S.A.

- Fully synthesized on each band, 5 kHz steps, digital read-out.
- Fm coverage on complete 144, 220 and 440 Amateur Bands, depending on model purchased. Completely band-switched from front panel.
- Four extra diode programmable fixed channels, with offsets, available for each band, *in addition to the* synthesizer.
- Diode programmable non-standard offsets available for each band.
- Separate SO-239 Antenna Connector for each band.
- Outstanding receiver front-end performance. Ideal for use in metropolitan areas where many repeaters are in use.
- Squelch.
- Hi-lo power, with lo-power adjustable.
- Priority scan feature:
  —scan a programmed fixed channel from any synthesizer frequency.
  —scan any synthesizer frequency *from a programmed fixed channel.*
  —scan a specific programmed fixed channel from another programmed fixed channel.
- Plug-in modular construction.
- *Remote operation. Removable control* head will operate radio in trunk compartment from driver seat. (remote kit optional)
- No frequency mixing in transmitter. Transmitter frequency derived directly from VCO frequency. Provides extremely low spurious output.
- Companion ac power supply (PS-3).
- Operate mobile or fixed station. (13.8 V supply required)
- Small, compact, rugged construction utilizing aluminum extrusion sides and panel.
- Transmit audio custom tailored for maximum communications "punch."
- *Choice of one, two, or three band* coverage in a single transceiver. Basic models may be purchased, with factor installed add-on modules added later.

PS-3 Ac Power Supply

1525EM Encoding Microphone



220 and 440 Add-on Modules

UMK-3 Remote Trunk-Mount Kit

3-band system

Fully synthesized on each band

144 220 440 MHz

DRAKE UV-3 SPECIFICATIONS

GENERAL

Frequency Coverage: 144 .... 144-148 MHz*
220 .... 220-225 MHz
440 .... 440-450 MHz

Mode: Fm (5 kHz deviation)

Supply Voltage: 11.5-15.0 V dc negative ground

Supply Current: Receive .... 0.9 A Standby
Transmit .... 6 A High Power
1.3 A Low Power

Dimensions: Length (single unit) .... 9" (22.86 cm)
(two unit) .... 11.5" (29.2 cm)
(three unit) .... 14" (35.56 cm)
Width .... 8.1" (20.6 cm)
Height .... 3.5" (8.9 cm)

Weight: (One unit) .... 7 lbs. (3.17 kg)
(Two unit) .... 7.3 lbs. (3.31 kg)
(Three Unit) .... 7.6 lbs. (3.45 kg)

Operating Temperature: 0°C to 60°C

*Band overlap allows tuning of most Mars frequencies

FREQUENCY SYNTHESIZER

Type: Directly programmable, digital phase locked loop, 5 kHz steps

Reference: 5 MHz crystal oscillator

Frequency Accuracy: ±.0005% over a temperature range of 0°C to 60°C with a supply voltage variation of 11.5 to 15 V dc

RECEIVER

Type: Double conversion, 1st i-f @ 10.7 MHz, 2nd i-f @ 455 kHz, 6 pole crystal filter @ 10.7 MHz and 8 pole ceramic filter at 455 kHz

Selectivity: 12 kHz @ −3 dB

Sensitivity: 146-148 MHz, 222-225 MHz, 442-447 MHz: Typically less than .35μV for 12 dB SINAD
144-148 MHz, 220-225 MHz, 440-450 MHz: .5 μV (max.) for 12 dB SINAD

Adjacent Channel Rejection: 144 .... greater than 80 dB min. @ ± 30 kHz
220, 440 .... greater than 70 dB min. @ ± 30 kHz
144, 220, 440 .... greater than 60 dB min. @ ± 15 kHz

Intermodulation Attenuation: (EIA RS-204-A) 144 .... 80 dB (referenced to 12 dB SINAD)
220 .... 75 dB (referenced to 12 dB SINAD)
440 .... 65 dB (referenced to 12 dB SINAD)

Image Rejection: 144 .... 80 dB
220 .... 60 dB
440 .... 50 dB

I-f Rejection: Greater than 95 dB

Audio Output: 2.5 watts @ less than 10% THD, 2 watts @ less than 5% THD

Squelch Sensitivity: Less than 0.2 μV

Meter: Indicates relative signal level

TRANSMITTER

Power Output (13.8 V dc): High Power: 144 .... 25 watts nom. (144-148 MHz)
220 .... 10 watts min. (220-225 MHz)
440 .... 10 watts min. (440-450 MHz)
Low Power: Approx. 10% of high power (adjustable)

Harmonic and Out of Band Spurious: 144, 220 .... −60 dB (min.) referenced to carrier
440 .... −40 DB (min.) referenced to carrier

Spurious in Band: −75 dB (min.) referenced to carrier

Modulation: Direct fm, pre-set to ±5 kHz deviation

Hum and Noise: Greater than 40 dB below maximum deviation

Model 1346 Drake UV-3 (144-220-440)
Model 1344 Drake UV-3 (144-440)
Model 1340 Drake UV-3 (144)

(Models above include factory installed modules for bands as listed, standard dynamic mike, and mobile mounting bracket.)

Add-on modules expand band coverage of models which may have been purchased in a single band or two band configuration. Prices include factory installation which is necessary to meet FCC receiver certification requirements.

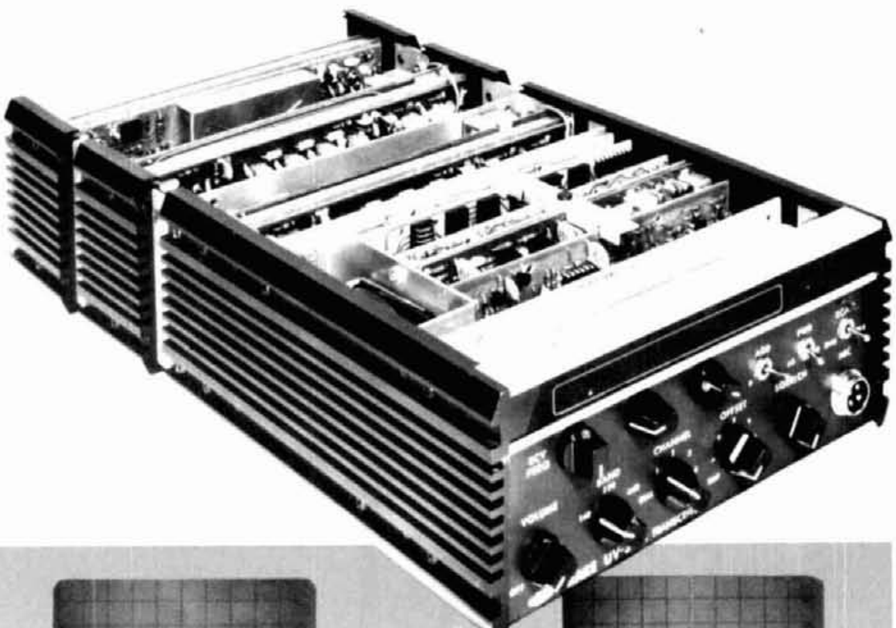
220 Add-on Module
440 Add-on Module
144 Add-on Module

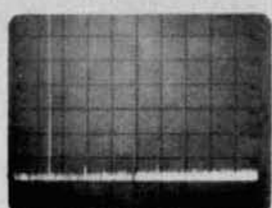
Model 1504 Drake PS-3 AC Power Supply

Model 1525 Drake 1525EM Encoding Mike (see next page)

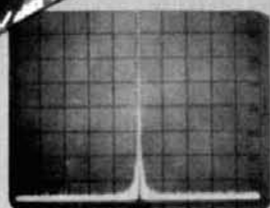
Model 1330 Drake UMK-3 Remote Trunk-Mount Kit

Model 385-0002 Drake UV-3 Service/Schematic Book




100 MHz/Division

UV-3 Frequency Spectrum
146.520 MHz
25 Watts


500 kHz/Division

DRAKE® MICROPHONES

Drake 1525EM Push Button Encoding Mike

Model 1525

- Microphone and auto-patch encoder in single convenient package with coil cord and connector. Fully wired and ready for use.
- High accuracy IC tone generator, no frequency adjustments.
- High reliability Digitran® keyboard.
- Power for tone encoder obtained from transceiver through microphone cable. No battery required. Low current drain.
- Low output impedance allows use with almost all transceivers.
- Four pin microphone plug: directly connects to Drake UV-3 without any modification in transceiver. Compatible with all previous Drake and other 2 meter units with minor modifications.
- Tone level adjustable

Drake 7077 Dynamic Desk Microphone

- **Audio and level characteristics** custom designed to match the transmit audio requirements of the Drake TR-7.
- **Features both VOX and PTT** operation without modification.
- **High Impedance**
- Includes coil cord and plug wired for direct installation to the Drake TR-7.
- Style and color provide a beautiful match to the Drake 7-Line.
- Size 4.3"W x 5.8"D x 9.3"H (10.9 x 14.7 x 23.6 cm). Wt. 1 lb. 7 oz (650 g).

R. L. DRAKE COMPANY DRAKE®
540 Richard St., Miamisburg, Ohio 45342
Phone: (513) 866-2421 • Telex: 288-017

Drake L-7 Continuous Duty 160-10* Meters 2kW Linear Amplifier

Temperature controlled design for "key-down" operation over a wide frequency range. Newly engineered for coverage of any new or expanded hf amateur bands within FCC amplifier rules. Also features wide frequency coverage for MARS, and other services authorized for this type of amplifier.

2 kW PEP, 1 kW cw, RTTY, SSTV operation—all modes, full rated input, continuous duty cycle.

160-10* meter amateur band coverage, plus expanded ranges for any future hf band expansions or additions within FCC rules. These ranges also include increased coverage for MARS, embassy, government, or other such services.

The Drake L-7 utilizes a pair of Eimac 3-500 Z triodes for rugged use, and lower replacement cost compared to equivalent ceramic types. Tubes are included.

Accurate built-in rf wattmeter, with forward/reverse readings, is switch selected. Calibrated 300/3000 watt scales.

Temperature controlled two speed fan is a high volume low noise type and offers optimum cooling.

Adjustable exciter agc feedback circuitry permits drive power to be automatically controlled at proper levels to prevent peak clipping and cw overdrive. Front panel control.

By-pass switching is included for straight through, low power operation without having to turn off amplifier.

Bandpass tuned input circuitry for low distortion and 50 ohm input impedance.

Amplifier is comprised of two units—rf deck for desk top and separate power supply.

Operates from 120/240 V ac, 50/60 Hz primary line voltage.

DRAKE L-7 SPECIFICATIONS

Frequency Coverage*: Ham bands 160 through 15 meters. Non-amateur frequencies between 6.5 and 21.5 MHz may be covered with some modification of the input circuit.

Plate Power Input: 2000 Watts PEP on SSB and 1000 Watts DC on CW, AM, RTTY, and SSTV.

Drive Power Requirements: 100 Watts PEP on SSB and 75 Watts on CW, AM, RTTY, and SSTV.

Input Impedance: 50 Ohms. (Bandpass tuned input)

Output Impedance: Adjustable pi-network matches 50 Ohm line with SWR not to exceed 2:1.

Intermodulation Distortion Products: In excess of −33 dB.

Wattmeter Accuracy: 300 Watts forward and reflected, ± (5% of reading + 3 Watts). 3000 Watts forward, ± (5% of reading + 30 Watts).

Power Requirements: 240 Volts 50-60 Hertz 15 Amperes, or 120 Volts 50-60 Hertz 30 Amperes.

Tube Complement: Two of 3-500Z or 8802/3-500Z or 8163 or 3-400Z.

Dimensions: Amplifier 13.69"W x 6.75"H x 14.25"D (34.8 x 17.1 x 36.2 cm). **Power Supply** 6.75"W x 7.88"H x 11"D (17 x 20 x 28 cm).

Weight: Amplifier 27 lbs (12.25 kg). **Power Supply** 42.5 lbs (19.3 kg).

*Export model includes coverage of the 10-meter Ham Band.

Drake R-7 Synthesized, General Coverage Receiver

Model 1240

Full general coverage receptio 0-30 MHz, with no gaps or range crystals required.

Continuous tuning all the way from vlf thru hf. Superb state-of-the-art performance on a-m, ssb, RTTY, and cw —and it transceives with the Drake TR-7.

100% solid state broadband design, fully synthesized with a permeability tuned oscillator (PTO) for smooth, *continuous tuning.*

Covers the complete range 0 to 30 MHz with no gaps in frequency coverage. Both digital and analog frequency readout.

Special front-end circuitry employing a high level double balanced mixer and 48 MHz "up-converted" 1st i-f for superior general coverage, image rejection and strong signal handling performance.

Complete front-end bandpass filters are included that operate from hf thru vlf. External vlf preselectors are not required.

10 dB pushbutton-controlled broadband preamp can be activated on all ranges above 1.5 MHz. Low noise design.

Various optional selectivity filters for cw, RTTY and a-m are switch-selected from the front panel. Ssb filter standard.

Special new low distortion **"synchro-phase" a-m detector** provides superior international shortwave broadcast reception. This new technique permits 3 kHz a-m sideband response with the use of a 4 kHz filter for better interference rejection.

Tunable i-f notch filter effectively reduces heterodyne interference from nearby stations.

The famous Drake full electronic passband tuning system is employed, permitting the passband position to be adjusted for any selectivity filter. This is a great aid in interference rejection.

Three agc time constants plus "Off" are switch-selected from the front panel.

Complete transceive/separate **functions** when used with the Drake TR-7 transceiver are included, along with separate R-7 R.I.T. control.

Special multi-function antenna selector**/50 ohm splitter** is switch-selected from the front panel, and provides simultaneous dual receive with the TR-7. This makes possible the reception of two different frequencies at the same time. Main and alternate antennas and vhf/uhf converters may also be selected with this switching network.

The digital readout of the R-7 may be used as a 150 MHz counter, and is switched from the front panel. Access thru rear panel connector.

The built-in power supply operates from 100, 120, 200, 240 V-ac, 50/60 Hz, or nominal 13.8 V-dc.

The R-7 includes a built-in speaker, or an external Drake MS-7 speaker may be used.

Built-in 25 kHz calibrator for calibration of analog dial.

Low level audio output for tape recorder.

Up to eight crystal controlled fixed channels can be selected. (With Drake Aux-7 installed.)

Optional Drake NB-7A Noise Blanker available. Provides true impulse type noise blanking performance.

DRAKE® R. L. DRAKE COMPAN

Optional accessories available

Model 1531 Drake MS-7 Speaker
Model 7021 Drake SL-300 Cw Filter, 300 Hz
Model 7022 Drake SL-500 Cw Filter, 500 Hz
Model 7023 Drake SL-1800 Ssb/RTTY Filter, 1800 Hz
Model 7024 Drake SL-6000 A-m Filter, 6.0 kHz
Model 7026 Drake SL-4000 A-m Filter, 4.0 kHz
Model 1532 Drake NB-7A Noise Blanker
Model 1536 Drake Aux-7 Range Program/Fixed-Frequency Board

DRAKE R-7 SPECIFICATIONS

Frequency Coverage, continuous tuning (With Drake DR-7 Digital R/O, General Coverage Board)
0 to 30 MHz continuous (With or without Aux-7 board) (No gaps in frequency coverage)

Frequency Coverage, continuous tuning (Without DR-7 Board installed)

Without Aux-7 Board		
0.01 to 0.5 MHz	Without Aux-7 Board	5.0 to 5.5 MHz
0.5 to 1.0 MHz		7.0 to 7.5 MHz
1.0 to 1.5 MHz		14.0 to 14.5 MHz
1.5 to 2.0 MHz		21.0 to 21.5 MHz
2.5 to 3.0 MHz		28.5 to 29.0 MHz
3.5 to 4.0 MHz		

Plus any eight additional 500 kHz segments between 0 and 30 MHz when programmed into Aux-7 Board.

Crystal Controlled Fixed Frequencies: Up to eight crystal-controlled fixed frequencies within the 0-30 MHz range with Aux-7 Accessory Board. Proper 500 kHz range for desired fixed frequency is also programmed into Aux-7.

Frequency Stability: Less than 100 Hz drift after temperature stabilization including ±10% line voltage variation.

Digital Readout Accuracy: (DR-7 installed) 15 PPM ± 100 Hz

Analog Dial Accuracy: Better than ±1 kHz when calibrated to nearest calibrator marker.

Modes of Operation: Ssb, cw, RTTY, SSTV, a-m.

Sensitivity (ssb): 1.8-30 MHz Less than .20μV for 10dB S+N/N with preamp on (typically .15μV) (Noise floor typically −134 dBm) Less than .50μV for 10 dB S+N/N without preamp (typically .30μV) (Noise floor typically −128 dBm). .01-1.5 MHz Less than 1.0μV for 10 dB S+N/N

Sensitivity (a-m): 1.8-30 MHz Less than 1.2μV for 10dB S+N/N @ 30% modulation, preamp on. Less than 2.0μV for 10 dB S+N/N @ 30% modulation, preamp off. .01-1.5 MHz Less than 4.0μV for 10 dB S+N/N @ 30% modulation.

Selectivity (2.3 kHz filter supplied): 2.3 kHz at −6 dB, 4.2 kHz at −60 dB (1.8:1) shape factor. Optional 300 Hz, 500 Hz, 1800 Hz and 4 kHz filters are available as follows:

Ultimate Selectivity: Greater than 100 dB

Accessory Crystal Filters

SL-300 cw filter: 300 Hz @ 6 dB, 700 Hz @ 60 dB
SL-500 cw, RTTY Filter: 500 Hz @ 6 dB, 1100 Hz @ 60 dB
SL-1800 ssb/RTTY Filter: 1800 Hz @ 6 dB, 3600 Hz @ 60 dB
SL-4000 a-m Filter: 4 kHz @ 6 dB, 8 kHz @ 60 dB
SL-6000 a-m Filter: 6 kHz @ 6 dB, 12 kHz @ 60 dB

Strong Signal Handling

Two-tone dynamic range: 99 dB* — 1.8-30 MHz
Third order intercept point: +20 dBm — preamp off
Two-tone dynamic range: 95 dB* — 1.8-30 MHz
Third order intercept point: +10 dBm — preamp on
Blocking: >145 dB above noise floor

**(At tone spacings of 50 kHz and greater)*

I-f and Image Rejection: Greater than 80 dB (48.05 MHz 1st i-f) (5.645 MHz 2nd i-f) (50 kHz 3rd i-f)

Agc Performance; Less than 4 dB audio output variation for 100 dB input signal change above agc threshold. Agc threshold is typical .8μV with preamp off and .25μV with preamp on.

Attack time: 1 millisecond. Three selectable release times: Slow—2 seconds; Med—400 m sec; Fast—75 m sec. Also, "Off" position is provided.

Antenna Input Impedance: Nominal 50 ohms

Audio Output: 2.5 watts with less than 10% T.H.D. into nominal 4 ohm load.

Power Requirements: 100/120/200/240 V-ac ±10%, 50/60 Hz, 60 watts or 11.0 to 16.0 V-dc (13.8 V-dc nominal), 3 amps

External Counter Mode (DR-7 installed): Readout: to 100 Hz. Accuracy: 15 PPM ±100 Hz. Maximum input frequency: 150 MHz. Input level range: 50 mV to 2 V rms.

Dimensions/Weight:

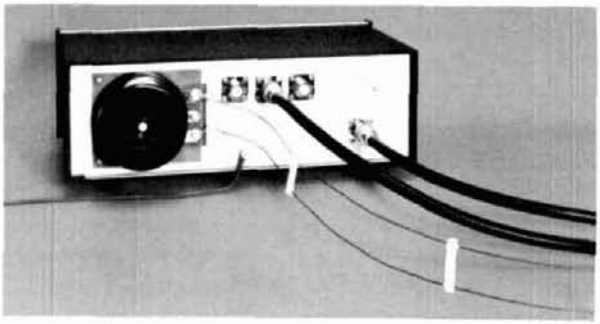
Depth— 13.0 in (33.0 cm) excluding knobs and connectors.
Width— 13.6 in (34.6 cm)
Height— 4.6 in (11.6 cm) excluding feet
Weight— 18.4 lbs (8.34 kg)

DRAKE® MATCHING NETWORKS

Precision instruments providing rf radiation control and measurement *for your communication system*

Drake MN-2700 2kW Matching Network

Model 1539



Drake B-1000 Balun Model 1510 installed on Matching Network.

The Drake MN-2700 manages rf radiation in the areas of impedance match to the antenna, rf power measurement, VSWR measurement, reduction of harmonic radiation, and antenna selection.

DRAKE MN-2700 FEATURES

160 thru 10 Meters Frequency Coverage—With out-of-band coverage for MARS, future band expansions and other applications.

Antenna Choice—Matches antennas fed with coax, balanced line, or random wire. (For balanced line use optional Drake B-1000 Balun, which mounts on rear panel of MN-2700.)

Antenna By-pass Switching—Unique design allows unit to be switch-by-passed regardless of which antenna is in use, whether coax or wire type. No need to manually disconnect feedlines. Switch also selects various antennas.

Extra Harmonic Reduction to help fight TVI—Drake Matching Networks employ special "pi-network" low-pass filter type circuitry for maximum harmonic rejection. This feature alone makes the MN-2700 a worthwhile investment; it is a Drake exclusive.

Built-in Metering—Accurate rf wattmeter/VSWR bridge is pushbutton controlled from front panel.

Power Capability—2000 watts PEP, 1000 watts average. Continuous Duty.

Dimensions—13.09"W x 4.53"H x 13"D including connectors (33.26 x 11.5 x 33 cm); **Weight** 11 lbs. (5 kg).

Drake MN-7 Matching Network Model 1538

Same features and specifications as the Drake MN-2700, but rated at 250 watts continuous. Same width and height, but only 8.5" (21.6 cm) in depth, and weighs 10 lbs. (4.55 kg). Meter reads 0-300 watts forward power or VSWR.

DRAKE MN-2700 SPECIFICATIONS

• **Frequency Coverage:** 1.8 to 30 MHz. Band Switch marked for 160, 80, 40, 20, 15, and 10 meter amateur bands; however, frequency coverage between amateur bands is possible by using the nearest band positions with a small reduction in matching capability. • **Input Impedance:** 50 ohms (resistive). • **Load Impedance:** 50 ohm coaxial with VSWR of 5:1 or less at any phase angle (3:1 on 10 meters). 75 ohm coaxial at a lower VSWR can be used. • **Balanced Feedlines:** With the Drake B-1000 accessory balun, which mounts on rear panel, tunes feed point impedances of 40 to 1000 ohms, or 5:1 VSWR referenced to 200 ohms (3:1 on 10 meters). • **Long-Wire Antennas** Feed point impedances up to 5:1 VSWR referenced to 50 ohms. Also, 5:1 referenced to 200 ohms with the Drake B-1000 accessory balun (3:1 on 10 meters). • **Meter:** Reads VSWR or forward power, 0-200 watts or 0-2000 watts. • **Wattmeter Accuracy:** ±5% of reading ±1% of full scale. • **Insertion Loss:** 0.5 dB or less on each band after tuning. • **Front Panel Controls:** Provide for the adjustment of resistive and reactive tuning, antenna switching, band switching, VSWR calibration, and selection of watts or VSWR functions of the meter. • **Rear Panel Connectors:** The rear panel has four type SO-239 connectors (one for input and 3 for outputs), three screw terminal connections (for long-wire and open-wire feeder systems), and a ground post.

really a "secret weapon" for 160 meter enthusiasts!

The Drake MN-2700 and MN-7 Matching Networks have a truly unique antenna feed switching design

Both matching networks will completely change the mode of a balanced-line fed 135 foot doublet to a special configuration that provides very effective 160 meter performance. And best of all, it's done with the simple flip of a switch on the front panel.

Consider a typical all-band antenna set-up—a 135 foot doublet, center-fed with 60 to 70 feet of balanced line at a height of 45 to 60 feet. The Drake MN-2700/B-1000 or MN-7/B-1000 will match this as a true balanced system on 80 thru 10 meters. (Fig. 1)

But what about 160 meters? Many amateurs recommend tying the feeders together and using the antenna as a vertical with a "top-hat." In fact, we suggest this ourselves in our manual.

However, the use of this, or any vertical, assumes you have a good ground or radial system for efficient operation. If you do not have enough room or do not wish to install such a radial system, performance may suffer. And if you do have radials, you still have to change the feeder connections each time you operate 160 meters.

On the other hand, when you use the MN-2700/B1000 or MN-7/B-1000 simply leave the feeders in the balanced connection as you would for 80 thru 10, and move the special antenna selector switch to Position No. 4. This automatically converts half of the antenna and feedline to an inverted "L", fed through a 4:1 impedance transformer, with the other half operating as a counterpoise. (Fig. 2)

This system offers the convenience of "stay in your chair" operation, while providing an effective means of operating 160 meters with a relatively small antenna.

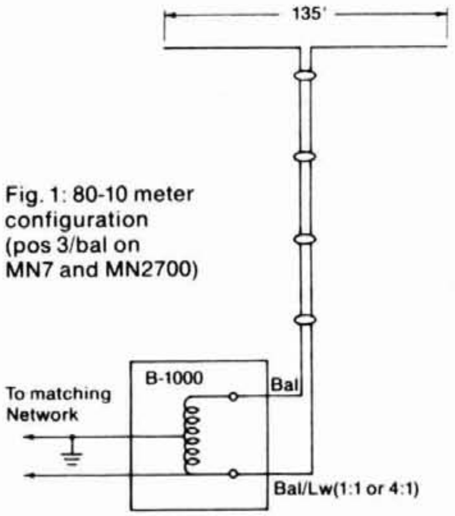


Fig. 1: 80-10 meter configuration (pos 3/bal on MN7 and MN2700)

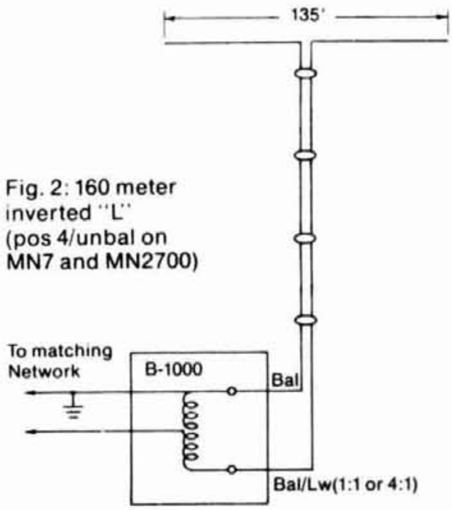


Fig. 2: 160 meter inverted "L" (pos 4/unbal on MN7 and MN2700)

short circuits

matchbox plus two

In the pictorial diagram of the switches in the July issue of *ham radio* (**fig. 2**, page 46), the wire from the balanced line terminal to the front deck ground terminal should be removed. The switch contact should remain grounded.

memory keyer

The schematic and PC layout of the deluxe memory keyer (**figs. 2** and **6**) in the April issue should show the three display-driver counter ICs (U5D, U6D, and U7D) as 7490s, *not* 7493s.

biquad bandpass filter

Author NØDE has written to point out that the two values of C in the biquad bandpass filter (June issue, **fig. 1**, page 70) should be equal for **eq. 1** to be correct. In **fig. 1** both values of C should be 0.1 μF. Also, placement of the left-hand IC on the PC board (**fig. 2**) is reversed; the notch should be oriented toward the lower edge of the circuit board.

rotator starting capacitors

Dear HR:

In "Ham Notebook" in *ham radio* magazine for June, 1979, W1JR reports on a common Ham-M rotator problem involving the electrolytic starting or phasing capacitor. The capacitor is in an ac circuit and is made up of two electrolytics connected in series, back to back in polarity. Each ac cycle places the full reverse potential across each section, in turn, which leads to eventual failure. At the suggestion of my friend Tony Abate, I connected two diodes across a pair of 120 μF electrolytic capacitors as shown below. In this circuit the individual capacitors are not subjected to a reverse polarity. I have used this arrangement for over seven years without any problems.

I. L. McNally, K6WX
Sun City, California

TIMETRAC

By Comus

NEW!

FIRST TIME OFFER

NOT A KIT

the microcomputer controlled appointment clock

Regular Price $79.95

Introductory Offer By Hal-Tronix only $69.95

RELAX AND LET TIMETRAC DO YOUR REMEMBERING.

TIMETRAC helps manage your busy schedule, increases your efficiency. Every home and office needs TIMETRAC.

TIMETRAC — THE CLOCK THAT REMEMBERS

This is the exciting, all new time-minder that combines space-age technology with everyday practicality. It remembers and reminds you of everything that you might forget. TimeTrac combines smart, modern design with precision and performance. Its vacuum fluorescent display provides readability from a distance (the largest display on the market today). You control the display brightness with a dimmer switch.

A NEW SOLUTION FOR SOME OLD PROBLEMS

Your spouse will never be upset with you for missing a birthday.

Your business associates will be pleased when you're never late for appointments.

Your doctor will be confident that you are taking your medication at the time prescribed.

FOR THE BUSY EXECUTIVE

Controls length of business meetings. Reminds you 10 minutes ahead of time to prepare for meeting and gives you time to clear your desk. Reminder of wife's birthday. Reminder to catch plane for important business trip.

FOR THE HOMEMAKER

Reminder to take meat out of freezer for dinner. Kitchen timer. Reminder of tennis dates and hair dresser appointments.

FOR THE MOTHER

Time children's phone calls, homework, music practice. Wake children for school.

FOR THE SENIOR CITIZEN

Medication reminder. Reminder of grandchildren's birthdays, doctor appointments. Easy-to-read large display. A wonderful gift for Mom and Dad.

FOR THE STUDENT

Timer for chemistry lab, bio lab. Timer for solving problems or preparation for exams.

FOR THE GOURMET COOK

Alarms to tell you when to start next step in meal preparation. By programming the timer alarm, you'll know just when each course of an elaborate meal must be prepared so everything will be ready at the same time. Helps you keep track of recipe timing.

FOR THE SALESMAN

Stores up to 30 future appointments — easy to see at the touch of a key when next appointment is scheduled.

FOR THE PHOTOGRAPHER

Timer for photographic development chain. Can insert red digital display filter to avoid damaging film.

FOR THE ATTORNEY

Records client's time charges, meetings, phone calls, research. Timer with built-in pause capability provides accurate way of timing speech presentations.

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The secretary's best friend. Remembers to remind the boss of key appointments. Times length of phone calls.

TIMETRAC FEATURES

- Sleek modern styling to complement any home or office decor.
- Tells the time.
- Tells the date and year.
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- Daily appointment sets appointments for the next 23 hours, 59 minutes.
- Future appointments up to one year.
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- Lithium power cell to retain memory during power outage.
- Appointments entered out of chronological order will be stored in chronological order.
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KDK 2015R KDK 2016A
MIDLAND 13-510 13-513 CLEGG FM-28
YAESU FT227R ICOM IC22S KENWOOD TR7400A

- AED continues to expand its line of quality scanners. Each of the above scanners are custom designed for their respective rigs.
- All scanners install completely inside rig. No obtrusive external connections.
- All are easy to assemble and come complete with detailed instruction manual.
- Scanned frequency displayed on digital readout (except IC22S).
- In the scanner OFF mode the rig operates normally. In the scanner ON mode the scanner locks on an occupied frequency, pauses for a preset time (about 5 secs.) and then resumes scanning.
- This gives you the ability to eavesdrop all over the band without lifting a finger. When you hear something interesting, you flip the switch to the LOCK mode and the rig is ready to transmit.

AED SCANNER SPECIFICATIONS

	KDK 2015R	KDK 2016A	KENWOOD TR7400A	YAESU FT227R	MIDLAND 13-510 13-513	CLEGG FM-28	ICOM IC22S
SCAN RATE	Adjustable 100 kHz/sec-1 MHz/sec		50 kHz/sec	200 kHz/sec	100 kHz/sec	100 kHz/sec	100 kHz/sec
SWEEP WIDTH	144-148 or only the MHz segment you select on MHz switch	142-149 995	complete band or MHz you want	adjustable eg 146-148 144-146 146-147	scans the MHz seg selected by the MHz switch	same as Midland	145 35-147 99
SCAN CONTROLS	2 mini toggle switches mounted on rig - LOCK switch may be mounted on mic		2 mini toggle switches mounted on rig	1 mini toggle switch mounted on mic or rig	2 mini toggle switches mounted on rig	same as Midland	1 mini toggle switch mounted on mic or rig
Price per kit	$39 95		$39 95	$34 95	$39 95	$39 95	$34 95
Price pre-assembled	$59 95		$59 95	$54 95	$59 95	$59 95	$54 95

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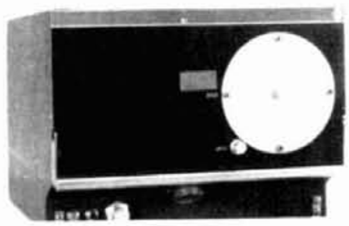

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1108	Antenna Adaptor Cord	$ 9.95
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1111	Carrying Case (Vinyl)	$ 9.95
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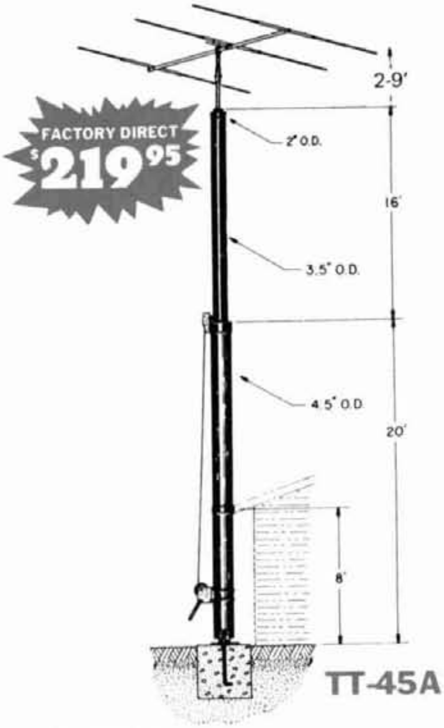
New, Improved Wilson Towers

Hinged Base Plate - Concrete Pad, Heavy Duty Winch

Mounting the House Bracket

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



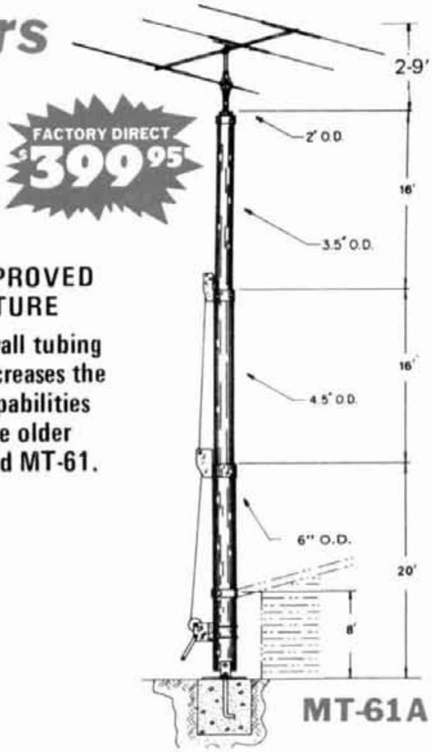
FEATURES:

- Maximum Height 45' (will handle 10 sq. ft. at 38') @ 50 mph
- 800 lb. winch
- Totally freestanding with proper base
- Total Weight, 189 lbs.

The TT-45A is a freestanding tower, ideal for installations where guys cannot be used. If the tower is not being supported against the house, the proper base fixture accessory must be selected.

NEW IMPROVED FEATURE

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



FEATURES:

- Is freestanding with use of proper base
- Maximum Height is 61' (will handle 10 sq. ft. at 53') @ 50 mph
- 1200 lb. brake winch
- 4200 lb. raising cable
- Total Weight, 350 lbs.

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

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

GENERAL FEATURES

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

TILT-OVER BASES FOR TOWERS

FIXED BASE

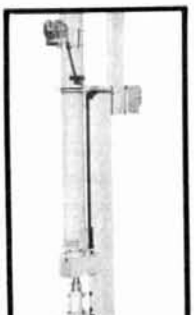
The FB Series was designed to provide an economical method of moving the tower away from the house. It will support the tower in a completely free-standing vertical position, while also having the capabilities of tilting the tower over to provide an easy access to the antenna. The rotor mounts at the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

FB-45A . . . $ 79.95
FB-61A . . . 109.95

ROTATING BASE

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

RB-45A . . . $119.95
RB-61A . . . 179.95



Tilting the tower over is a one-man task with the Wilson bases.
(Shown above is the RB-61A.)

WSI **WILSON SYSTEMS, INC.**

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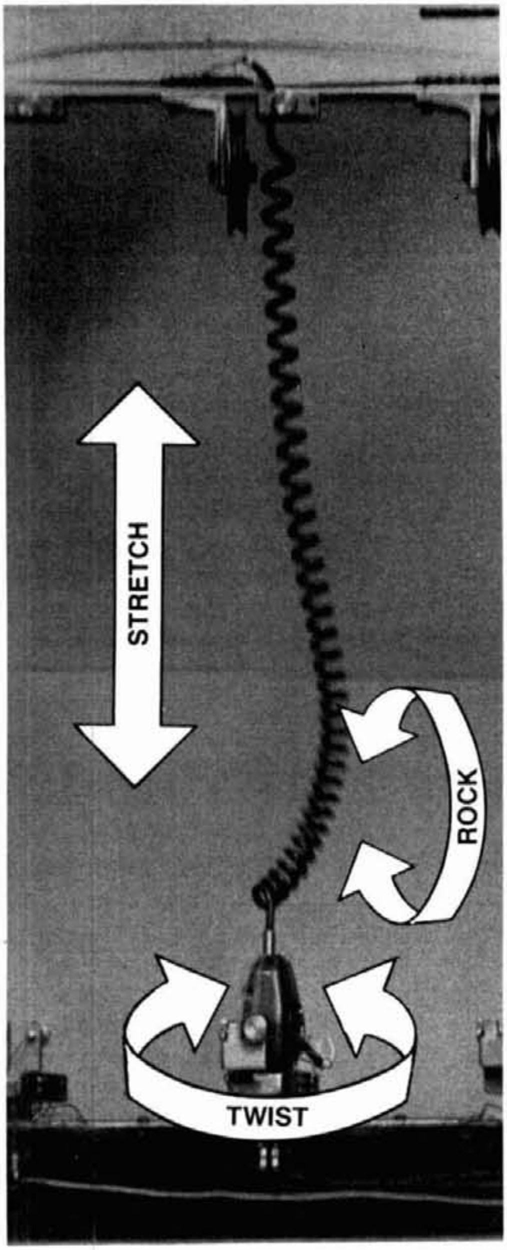
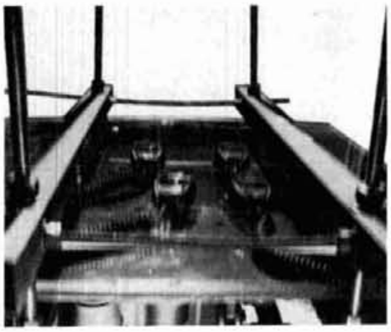


Originally designed for battlefield ruggedness, the microphone elements in Shure mobile and communications microphones offer unequaled reliability. Our quality control engineers anticipate the worst possible field conditions. These microphones have been subjected to the most rigorous tests in the industry, including six-foot drops onto hard floors; violent vibration tests; temperature variation tests ranging from a bitter -54°F. to a searing 185°F.; and 100% humidity tests. We've even dragged them behind automobiles on open roads and subjected them to a battery of corrosion tests. And yes, they really work after all that!

Exclusive Three-Way Flex Tester

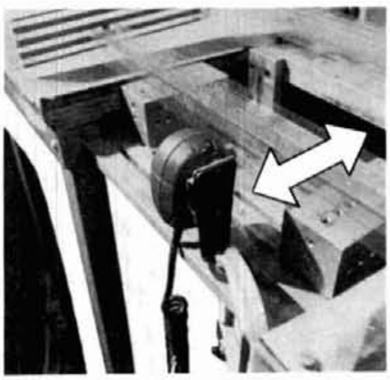
Shure knows that the single most common cause of microphone malfunction is failure of the cord. An exclusive Shure-designed story-and-a-half tall microphone cord tester dishes out more abuse than the average microphone gets in a lifetime.

Stretch, rock, and twist: first, the cord tester stretches the microphone to the full length of the cord. Then it simultaneously rocks the microphone 270° at the end of the cord while it gives the microphone a violent 90° twist in two directions. And this goes on day after day!



3-D Shake Tester

A microphone that fails spells disaster for a mobile communications system. Every Shure microphone is designed to withstand hours in our brutal 3D Shake Tester — simulating years of driving over rough, bumpy roads.



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TELEMETRY COMMUNICATIONS & INSTRUMENTATION, CORP.

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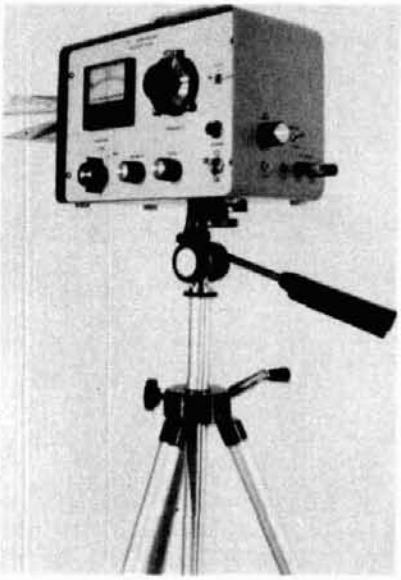

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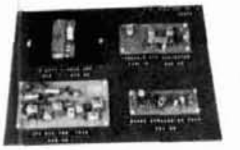

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RTTY AFSK Modulator PC board. See Feb. 79 Ham Radio. Drilled $5.00 F. E. Hinkle, 12412 Mossy Bark, Austin, TX 78750.

EZ deals are the best! Try me and see for Yaesu, Drake, KLM, Swan, Cushcraft, DenTron, VHF Eng., ICOM, CDE, Hustler, Wilson and more. Call, see or write WØEZ, Bob Smith Electronics, RFD 3, Hwy 169 & 7, Fort Dodge, IA 50501. (515) 576-3886.

WANTED: Motorola KXN 1024 and KXN 1052 channel elements. WA6COA, 4 Ajax Place, Berkeley, CA 94708.

THE MEASUREMENT SHOP has used/reconditioned test equipment at sensible prices; catalog. 2 West 22nd St., Baltimore, MD 21218.

MANUALS for most ham gear 1937/1970. Send 25¢ for "Manual Catalog." HI, Inc., Box H864, Council Bluffs, Iowa 51502.

More Details? CHECK — OFF Page 126

PORTA PAK — Make your FM mobile a self-contained portable. Models in stock for most popular makes. 4.5 amp hr model $80.00. 9 amp hr $103.00. Charger included, shipping extra. P.O. Box 67, Somers, WI 53171.

TEST EQUIPMENT CATALOG listing used Tektronix, HP and GR equipment at bargain prices. PTI, Box 8699, White Bear Lake, MN 55110. Price $1.00 refundable with first order.

SATELLITE TELEVISION — movies, sports, etc. Build or buy your own Earth Station. Send $3.00 for information Satellite Television, Box 140, Oxford, N.Y. 13830.

VERY in-ter-est-ing! Next 3 issues $1. "The Ham Trader", Wheaton, IL 60187.

Coming Events

RADIO EXPO '79, September 15 and 16, Lake County Fairgrounds, Routes 120 and 45, Grays Lake, Illinois. Manufacturers' displays, flea market, seminars, ladies' programs. Advance tickets $2.00. Write EXPO, P.O. Box 305, Maywood, IL 60153. Exhibitors inquiries: EXPO Hotline (312) 345-2525.

CINCINNATI HAMFEST: 43rd Annual — Sunday, September 16, 1979 at Strickers Grove on State Route 128, one mile west of Ross (Venice) Ohio. Exhibits, prizes, food and refreshments available, Flea Market (radio related products only), Music, Good Fellowship, Hidden Transmitter Hunt and Sensational Airshow. Admission and Registration $4.00. For further information: Lillian Abbott, K8CKI, 1424 Main St., Cincinnati, Ohio 45210.

GEORGIA: The Northwest Georgia Amateur Radio Club's Annual Hamfest, October 7, Coosa Valley Fairgrounds, Rome. Gates open 9:00 AM. Talk-in 146.34/94 and 146.085/685. For info: WB4AEG, Box 274, Adairsville, GA 30103.

SOUTH DAKOTA: 1979 ARRL Dakota Division Convention, October 5, 6, 7. Sioux Falls Airport Ramada Inn, off exit 81 on I-29. Forums, ladies' programs, ARRL forum, exhibits, banquet. Advance registration prize: DenTron GLA-1000 amplifier. Grand Prize: Kenwood TS-820S. Second Grand Prize: Wilson SY-1 antenna and WR-500 rotor. Registration: $15.00 ($16.00 after Sept. 1). Convention only: $6.00 ($7.00 after Sept. 1). Talk-in on 146.16/76. For further information: Sioux Falls Amateur Radio Club, Box 91, Sioux Falls, SD 57101.

INDIANA: The Marshall County Amateur Radio Club's 4th Annual Swap and Shop, October 21, 8 AM (1300Z), National Guard Armory, Plymouth. Electronics items only. Admission: $2.00 door. $3.00 per 6-ft. table. Main door prize: Cash plus hourly drawings. Advance tickets $1.50. Table/space/ticket requests: MCARC, P.O. Box 151, Plymouth, IN 46563 or Paul DeVos, 109 Maple Avenue, North Liberty, IN 46554.

CALIFORNIA: Third Annual Smoked Hamfest, San Antonio Mission (near King City) Sunday, September 30. The usual Barbeque, good fun, flea market, prizes, ? dancing girls ? Commemorating the Hams who worked the Marble Cone fire, 1977, plus all their helpers plus all other Hams who enjoy a good time. Reply to: WB6IZF, 51280 Pine Canyon Road, King City, CA 93930. (408) 385-6164.

MASSACHUSETTS: The Hampden County Radio Association's Annual Auction, Friday, October 5, Feeding Hills Congregational Church, intersection of routes 57 and 187, Feeding Hills (west of Springfield). Doors open 7 PM, auction starts at 8 PM. Club takes 10%. For more info: Jeffrey J. Duquette, K1BE at (413) 569-6739.

GEORGIA: The Colquitt County Ham Radio Society will be operating club station WD4KOW from the site of the second annual Sunbelt Agricultural Exposition, October 9, 10, 11, 0900 to 1600 EDST each day. Operations will be mostly on 40 and 20 meters around 7.250 and 14.300 MHz with some operations on other HF bands. Members will listen for visiting Hams on local repeater 146.19/79. Special QSL cards will be available for those making a contact.

KENTUCKY: The ninth annual Greater Louisville Hamfest and State ARRL convention, September 29 and 30, West Hall of Kentucky Fair and Exposition Center, Louisville. Indoor, air-conditioned exhibitors area and flea market. Meetings and forums. Ladies' programs. FCC exams September 29. Banquet Saturday night. Admission: Adults $3.00 advance — $3.50 at door. Banquet $9.75. For info: The Greater Louisville Hamfest, P.O. Box 34444, Louisville, KY 40232. (502) 634-0619.

SPECIAL EVENT: Miss America Pageant. Station K2BR will be operating from the Miss America Pageant Headquarters in Atlantic City, September 1-8. Sponsored by Southern Counties Amateur Radio Association. Approximate frequencies: CW: 3560, 7060, 14060, 21060. Novice: 3730, 7130, 21130. Phone: 3935, 7235, 14280, 21380. QSL to K2BR, 591 White Horse Pike, Egg Harbor, N.J. 08215. (SASE please). Traffic to and from the Miss America Contestants will be accepted.

THE LONDON AMATEUR RADIO CLUB will hold its 2nd Annual Swap & Shop on October 28 from 8:00 until 4:00 at Lord Dorchester High School in Dorchester just off the 401. Lots of displays, prizes, etc., with a talk-in on 78/18. Tables are $2.00 and admission is $2.00. For more info write VE3CSK, R.R. #1, Ailsa Craig, On. N0M 1A0.

NEW YORK: The Hall of Science Amateur Radio Club's 2nd annual indoor/outdoor, rain/shine Electronic Hamfest, September 16, 9 AM to 4 PM, Municipal Parking Lot, 80-25 — 126th St., Queens. Free parking. Sellers $2.00. Buyers $1.00. Refreshments. Free prizes. Talk-in .52-.52 & 96-36.

OHIO: The Findlay Radio Club's 37th Annual Hamfest, Sunday, September 9, Riverside Park — 45 miles south of Toledo — Findlay. Admission and prize tickets: $1.50 advance, $2.00 door. 1st Prize Kenwood TS-120S. For tickets and further info: SASE to Clark Foltz, W8UN, 122 W. Hobart Avenue, Findlay, OH 45840.

PENNSYLVANIA: Pack Rat Hamarama '79 weekend — Third Annual Mid-Atlantic States VHF Conference, Saturday, October 6, Warrington Motor Lodge, Route 611, Warrington. Advance registration $3.00 ($4.00 at door) includes admission to Flea Market at Bucks Co. Drive-In Theater, Route 611, Warrington, Sunday, October 7. 8 AM to 4 PM. Flea Market registration: $2.00, tailgating $2.00/space (own table). Talk-in via W3CCX/3 on 146.52 MHz. Information on both events: Ron Whitsel, WA3AXV, Chairman, P.O. Box 353, Southampton, PA 18966.

PENNSYLVANIA: The Butler County Amateur Radio Association's "Ye Olde-Fashioned Hamfest", September 9, 10 AM to 4 PM, Butler Co. Farm Show Grounds, next to Roe Airport on Rt. 68, Butler. Admission: $1.00. Free parking, free outdoor flea market. Under 12 free. Overnight camping. Refreshments. Mobile check-ins on .52 simplex and 147.90/.30 (W3UDX). Fly-ins welcomed. 80 + 100 AV gas available. YL, mobile and fly-in prizes. Six main prizes plus others drawn every 15 minutes. For info: Fred Young, WB3HGC, 195 Robbie Way, Portersville, PA 16051. (412) 368-3386.

NORTHEASTERN STATES 160 METER AMATEUR RADIO ASSOCIATION Annual election and banquet on Sunday, October 14, 1979, at Kozel's Restaurant, Route 9H, West Ghent, New York. Flea Market in Rear Parking Lot at 1:00 PM. Roast Beef Dinner at 5:00 PM. All HAMS and XYLS are welcome. For reservations and details contact WA5IOD — Sec/Treas. William Derby, 14 Plain Street, Medfield, MA 02052.

KANSAS: Sand Hills A.R.C. Swapfest, Sunday, September 16, 1979, 9 AM to 5 PM at the Four-H building on the Garden City Fairgrounds. Talk-in on 3935 kHz (75 meter SSB) 146.31/.91 and 146.52 (simplex) FM. For information, write S.A.R.C., P.O. Box 811, Garden City, Kansas 67846.

TENNESSEE: Memphis Hamfest and ARRL Tennessee State Convention, Saturday and Sunday, October 13th and 14th, 1979 at Youth Building, Mid-South Fairgrounds. Exhibits, forums, FCC exams at 8 AM sharp on Saturday (bring completed Form 610 and copy of license), flea market ($3/day — tables free), dealers and manufacturers' displays, ladies' activities, and Saturday night party. Hourly prizes. Admission: $3 adults; children 14 and under, free. Trailer hookups available on site. Talk-in 146.34/.94 and 146.25/.85. For information including special motel accommodations, write Memphis Hamfest, P.O. Box 3845, Memphis, TN 38103.

THE SANGAMON VALLEY RADIO CLUB of Springfield, Illinois holds its Fourth Annual Hamfest on Sunday, September 23rd. Location — Sangamon County Fairgrounds in New Berlin. 16 miles west of Springfield. Indoor display area and covered pavilion. Hear Randy Rowe NØTG talk on the Navassa DX-pedition! Various exhibits, kids activities and food available. Overnite camping. First Prize: Atlas RX110/TX110 with power supply. Tickets: $1.50 advance, $2.00 gate. Information: John Sams, WA9KRL, S.V.R.C., 1025 South Sixth, Springfield, IL 62703.

NEW JERSEY: South Jersey Radio Association's Hamfest, September 9, Pennsauken Senior High School, Hylton Road at routes 130 and 70, Pennsauken. Entry tickets: $2.00; tailgate: $3.00; inside tables: $5.00. For info: South Jersey Radio Association, P.O. Box 1734, Cherry Hill, NJ 08034.

More Details? CHECK — OFF Page 126

YLRL ANNIVERSARY QSO PARTY: Phone: Wednesday, October 10, 1979 at 1800 UTC; ends Thursday, October 11, 1979 at 1800 UTC. CW: Thursday, November 1, 1979 at 1800 UTC; ends Friday, November 2, 1979 at 1800 UTC. All licensed women operators throughout the world are invited to participate. Call CQ YL on any band 160-10 meters, exchange call, QSO number, RS(T), ARRL section or country. Log time, band, date and power. More information from Margaret Williams, KI4W, 965 Redwood Circle, Virginia Beach, VA 23462.

HAMFESTERS RADIO CLUB WAHM AWARD: Illinois Amateurs to Illinois Amateurs must reach ten HAMFESTER members; outside Illinois, but inside USA, you must reach five HAMFESTER members in Illinois; outside USA, you must contact three HAMFESTERS in Illinois. Submit a list of contacts containing call, frequency, date, and handle of HAMFESTER member contacted to: Hamfesters R.C., Box 42792, Chicago, IL 60642.

NEW YORK: Hamburg Ham-O-Rama, September 14th and 15th, 1979 at the Hamburg Fairgrounds. Technical meetings, displays, FCC examinations, inside-outside flea markets, contests, CD activities and more. Fun for the whole family. Contact Jim Ciurczak, 10404 Cayuga Drive, Niagara Falls, N.Y.; Carl Leisner, 138 Louis Street, Cheektowaga, N.Y., or Ron Chmiel, 278 Dan-Troy, Amherst, N.Y. for more information.

NORTH CAROLINA QSO PARTY: 1900Z December 1, 1979 through 0100Z December 3, 1979. Suggested frequencies +/− 10 kHz are: CW: 3560, 7060, 14060, 21060, 28060; Novice: 3720, 7120, 21120, 28120; SSB: 3900, 7270, 14290, 21390, 28590. Out-of-state stations transmit RS(T) and state, province or country. North Carolina stations send RS(T) and county. More information from Almance ARC, Inc., 2822 Westchester Drive, Burlington, N.C. 27215.

DELTA QSO PARTY: sponsored by the ARRL DELTA DIVISION, 1800Z, September 29th to 2400Z September 30th. No time or power restrictions. Amateurs outside the Delta Division will try to contact as many Amateurs inside the Delta Division as possible, while Division hams try to contact as many others as possible both inside and outside the division. Exchange QSO number, RS(T), and QTH (ARRL section for non-Delta hams, and county and state for Delta Division hams. Suggested frequencies are: CW: 3550, 7050, 14050, 21050, 28050; SSB: 3990, 7290, 14290, 21390, 28590; Novice: 3725, 7125, 21125, 28125 kHz. More information from Malcolm P. Keown, W5XX, 213 Moonmist, Vicksburg, MS 39180.

COLLEGE RADIO SCRIMMAGE: open to all Radio Amateurs, 1900Z, September 30th, through 0100Z, October 1st. Entry classes alumni and club station, SSB only. Exchange name of college, junior college or university you last attended, and the last two digits of the year you graduated, will graduate, or last attended. Club stations substitute the words "Amateur Radio Club" for number. Work stations once per band. Suggested frequencies: 1815, 3895, 7230, 14280, 21355, 28560 kHz. For more info write Penn State A.R.C., K3CR, 202 Engineering Unit E, University Park, PA 16802.

ARROWHEAD RADIO AMATEURS QSO PARTY: open to all Amateurs. Amateurs within 50 air miles of Duluth/Superior are considered Arrowhead Amateurs, and may work anyone. Amateurs outside this area may work only Arrowhead Amateurs for contact points. Call CQ ARAC-50. Frequencies: CW: 3535, 3725, 7035, 7125, 14035, 21035, 21125, 28035, and 28125 kHz. Phone: 3980, 7280, 14280, 21360, and 28560 kHz. Times: 1500 UTC, October 6, 1979 to 0300 UTC, October 7, 1979; and 1500 UTC to 2359 UTC, October 7, 1979. Arrowhead Amateurs exchange RS(T), county and state; all others send RS(T), ARRL section or country. More info from ARAC-50, 123 East First Street, Duluth, MN 55802.

IOWA: 1979 Midwest Division ARRL Convention and CVARC Hamfest, October 19, 20, 21, Five Seasons Center, Cedar Rapids. Grand Prize: Deluxe HF Transceiver, TH6-DXX Antenna, Ham III Rotor, 60 ft. Rohn 25G Tower. Thousands of dollars in door prizes. Forums, flea market, tours, ladies' programs, nearby theaters, entertainment, shopping, camping. Saturday evening banquet. Tickets: $4.00 advance; $5.00 at door. Talk-in on 146.16/76, 146.52, 223.34/94. Write: Convention Cedar Valley Amateur Radio Club, Box 994, Cedar Rapids, IA 52406.

NORTH CAROLINA: The Western Carolina Amateur Radio Society's Fourth Annual Autumnfest, Saturday, October 13, Asheville Civic Center, Asheville. Come enjoy the beautiful Smokey Mountain scenery. Displays, dealers, flea market. For info: Western Carolina ARS, P.O. Box 1488, Asheville, NC 28802.

OHIO: Fourth Annual Cleveland Hamfest, Sunday, September 23, Cuyahoga County Fairgrounds, Berea. For info: David L. Kersten, N8AUH, 2197 McKinley Avenue, Lakewood, OH 44107.

BLOSSOMLAND: Fall Swap Shop, October 7, Berrien County Youth Fairgrounds, Berrien Springs, Michigan. Large convenient facilities and refreshments. Tables restricted to radio and electronic items. Advance ticket donation $1.50. Tables $2.00. Write Charles White, 1940 Union Ave., Benton Harbor, MI 49022. Make checks payable to Blossomland ARA.

VIRGINIA: Fourth annual Tidewater Hamfest — Computer Show — Flea Market will be held in the Norfolk, VA Cultural and Convention Center SCOPE October 20 and 21, 1979. 60,000 square feet of airconditioned exhibit and Flea Market tailgating space are available. Doors open at 9:00 AM. ARRL meetings, DX, Traffic forums, plus a CW contest are scheduled. FCC Exams are planned for amateur upgrading Saturday 9-12 AM. A special feature will be a dinner cruise and banquet on the Spirit of Norfolk Cruiseship Saturday night. Advance registrations $2.50 (SASE), $3.50 at the door. Flea Market tailgate spaces $3/day. Cruise and banquet $16 per person, $30 per couple. Tickets and information — TRC P.O. Box 7101, Portsmouth, VA 23707.

JAMBOREE ON THE AIR: Scouts, former Scouts and interested hams, October 20-21, 1979. Stations are free to select times of operation. Hams may invite Scouts, Cub Scouts and Explorers to their shacks for a look at ham radio and to talk with Scouts in some 100 countries, wherever permitted by regulations. Phone (US): 3940, 7290, 14290, 21360, 28990, 50500; Phone (International): 3740, 7090, 14290, 21360, 28990. CW (US & Int'l): 3590, 7030, 14070, 21040, 28190, 50050; CW (Novice): 3750, 7125, 21140. Acorn and Oak participation certificate for hams and Scouts. More information from Harry A. Harchar, W2GND, Boy Scouts of America, North Brunswick, New Jersey 08902; Tel. (609) 448-4717.

PENNSYLVANIA: The RAE Annual Hamfest, September 23, 8 AM to 4 PM, Rainbow Gardens, Waldameer Beach Park. Admission: $3.00 includes chances on main prize, YL prizes, and half-hour drawing on door prizes. Flea market vendors $1.00 per car. Exhibits, refreshments. Talk-in on .34/.94 and .52 simplex. For info: Radio Association of Erie, Box 844, Erie, PA 16512.

MICHIGAN: Adrian Amateur Radio Club's 7th Annual Hamfest, Sunday, September 23, Lenawee County Fairgrounds, Adrian. Prizes: Main and hourly. Bingo and much more. Talk-in 146.31/91 and 146.52. For ticket, tables info: Adrian ARC, Inc., P.O. Box 26, Adrian, MI 49221. Telephone info: Bob/Sally Fay, Sword Enterprises — (517) 263-3597.

MICHIGAN: L'Anse Creuse ARD's 7th Annual Swap and Shop, September 16, L'Anse Creuse High School, Mt. Clemens, 0900-1500. Cash prizes — hourly drawings. $2.00 door, $1.00 advance. Talk-in 14769.09 and 146.52. For info or tickets SASE to WD8ITS, 3488 Ashley, Pontiac, MI 48055.

NEW YORK: Radio Amateurs of Greater Syracuse 15th Annual Hamfest, Saturday, October 13, 9:00 AM to 6:00 PM, New York State Fairgrounds, Arts and Home Center, Syracuse. Exhibitors booths, indoor/outdoor flea market, awards, films and ladies' programs. Tickets before October 1 at reduced rates. Under 12 free. Overnight and trailer parking available. Talk-in on 90/30 and 31/91. Refreshments. For tickets or info: RAGS, P.O. Box 88, Liverpool, N.Y. 13088.

GEORGIA: The Central Georgia ARC's First Annual Hamfest, September 30, 8:00 AM to 5:00 PM, City Recreation Center, Watson Blvd., Warner Robins. Indoor flea market. Talk-in 3.975 LSB, 146.25/85 and 146.52 simplex. Also Georgia SSB Assn. and Georgia CW Assn.'s annual meetings. For info: Bill Atkins, WD4ASB, 201 Avalon Drive, Warner Robins, GA 31093. (912) 923-3454 or check in on 3.975 MHz nightly at 2330 UTC.

PENNSYLVANIA: Uniontown Amateur Radio Club's 30th Annual Gabfest, Saturday, September 8, Club Grounds, Old Pittsburgh Road, Uniontown. Noon on. Refreshments, free coffee. Registration: $2.00. Prizes. Main Registration Prize: Tempo Syncon S1. Swap and Shop. For info: W3PIE, Uniontown Amateur Radio Club, Inc., 438 Braddock Avenue, Uniontown, PA 15401. (412) 438-9488.

SOUTH CAROLINA: York County Amateur Radio Society's 28th Annual Hamfest, Sunday, October 7. 0800 until —, Joslin Park, Rock Hill. Registration: $2.75 each or 2/$5.00 advance); $3.00 gate. Main prize: Yaesu 901-DM. Barbeque dinner available at park. Talk-in 146.43/147.03 and 146.52 direct. For info: P.O. Box 4141 CRS, Rock Hill, SC 29730.

CANADA: Radio Society of Ontario's Eleventh Annual Convention, October 12, 13, 14, Skyline Hotel, Ottawa. Friday evening buffet and dance. Saturday tech. sessions, demonstrations, forums, displays. Saturday night banquet and dance. Sunday indoor flea market. P.O. Box 5076, Stn. F, Ottawa, Ontario K2C 3H3.

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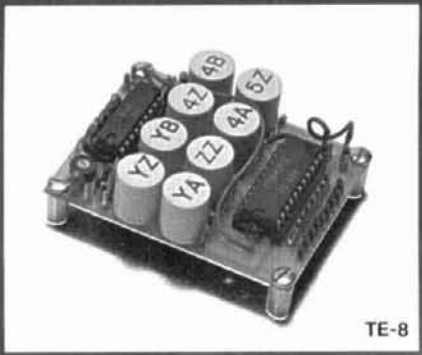
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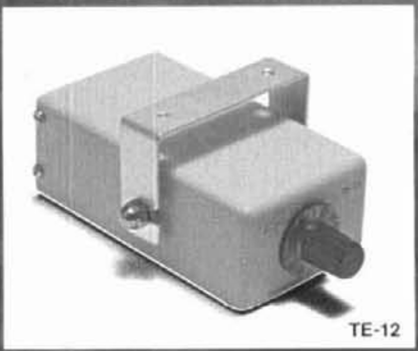
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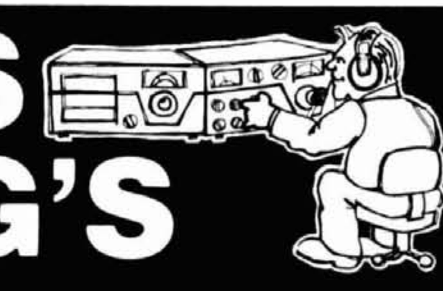

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Remember, you can Call Toll Free: **1-800-633-3410** in the U.S.A. or call **1-800-292-8668** in Alabama for our low price quote. Store hours: 9:00 AM til 5:30 PM, Monday thru Friday.

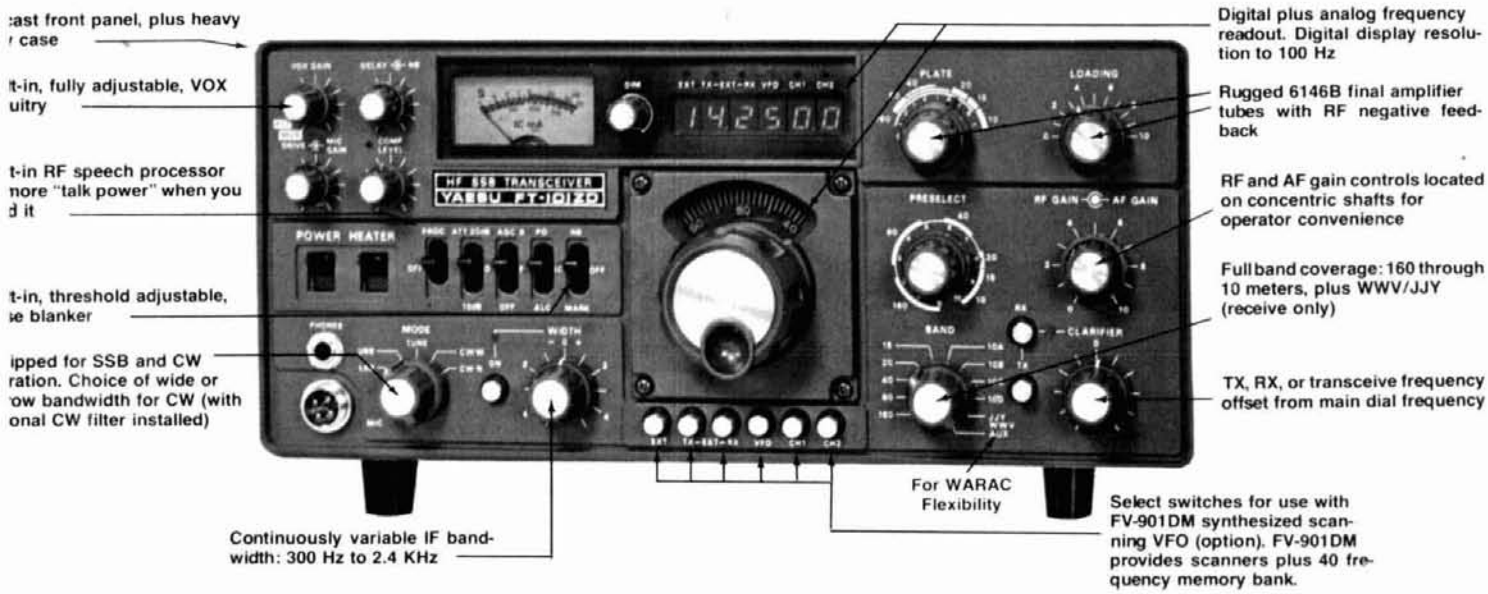
More Details? CHECK — OFF Page 126

ALL NEW

FT-101ZD

HIGH-PERFORMANCE HF TRANSCEIVER

Today's technology, backed by a proud tradition, is yours to enjoy in the all-new FT-101ZD transceiver from YAESU. A host of new features are teamed with the FT-101 heritage to bring you a top-dollar value. See your dealer today for a "hands on" demonstration of the performance-packed FT-101ZD.



SPECIFICATIONS

TRANSMITTER

PA Input Power:
180 watts DC
Carrier Suppression:
Better than 40 dB
Unwanted Sideband Suppression:
Better than 40 dB @ 1000 Hz, 14 MHz
Spurious Radiation:
Better than 40 dB below rated output
Third Order Distortion Products:
Better than −31 dB
Transmitter Frequency Response:
300-2700 Hz (−6 dB)
Stability:
Less than 300 Hz in first 30 minutes after 10 min. warmup; less than 100 Hz after 30 minutes over any 30 min. period
Negative Feedback: 6 dB @ 14 MHz
Antenna Output Impedance:
50-75 ohms, unbalanced

GENERAL

Frequency Coverage:
Amateur bands from 1.8-29.9 MHz, plus WWV/JJY (receive only)
Operating Modes:
LSB, USB, CW
Power Requirements:
100/110/117/200/220/234 volts AC, 50/60 Hz; 13.5 volts DC (with optional DC-DC converter)
Power Consumption:
AC 117V: 75 VA receive (65 VA HEATER OFF) 285 VA transmit; DC 13.5V: 5.5 amps receive (1.1 amps HEATER OFF), 21 amps transmit
Size:
345 (W) × 157 (H) × 326 (D) mm
Weight:
Approximately 15 kg.

COMPATIBLE WITH FT-901DM ACCESSORIES

RECEIVER

Sensitivity:
0.25 uV for S/N 10 dB
Selectivity:
2.4 KHz at 6 dB down, 4.0 KHz at 60 dB down (1.66 shape factor); Continuously variable between 300 and 2400 Hz (−6 dB); CW (with optional CW filter installed): 600 Hz at 6 dB down, 1.2 KHz at 60 dB down (2:1 shape factor)
Image Rejection:
Better than 60 dB (160-15 meters); Better than 50 dB (10 meters)
IF Rejection:
Better than 70 dB (160, 80, 20-10 m); Better than 60 dB (40 m)
Audio Output Impedance:
4-16 ohms
Audio Output Power:
3 watts @10% THD (into 4 ohms)

member arma amateur radio manufacturer's association

Price And Specifications Subject To Change Without Notice Or Obligation

YAESU **The radio.**

YAESU

379X

YAESU ELECTRONICS CORP., 15954 Downey Ave., Paramount, CA 90723 ● (213) 633-4007
YAESU ELECTRONICS Eastern Service Ctr., 9812 Princeton-Glendale Rd., Cincinnati, OH 45246

Heathkit SB-221 linear amplifier uses EIMAC 3-500Zs for efficiency, economy and performance.

Designed for rugged service.

The new desktop Heathkit SB-221 linear amplifier provides up to 2000 watts PEP input for SSB and 1000 watts input for CW service. Only 100 watts drive power is required to achieve these power levels.

Designed for rugged contest and traffic service, the SB-221 uses the highest grade components including two EIMAC 3-500Z high gain power triodes, well-known for their reliable, efficient performance. One thousand watts of plate dissipation is available from the two tubes, providing ample safety factor for long life service.

The designer's choice.

Top-notch equipment designers, such as Heathkit, choose EIMAC power tubes for commercial as well as amateur products. The 3-500Z power tube used in the SB-221 also serves in many commercial broadcast, FM and point-to-point radio systems where reliability and long life are paramount.

Make sure this fine EIMAC 3-500Z is in your equipment. For full details and a data sheet on the 3-500Z, write Varian, EIMAC Division, 301 Industrial Way, San Carlos, CA 94070. Or contact any of the more than 30 Varian Electron Device Group Sales Offices throughout the world.

varian