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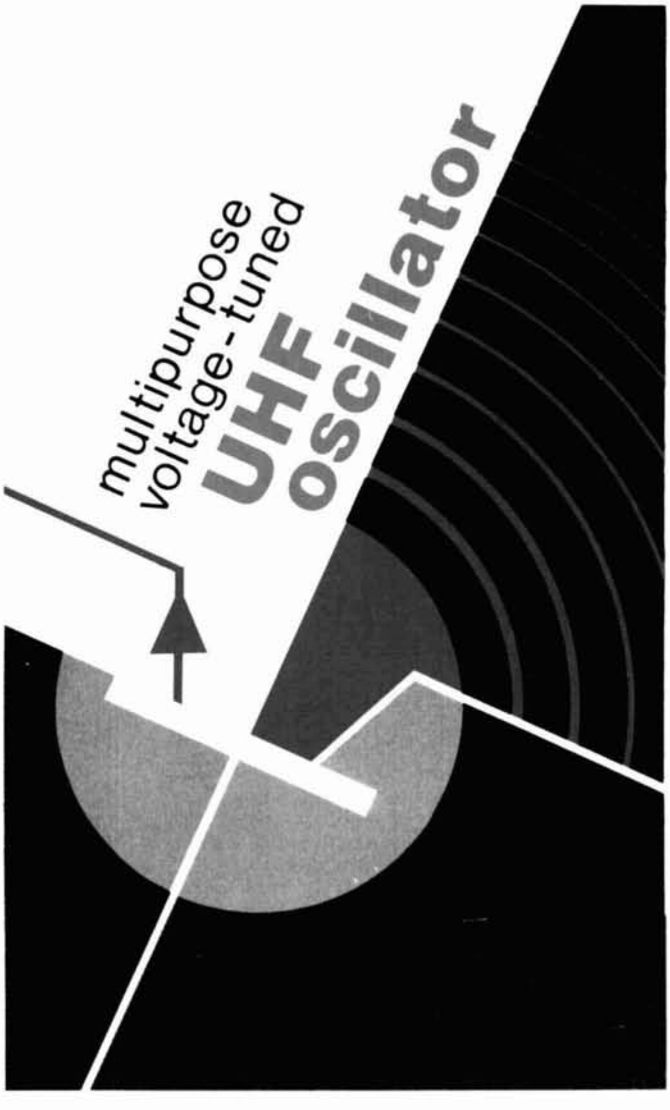
ham radio magazine

DECEMBER 1980



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The S-4...$349.00
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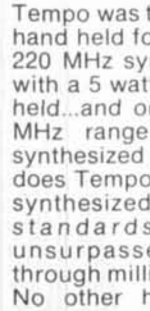
With 16 button touch tone pad...$419.00
S-40 matching 40 watt output
13.8 VDC power amplifier...$149.00



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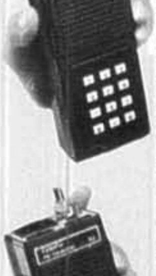
The first and most thoroughly field tested hand held synthesized radio available today. Many thousands are now in use and the letters of praise still pour in. The S-1 is the most simple radio to operate and is built to provide years of dependable service. Despite its light weight and small size it is built to withstand rough handling and hard use. Its heavy duty battery pack allows more operating time between charges and its new lower price makes it even more affordable.



Tempo S-5

Offers the same field proven reliability, features and specifications as the S-1 except that the S-5 provides a big 5 watt output (or 1 watt low power operation). They both have external microphone capability and can be operated with matching solid state power amplifiers (30 watt or 80 watt output). Allows your hand held to double as a powerful mobile or base radio.

S-30...$89.00* S-80...$149.00*
*For use with S-1 and S-5

Tempo S-2

With an S-2 in your car or pocket you can use 220 MHz repeaters throughout the U.S. It offers all the advanced engineering, premium quality components and features of the S-1 and S-5. The S-2 offers 1000 channels in an extremely lightweight but rugged case.

If you're not on 220 this is the perfect way to get started. With the addition of the S-20 Tempo solid state amplifier it becomes a powerful mobile or base station. If you have a 220 MHz station, the S-2 will add tremendous versatility.

Price...$349.00 (With touch tone pad installed...$399.00)
S-20...$89.00

Please note, as of Dec. 1, 1980 we will occupy our new world headquarters building with a new Los Angeles address and phone number.

Specifications:

Frequency Coverage: 440 to 449.995 MHz
Channel Spacing: 30 KHz minimum
Power Requirements: 9.6 VDC
Current Drain: 17 ma-standby 400 ma-transmit (1 amp high power)
Antenna Impedance: 50 ohms
Sensitivity: Better than .5 microvolts nominal for 20 db
Supplied Accessories: Rubber flex antenna 450 ma ni-cad battery pack, charger and earphone
RF output Power: Nominal 3 watts high or 1 watt low power
Repeater Offset: ± 5 MHz

Optional Accessories for all models

12 button touch tone pad (not installed): $39 • 16 button touch tone pad (not installed): $48 • Tone burst generator: $29.95 • CTCSS sub-audible tone control: $29.95 • Leather holster: $20 • Cigarette lighter plug mobile charging unit: $6

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Boost your signal. . . give it the range and clarity of a high powered base station. VHF (135 to 175 MHz)

Drive Power	Output	Model No.	Price
2W	130W	130A02	$209
10W	130W	130A10	$189
30W	130W	130A30	$199
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10W	80W	80A10	$149
30W	80W	80A30	$159
2W	50W	50A02	$129
2W	30W	30A02	$ 89

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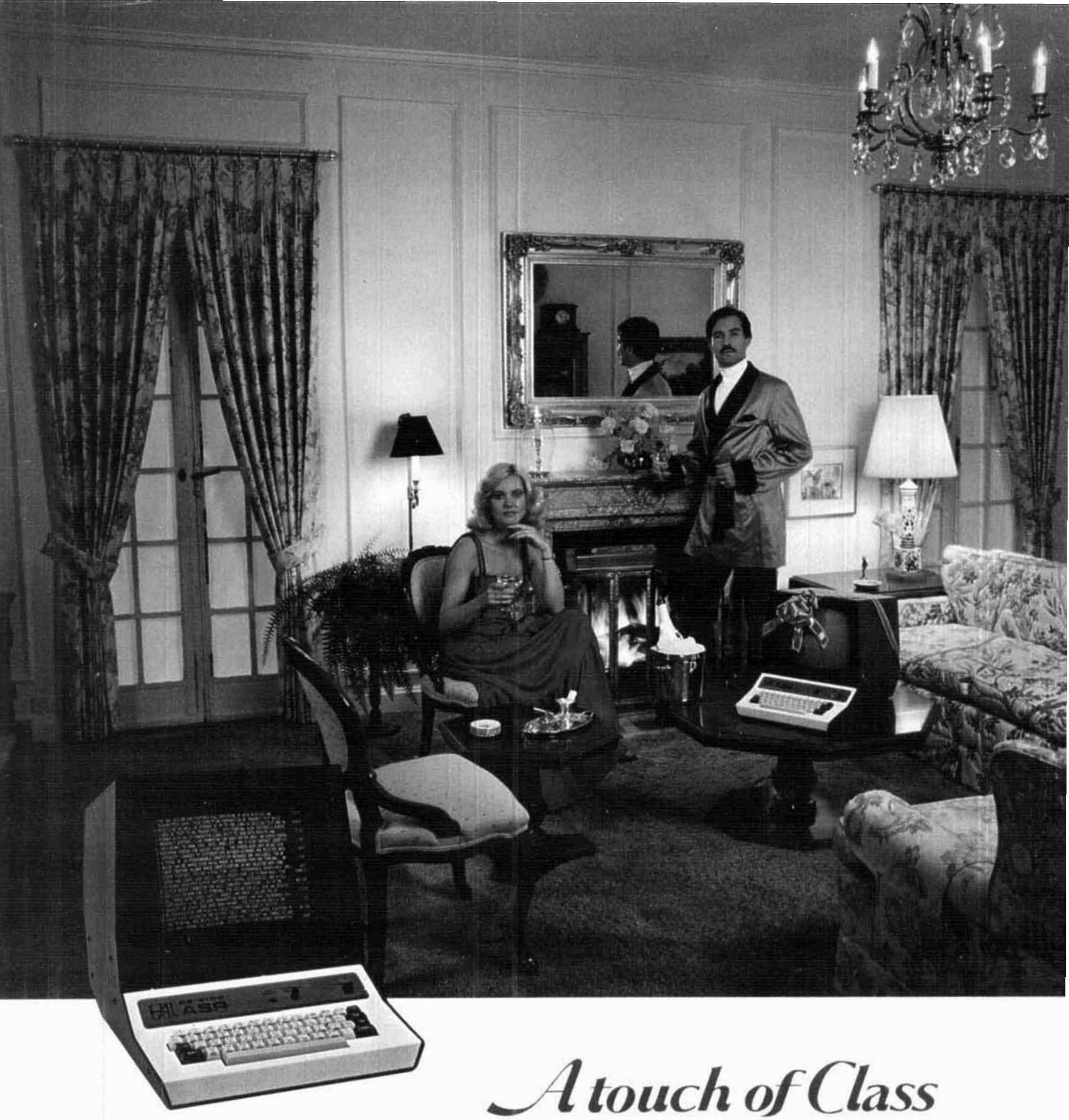

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Switch to "seconds" readout. For the times when you need the utmost accuracy.
Switch to ID timer. Alerts every 9 minutes after you tap the button (also functions as a snooze alarm).
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Switch to regular alarm. For skeds reminder or wake-up use (has alarm-on indicator).
Synchronize with WWV. Now you can adjust the MFJ clock to WWV accuracy.
Fast/Slow set buttons for easy setting of time and alarm.
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Operates on 110VAC, 60 Hz (50 Hz with simple modification). UL approved.
Handsome styling with rugged black plastic case with brushed aluminum top and front. Front has sloping surface for easy viewing. Cabinet measures 6x2x3".
Put this new improved MFJ digital clock to work in your shack.

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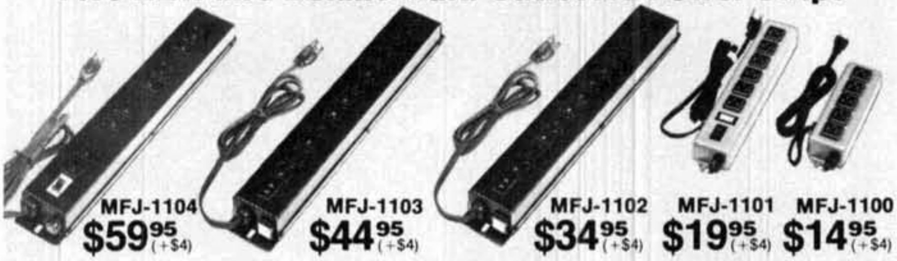

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Lighted switch shows circuits are "on."
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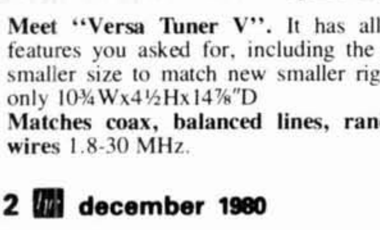
MFJ-1102, similar to 1103 but no RFI filter.
MFJ-1101: 6 sockets, all 3-prong type. Fuse protected, 15A, 125VAC. On-off switch. Lighted "On" indicator. 3-wire 6' power cord. Steel case, finished in gray hammertone, has mounting slots, measures 13⅛"L x2⅜"Wx1½"H.
MFJ-1100, similar to 1101 but 5 sockets, less switch, light, and is 8⅝"L.

NEW MFJ Compact 3 KW Antenna Tuner Has Roller Inductor



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$279.95 (+$10)

Meet "Versa Tuner V". It has all the features you asked for, including the new smaller size to match new smaller rigs — only 10¾Wx4½Hx14⅞"D
Matches coax, balanced lines, random wires 1.8-30 MHz.
3 KW PEP — the power rating you won't outgrow. (250 pf-6KV caps).
Roller inductor with a *3-digit turns counter* plus a *spinner knob* for precise inductance control to get that SWR down to minimum every time.
Built-in 300 watt, 50 ohm dummy load.
Built-in 4:1 ferrite balun.
Built-in lighted 2% meter reads *SWR plus forward and reflected power in 2 ranges* (200 & 2000 w).
6-position antenna switch (2 coax lines, through tuner or direct, random/balanced line or dummy load). SO-239 coax conn., ceramic feed-throughs, binding post ground.
Deluxe aluminum low-profile cabinet with sub chassis for RFI protection, black finish, black panel with raised letters; tilt bail; requires 12 VDC for meter light.

NEW MFJ VHF SWR/ Wattmeter/Field Strength Meters

MFJ-812 $29.95 (+$4) **MFJ-810 $24.95** (+$4)

New low cost VHF operating aids.
MFJ-812: Reads SWR from 14-170 MHz to keep you informed about antenna/ feedlines. SO-239 coax conn.
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MFJ-732 Puts more presence in SSB/ AM/FM voice communications, brings more signals out of the "mud."
Easy to use, just push up to 4 buttons.
10-pole (5-stage) circuit with *Chebyshev superfast roll-off* (up to 58 dB/octave). First button: On/Off-Bypass, response 300-3000 Hz; second: 500 Hz lower cutoff; third: 2200 Hz upper cutoff; fourth: 1500 Hz upper cutoff. *Built-in speaker,* 2 watt amplifier, LED, 9-18 VDC or 110VAC with optional AC adapter ($7.95+$2), 5x6x1⅝".

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More Details? CHECK—OFF Page 126

ham radio

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DECEMBER 1980

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contents

Observation & Opinion

It seems that a West Coast Amateur has decided to make some easy money by publishing material to aid prospective licensees in passing FCC Amateur examinations. His material is crafted so that mere memorization of answers to FCC exam questions practically guarantees a passing grade. His product apparently is derived from FCC exam materials. Such material is gleaned by a well-organized effort to collect questions verbatim from the various exams when they are administered by FCC representatives. Very often this has happened at Radio Amateur conclaves and conventions. We at *ham radio* magazine deplore such tactics. Amateur Radio has flourished because of its many established traditions. "In today, out tomorrow" publications, such as that referred to above, defeat the entire purpose of the Amateur Radio tradition, which has made our hobby one of the greatest in the nation for over 60 years.

Where do these questions and answers come from? From Radio Amateurs. The publisher in question solicits FCC test questions from those who have recently taken the exam, then publishes these questions along with the proper answers. Pretty neat. All one has to do is memorize the questions and answers, and the exam is a comparative cinch.

The publisher probably is making lots of money publishing the exam questions and answers without apparent legal sanctions (at least to date). But what about the long-range impact on the Amateur Radio Service and U.S. taxpayers at large? We lose.

An interesting sidelight is that the publisher justifies his action in the interest of "socially motivated" hams. His rationale for this rather obtuse reasoning is Part 97.1 (a) of the FCC rules and regulations, *Basis and Purpose*: "Recognition and enhancement of the value of the amateur service to the public as a voluntary noncommercial communication service, *particularly with respect to providing emergency communications*." (Italics mine.)

The publisher, however, conveniently overlooks Part 97.1 (b), which states: "Continuation and extension of the amateur's *proven ability to contribute to the advancement of the radio art*." (Italics mine.)

How can anyone in the Amateur Service comply with regulation 97.1 (b) if a license is obtained by memorizing answers to FCC questions? It is the purpose of this magazine to encourage Amateurs, by publishing articles on current technology, to "contribute to the advancement of the radio art." We believe that, for the most part, Amateurs who obtain their license using only the memorization technique are rarely in a position to contribute to part 97.1 (b) on a technical basis. There are exceptions, of course, but the method of preparing for exams to which we object seems to augur an increasingly less proficient operator in the midst of a rapidly increasing technical operating environment.

What can we Amateurs do to promote the technical integrity of Amateur Radio? Let's learn as much electronic theory as possible before taking the examination. It requires some effort, true, but when we pass the FCC exams based on knowledge rather than memorization we achieve a more significant accomplishment. After all, that's what ham radio is all about. Consider part three of "The Amateur's Code" by Paul Segal: "*The Amateur is Progressive* . . . He keeps his station abreast of science. It is well-built and efficient. His operating practice is above reproach."

ham radio continues to endorse this philosophy. The Amateur Radio Service cannot survive if licenses are obtained without due regard to technical knowledge: that is, passing FCC exams by learning the questions and answers by rote.

All prospective Amateurs should take a closer look at this problem. We licensed Amateurs who organize training classes and other tutorial endeavors have a special responsibility in this regard. Obtaining an Amateur license requires some effort. It is usually a difficult, time-consuming process. The successful license applicant will find the process rewarding for years to come.

What can the FCC do at this point to promote the technical integrity of Amateur Radio? We have some ideas, but we would like to hear from our readers on this point. Should the FCC look the other way while the abuse of Amateur exams continues? Should the FCC adopt an Amateur exam question series broadly similar to the FAA's several-hundred-question series for the Private Pilot license? More basically, why should newly updated exams be negated by one of us at the expense of us all? Consider this issue carefully, then discuss it among your Amateur Radio associates. Your views on the subject will be welcome at *ham radio*.

Alf Wilson, W6NIF
Editor

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IC-255A the most popular 2 meter FM rig on the air today.

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comments

RST feedback

Dear HR:

I read your comments on DL7DO's letter in "Observations and Comments," September, 1980, with some interest and a bit of confusion.

When I was running a '45 with 135 (not 90) volts on the plate, a signal report of S7 would have been somewhat meaningless: it did not gain significance until adoption of the RST system in the late thirties. The proper report prior to that would have been QSA (1-5), R (1-9). At the time of the adoption of the RST system most had converted to non-chirpy crystal control, and a-c on the plate supply brought an immediate citation from the newly formed FCC.

There is a definite need for accurate signal reporting, but if a report on tone is no longer needed (I for one disagree strongly with this reasoning), then let us not go the route of "inventing" a new system when the need is clearly covered in the international Q signals.

My personal feeling is that the RST system is performing admirably, with the exception of some contesters, and a change of the system would not change that. In other words, if it ain't broke, don't fix it!

Rue O'Neill, W0NN
St. Louis, Missouri

Dear HR:

I applaud the idea of junking the RST signal reporting system. But do we really need a new system? Why not simply make use of the existing QSA system which (with "copy" notes added) is as follows:

QSA 1 Scarcely perceptible — no copy
2 Weak — very little copy
3 Fairly good — partial copy
4 Good — almost full copy
5 Very good — full copy

Reports would simply be Q1, 2, 3, 4, or 5. Where the situation permits, an operator should do the other station the favor of reporting technical signal defects such as distortion, overdriving, VOX clipping, key clicks, poor tone, etc.

The difference between a signal received off the end of a dipole and the same signal received by a properly oriented high-gain beam is tremendous. The signal strength measured in the receiver depends almost entirely upon the character and orientation of the receiving antenna. A signal reported as S5 by a station with a mediocre antenna might easily be reported S9 or more by the station right next door having a superior antenna. So the popular "S" reports are all but meaningless anyhow!

J.W. Kennicott, W4OVO
Lexington, Tennessee

"circuit figure of merit"

Dear HR:

In reference to "Observations and Comments" in the September, 1980, issue of *ham radio*, I thought you might be interested in the "Circuit Figure of Merit" used by the State of New York in police two-way fm radio communications in the vhf and uhf ranges.

In writing specifications we usually ask the bidder to guarantee a Circuit Figure of Merit of 3 or better in a defined area of coverage from defined sites and with defined equipment parameters.

Byron H. Kretzman, W2JTP
Huntington, New York

The performance of a two-way radio circuit can be defined by grading the circuit in terms of a "Circuit Figure of Merit" using a scale of 1 to 5 under the following conditions:

circuit figure of merit	grade of circuit performance	voice frequency signal-to-noise ratio	typical receiver quieting
1	Unusable. Presence of speech barely discernible.	Below 8 dB	0 to 6 dB
2	Readable with difficulty. Requires frequent repeats. (Noncommercial)	8 to 16 dB	14 dB
3	Readable with only a few syllables missing. Requires occasional repeats. (Commercial)	14 to 22 dB	20 dB
4	Perfectly readable but with noticeable noise.	20 to 30 dB	25 dB
5	Perfectly readable; negligible noise.	Above 30 dB	Above 25 dB

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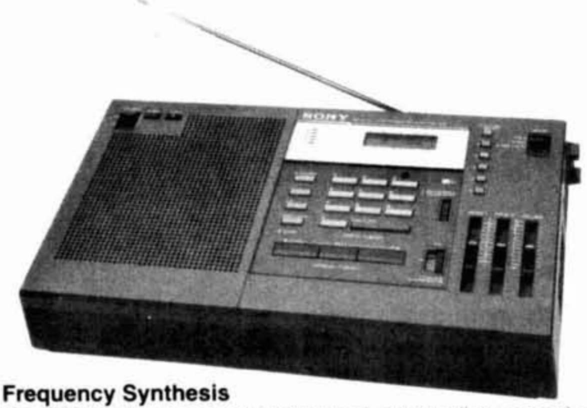

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Innovative design. Advanced technology. Digital key-touch tuning. The ICF-2001. It's a whole new breed of radio. A receiver that supplants the conventional multi-band concept, receiving a wide amplitude-modulated frequency range–shortwave, mediumwave and most longwave broadcasts. Plus FM, SSB and CW. Even more important, the 2001 replaces the ordinary tuning knob and dial with a direct-access tuning keyboard and a Liquid Crystal Display (LCD) for digital frequency readout. Which make the unit as easy to use as a pocket calculator. Instant, direct-access tuning modes and six memory-station presets assure maximum ease of use. And the quartz-crystal, frequency-synthesized circuitry behind them assures outstanding reception. Reception of local broadcasts and exciting news, music, sports, entertainment and information from around the world. You'll get the inside, local news stories from foreign countries ... exclusive coverage of world sports events ... plus everything from informal "ham" to marine communications. All at your fingertips.

Key-Touch Tuning

To tune a station manually, you simply punch in the station frequency numerals on the direct-access, digital tuning keyboard. Press the "Execute" key and the command is entered, the station is received and LCD readout confirms tuning. If you punch in an incorrect frequency by mistake, the ICF-2001 tells you to "Try Again" by flashing those words on the display. The instant, fingertip tuning provides total accuracy and convenience. And the LCD digital frequency display confirms the exact, drift-free signal reception.

Automatic Scanning

In auto-scan mode, the tuner can be set for continuous scanning of a given frequency range, which you set by means of upper and lower limit keys designated "L_1" and "L_2." You may want to scan an entire frequency range. For instance, the 76 to 108 MHz FM spectrum. If you want scanning to stop at any strong signal–one that reads "4" or "5" on the LED signal-strength indicator– switch on "Scan Auto Stop." For continuous scanning, leave the switch off, and just press the "Start/Stop" key to listen to a station or resume scanning.

Manual Tuning

Like the auto-scanning mode, manual tuning is useful for quick signal searching when you don't know particular station frequencies within a given range. You simply press the "Up" or "Down" key, and the tuner does the searching for you. And if you press the "Fast" key at the same time, the scanning rate increases for especially rapid station location. When you hear a broadcast you want to receive, just release the keys for instant reception, presssing the "Up" or "Down" key again if necessary for exact tuning.

Memory Presets

After you've tuned a station using punch-in, key-touch tuning or either scanning mode, you can enter it in the 2001's memory for instant, one-touch preset reception. Which means no retuning hard-to-find foreign broadcasts. Plus instant access to your favorite local stations for music and news. Six preset buttons allow up to six stations–in any wave range–to be memorized. And there's LCD digital readout of the memory buttons being used on each band. What's more, the upper and lower limit keys can be used as memory presets when they're not being used for scanning, allowing a total of eight frequencies to be memorized for instant, one-touch reception.

Frequency Synthesis

The 2001's direct-access tuning and outstanding reception quality are made possible by the unit's all-band quartz-crystal, PLL frequency synthesis. Instead of the conventional analog tuning system, with its variable tuning capacitor, the 2001 incorporates an LSI and a quartz-crystal reference oscillator. Which means that the local-oscillator frequencies used in superheterodyning are locked to the "synthesized" quartz reference frequencies. The result is the utmost in tuning stability, without a trace of tuning drift. In addition, dual-conversion superheterodyning for AM assures exceptionally clean, clear reception across the entire 150-to-29,999kHz spectrum.

Features

FM/AM/SSB/CW/wide spectrum coverage

Dual-conversion superheterodyne circuitry of AM assures high sensitivity and interference rejection

Quartz-crystal, phase-locked-loop frequency synthesis for all bands assures the utmost tuning stability, without a trace of tuning drift

Direct-access, digital tuning keyboard and LCD digital frequency readout for quick, key-touch station selection–maximum accuracy and ease of use

Manual tuning and automatic scanning for effortless signal searching, easy DXing

6-station presets, plus 2 auxiliary presets, for instant reception of memorized stations on any band–plus LCD memory indication.

5-step LED signal-strength indicator

Local/Normal/DX sensitivity selector for AM

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presstop

AN IMPORTANT ANTENNA VICTORY has not only restored the right of a Placentia, California, Amateur to use the antenna system of his choice, but has also reimbursed him his attorney's fees for defending that right. W6QOL, represented by attorney K6JAN, won his decision by taking the offensive and suing the city of Placentia in federal court for violating his civil rights by passing legislation aimed at his installation.

W6QOL's Tower, A 71-foot Crankup with several beams on it, had been constructed in 1977 with the approval of the city's planning commission, but prodding by an unhappy councilman who lived nearby led the city council to pass an emergency ordinance making such installations illegal and ordering W6QOL to take it down. His response was to file a suit charging civil rights violation in the Federal District Court for the Central District of California.

On May 2, 1978, Judge Robert M. Takasugi granted a preliminary injunction that prohibited Placentia from enforcing its ordinance but limiting the antenna to 50 feet. On December 11, 1978, the preliminary injunction was made permanent, noting that the ordinance had infringed W6QOL's right to free speech and ordering the city to review and revise its ordinance to conform with the Constitution. On June 3, 1980, the court awarded W6QOL his attorney's fees as "prevailing plaintiff in the Paragraph 1983 action pursuant to the Civil Rights Attorney's Fees Act."

W6QOL's Antenna Was Still Limited to 50 feet, however, until a September 26 ruling by Judge Takasugi that modified his permanent injunction by removing the height restrictions. Placentia has 30 days in which to appeal, but it's considered unlikely that it will. The city has already spent a great deal of money on this case, and an appeal would cost it a good deal more, with at best a marginal chance of success.

Details On This Unusual antenna case will be available from both the Personal Communications Foundation, which assisted K6JAN during the proceedings, and the ARRL.

THE COMMUNICATIONS ACT REWRITE IS DEAD for this session of Congress. The House Judiciary Subcommittee has voted unanimously to recommend delaying further Congressional consideration of the often stalled and controversial legislation until Congress's next term, essentially ensuring it's a dead issue for now. Biggest current problem with the rewrite was the possible effect its proposed restructuring of AT&T would have on the government's antitrust case against Bell Telephone.

Although Another Rewrite effort can surely be expected in the next Congress, there's a serious question as to just what it is likely to contain. Each rewrite attempt has some significant shifts in emphasis, and the next one should be no exception. One addition that can be expected, however, is a provision, similar to Rep. Preyer's bill and the current California legislation, to control or restrict unscramblers and other equipment designed to intercept pay TV signals.

Rep. Preyer's Bill has been modified by Congressmen Smith (Washington) and Waxman (California) in attempts to further strengthen protection for the subscription TV industry. Their new version is directed specifically at the "commercial piracy" firms, a move that apparently will resolve the potential threat to Amateurs who wish to work on homebrew gear, and their suppliers.

That California Bill Has Finally been signed by Governor Jerry Brown, making it illegal in California to manufacture, distribute or sell "any device or plan or part for the knowing purpose of facilitating an unauthorized interception or decoding of subscription TV signals." This bill is so broad in its scope that it's sure to be challenged in court—even one of the subscription TV firms is thinking of going after it.

ATTEMPTS BY RC MODELERS TO GET 6 meters for non-Amateur RC use was to come up for hearing before the FCC on Thursday, November 6. Unhappy with an earlier staff opinion that only licensed Amateurs could operate RC equipment in the 6-meter band, the Academy of Model Aeronautics petitioned for a formal review before the Commissioners and staff. They'd like to bring about a rules change to permit anyone to operate 6-meter RC transmitters under the supervision of "a licensed Amateur." However, Part 97 still requires an Amateur license to operate an Amateur transmitter, though a "third party" may communicate through an Amateur station with a "control operator" standing by. Since Radio Control is a one-way transmission the rules pertaining to third party communications should not apply, so any decision to permit someone not holding an Amateur license to operate a transmitter on Amateur frequencies—even under "supervision"—would be a departure.

COST OF AMATEUR GEAR IN CANADA should be dropping sharply, following the long hoped-for elimination of import duty on Amateur Radio equipment. New Tariff Item 44535-2, passed on October 28 and effective October 29, removed the 15 per cent tariff formerly charged Canadians on "Amateur transmitters, receivers, transceivers, transverters, assembled or in kit form, designed for use only on Amateur bands of the radio frequency as defined by regulations made pursuant to the Radio Act; linear amplifiers, VFOs and power supplies designed for use with the foregoing, parts of all the foregoing." The federal sales tax of 9 per cent still pertains, however, and equipment not specifically made for Amateur use—for example, general coverage receivers—is still subject to the 15 per cent bite.

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74.4 1500
71.9 1000
67.0 600
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118.8 2100
123.0 2150
127.3 2200
131.8 2250
136.5 2300
141.3 2350
146.2 2400
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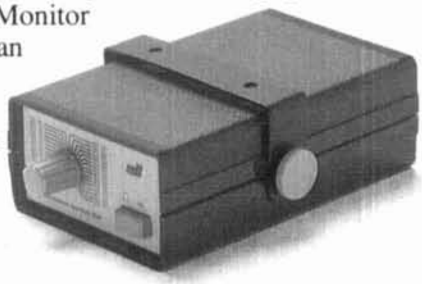

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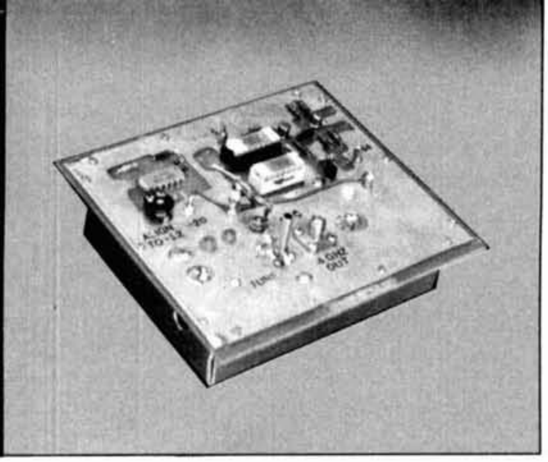


multipurpose voltage-tuned UHF oscillator

This easy-to-build oscillator features multiple-band application, remote tuning, and phase-lock capability

This uhf oscillator is the result of much experimentation. It has an outstanding record of utility and performance. Despite the opinion of many Amateurs, a good uhf oscillator *can* be built without a shop full of machine tools, expensive test equipment, and a high degree of manual dexterity. The PC boards that have been developed for the circuit described here will allow anyone to build a voltage-tuned uhf oscillator.

general description

This oscillator has many applications. It was originally intended for use as the local oscillator in a 1215-1300 MHz TV converter. Later, the board was modified so that the operating-frequency band could be moved up or down to satisfy various other applications. Finally, provisions were made to add either a doubler or tripler circuit to extend the useful output frequency range into the microwave region.

features

The fundamental tuning range of the circuit covers ≈ 1120-1300 MHz. However, by changing the lengths and locations of the frequency-determining circuit elements on the PC board, the operating-frequency range can be adjusted to about 900 MHz and 1400 MHz, giving coverage between 900-4200 MHz with the help of the multiplier circuits.

A varactor provides continuous tuning from a remotely located potentiometer. This feature may be important if you're interested in weak-signal detection, because it allows the entire converter, including the uhf local oscillator, to be located where it belongs — at the antenna.

For television applications, the oscillator may be operated either in the free-running mode or phase locked to a stable reference signal.

The addition of phase-lock capability is easy, because the basic oscillator already includes a tuning varactor. Remote tuning can be used with or without the phase-lock feature. The uhf oscillator is simple. No need for a crystal multiplier chain; therefore no need to struggle with unwanted crystal-oscillator harmonics. Also, if your interest lies in ATV, where crystal control may not be necessary, the design is a natural because of its simplicity.

A divide-by-40 prescaler is mounted on the PC board with the oscillator. The prescaler drives an external frequency counter to monitor the oscillator frequency. Not only is the counter useful as a frequency indicator, it's needed for setting and adjusting the oscillator. The prescaler also provides a signal for the phase detector.

Numerous techniques can be used to phase lock the uhf oscillator to a crystal reference to achieve a high degree of frequency stability; many articles have been written to describe them. In this article, attention is placed on a simple technique that uses a crystal clock as the phase-locked loop (PLL) reference and manual tuning to select the desired lock point. By the proper choice of crystal frequency and divider chains, the uhf oscillator may be locked to any one of a number of desired frequencies. Tuning is done with a ten-turn pot.

applications

Fig. 1 illustrates a typical ATV application that employs the uhf oscillator in the free-running mode as the local oscillator for the mixer. No phase-locked loop is associated with this circuit. A single shielded wire connecting the operating position with the converter serves for tuning, and the converter output is fed over a length of inexpensive transmission line to the receiver. This arrangement avoids the usual degradation in signal-to-noise ratio that generally results from transmitting the rf signal over a long transmission line.

By Norman J. Foot, WA9HUV, 293 East Madison Avenue, Elmhurst, Illinois 60126

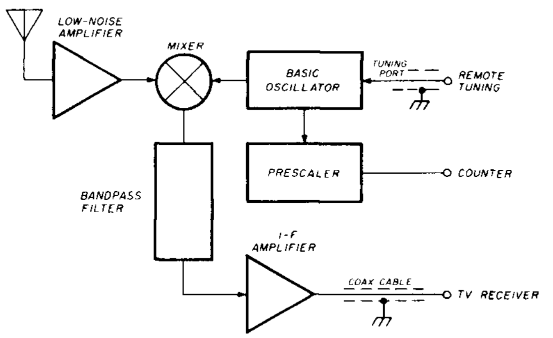


fig. 1. Functional block diagram showing the uhf oscillator in a typical ATV application (free-running mode).

In applications where frequency stability is important, or where a click-stop form of tuning is desired, the basic oscillator can be locked to a stable reference. A block diagram of such a scheme is illustrated in **fig. 2**. The i-f output from the mixer feeds a bandpass filter wide enough to pass the entire band of frequencies of interest, while a wideband fm or television receiver provides the necessary tuning and selectivity. A preselector may be needed between the low-noise preamplifier and the mixer, depending on the application and choice of intermediate frequency.

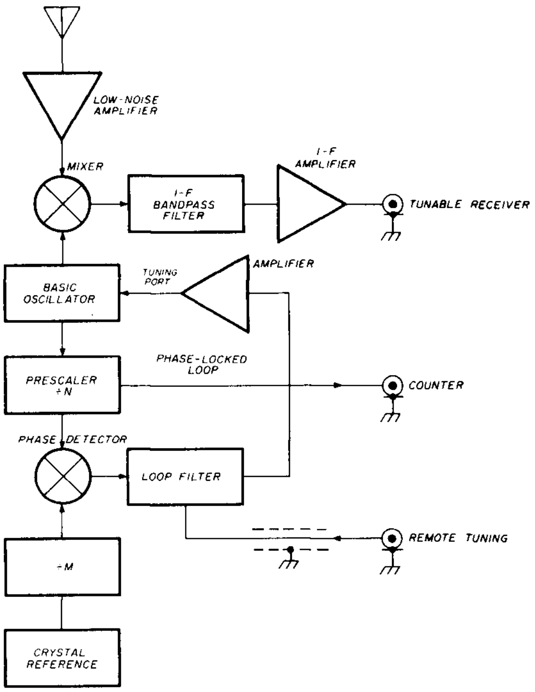
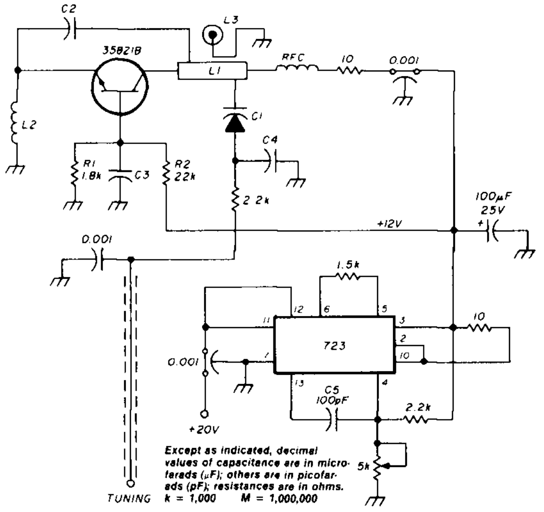


fig. 2. Basic uhf oscillator used in a phase-locked application. Crystal oscillator provides a stable reference for frequency stability.

In both of these arrangements, a frequency scaler drives a frequency counter to permit measurement and continuous monitoring of the uhf oscillator frequency. It's convenient to have this capability, whether the phase-lock feature is used or not. If a programmable counter is available, the readout can display the signal frequency rather than the oscillator frequency.

The advantages to be gained by use of the uhf oscillator described here are now apparent. In some applications the basic oscillator and prescaler alone may do the job, and continuous tuning from a remote location can be used; or a simple PLL may be added for bandswitching, with tuning and selectivity provided by an fm or TV receiver. In either case, a counter can monitor the oscillator (or the equivalent signal) frequency. Other applications can be accommodated using the same PC board with minor modifications, and frequency multiplication can be added for application up into the microwave region.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 3. Schematic of the uhf oscillator. Capacitor C1 is the varactor tuning diode (GHZ devices GC-1607 or equivalent — 3.3 pF at −4.0 volts).

the uhf oscillator

The transistor selected for the uhf oscillator (**fig. 3**) is the HP-35821B. It has an f_t of 4.5 GHz. In the commonbase configuration it's ideally suited for oscillator service. The 35821 has been around for over ten years and is inexpensive. As an oscillator, it can provide 50 mW or more of useful output power with good efficiency.

The base terminals of the 35821 are soldered di-

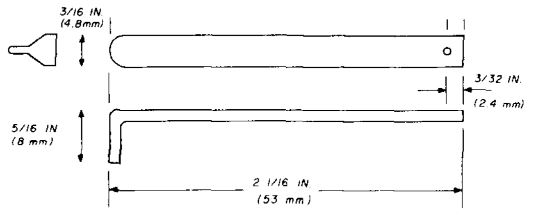


fig. 4. *Mechanical details of inductor L1 in fig. 3.* Material is 0.032-inch-thick (0.8 mm) brass stock.

rectly to the pad provided on the PC board. The board is G10, which is entirely satisfactory for use over the uhf oscillator fundamental tuning range. The board includes all the rf bypass capacitors associated with the oscillator circuit; no chip capacitors are needed.

Fig. 3 is the schematic of the uhf oscillator. There are four special rf circuit elements, L1, L2; C1 and C2. L1 and C1 are the most critical, because they are the principal frequency-determining components. L1 is made of flat brass strip elevated about 0.1 inch (2.5 mm) above the ground plane. The mechanical details of this inductance are illustrated in **fig. 4**.

Capacitor C1 is a varactor tuning diode connected in series with L1 (**fig. 3**). It returns to ground through the large pad under L1 but is electrically above ground to accommodate tuning and automatic phase control. The location of C1 sets the effective length of L1. Moving it back and forth adjusts the tuning range up and down in frequency. The distance between the transistor collector and the tuning varactor should be about 1-1/2 inches (38 mm) to tune the range 1120-1320 MHz. The rf ground pad on the PC board was made long intentionally to provide a wide choice of operating range.

Inductor L2 is a four-turn coil wound with No. 18 (1.0 mm) tinned copper busbar with a 1/8 inch (3 mm) inside diameter. The exact inductance of this coil isn't critical.

Capacitor C2 is a feedback capacitor made from 0.010-inch (0.25 mm) shim brass stock 1/2 inch (13 mm) long and 1/8 inch (3 mm) wide. It is soldered to the emitter and extends over the top of the transistor, parallel with the collector inductance, L1. The feedback capacitor is insulated from L1 with 0.001 inch (0.03 mm) Mylar tape. Feedback is controlled by bending the shim to position it closer or further away from L1. Note that the fixed bias divider consisting of R1 and R2 provides very little forward base bias; consequently, the collector current is primarily determined by the amount of feedback from emitter to collector. This is convenient, because it allows a simple means for properly adjusting the feedback. The correct feedback corresponds to the spacing that produces 30-40 mA collector current. Capacitor C3 is a printed-base bypass capacitor. Capacitor C4, which is the rf bypass for the series L1-C1 circuit, is also printed on the oscillator board.

The rf choke is an eight-turn solenoid wound with No. 24 (0.5 mm) enamel copper with a 1/8 inch (3 mm) ID. The junction of the rf choke and the 10-ohm resistor is supported by the terminal of a push-in Teflon standoff insulator.

power output

Overall converter performance can be degraded because of lack of sufficient local oscillator power. Many Amateurs don't have facilities to measure rf power accurately, in which case the adequacy of their local oscillator is unknown. Mixer noise figures less than 5 dB can be realized with 10 milliwatts of LO power. However, as the LO power is reduced below a few milliwatts, noise figure generally increases dramatically. If the mixer in your system needs the help of more than one low-noise preamplifier, chances are that the mixer noise figure is abnormally high. This is most likely the result of inadequate local-oscillator power. It's possible to reduce the mixer's appetite for LO power by various schemes, including applying dc forward bias to the diodes; but for most practical applications, a good design goal for mixer LO power is 10 milliwatts. This point was kept in mind during the design of the uhf oscillator.

The available power from the uhf oscillator described here is, fortunately, quite high, which allows the output to be loosely coupled; in turn this promotes good free-running stability. When the uhf oscillator is used to drive a doubler, power levels well above 10 milliwatts are easily obtained, with the doubler circuit providing the isolation. Power output from a fixed-tuned tripler was measured at +7 dBm minimum when used with an appropriate idler circuit.

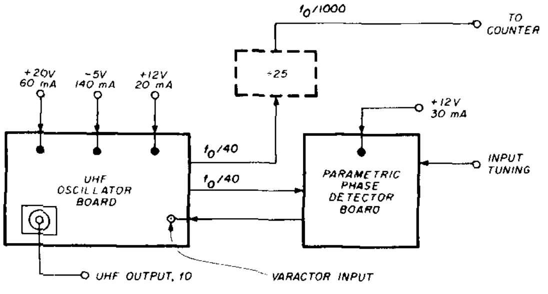


fig. 5. Interface wiring between uhf oscillator board and phase-detector board showing external signal and power requirements.

the phase-locked loop

To provide design flexibility, the oscillator is on one PC board and the phase detector on another. Input signals required by the phase detector are the prescaled signal from the uhf oscillator and the tuning voltage. A single output feeds the VTO (varactor-tuned oscillator) varactor diode for frequency control. **Fig. 5** is a wiring diagram showing a) how these two boards interface, and b) the external signal and power requirements.

The circuit on the phase detector PC board is identical in most respects to the parametric phase detector described in reference 1. This circuit provides considerable design flexibility. In the application here, it operates at about 30 MHz. The circuit (**fig. 6**) also includes provisions for the reference generator, consisting of a quartz crystal and a CD4060B oscillator and divider chain.

Fig. 6 shows the parametric phase detector. This board includes most of the PLL key components, which are the reference generator, spectrum generator, phase detector, and loop filter and dc amplifier. **Fig. 7** shows the phase detector foil and parts layout.

reference signal

The lock points for the uhf VTO are specified in terms of the reference-signal frequency and the prescaling factor. For example, assume the VTO is to be used as the local oscillator in a 23-cm ATV converter and 6-MHz lock-point separation is desired. If a 45-MHz i-f is to be used, the local oscillator frequencies will be 1206, 1212, 1218, and 1224 MHz, corresponding to signal frequencies of 1251, 1257, 1263, and 1269 MHz.

The lock points are 6 MHz apart at the oscillator frequency, but only 150 kHz apart at the phase detector because of the prescaler. The reference needed by the phase detector is therefore 150 kHz. Note that the 202nd harmonic of 150 kHz is 30.3 MHz, which is the spectral line recognized by the phase detector for the 1218-MHz phase lock. Thus, in this type of phase detector, the reference signal must be rich in harmonics. To accomplish this, the phase detector board includes a spectrum generator. On the other hand, if you're interested in a single operating frequency (1257 MHz for example), a crystal-controlled signal at 30.3 MHz is all that's needed. There are, of course, many other schemes that may be used depending on the application.

Tuning and locking to a particular point is easily accomplished by watching the counter. When unlocked, the units and tenths of kilohertz digits will fluctuate due to jitter. When locked, all counter digits will remain steady, and it will be possible to rock the tuning knob back and forth within the hold-in range with no apparent change in the counter status. The final setting should be near the center of the hold-in range.

The pull-in range of the PLL should be less than half the lock point separation; otherwise, if power is momentarily lost, the oscillator may end up locked to the wrong channel. Pull-in range can be controlled

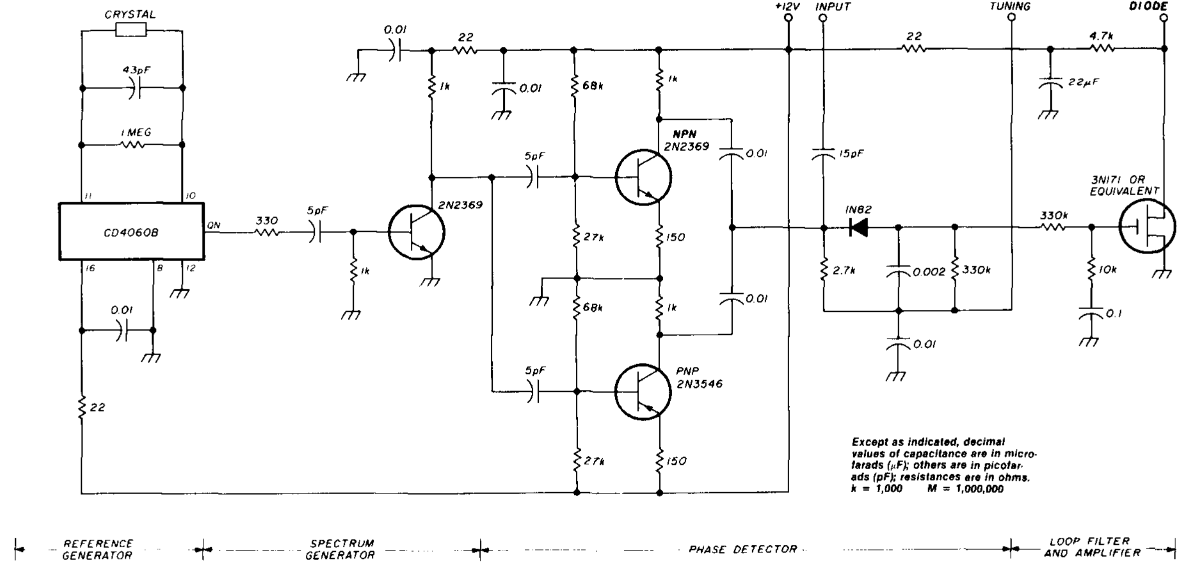


fig. 6. Parametric phase-detector circuit. This circuit includes most of the PLL key components, such as the reference generator, phase detector, and loop filter and dc amplifier.

by adjusting the power level of the prescaled uhf oscillator signal at the input of the phase detector.

prescaler

The Plessey SP-8610 is a 1-GHz divide-by-four prescaler that works well considerably above 1 GHz, even when mounted in a DIP socket. This chip, together with the Plessey 8636 decade divider, provides outputs in the 27-33 MHz frequency range. The circuit is simple and straightforward. One important consideration is that prescalers used at these frequencies require leadless bypass capacitors. Chip capacitors used initially performed satisfactorily from an electrical standpoint, but PC-board flexing caused them to work loose. To solve this problem, leadless capacitors were made by modifying dipped mica capacitors. The insulation was removed with a file, uncovering two metal clamps that hold the stack together. Connections were made directly to the clamps by soldering. This arrangement is entirely satisfactory and considerably less expensive.

The output from the SP-8636 drives a 2N5179 NPN transistor amplifier, which, in turn drives a 2N918 splitter to provide dual low-impedance outputs. One of these is intended to drive the phase detector, while the other can be used to operate the frequency counter. I suggest that an external divide-by-25 circuit be added to increase the overall division factor to 1,000 for the counter. This circuit adds a convenience that relates counter kilohertz to oscillator megahertz. For example, the counter will display 1200 kHz when the uhf oscillator frequency is 1200 MHz.

A schematic of the prescaler is shown in **fig. 8**. An input signal is coupled to the SP8610 by a small probe bent in an L shape and soldered to pin 4. The bent part of the probe is approximately 1/4 inch (6 mm) long and spaced 3/32 inch (2.4 mm) from L1. The probe should be carefully insulated with Mylar tape to prevent it from coming into contact with +12 volts on L1. Also, to prevent damage, do not overcouple the 8610. The proper procedure is to tune the oscillator to the high end of its range and couple the probe sufficiently for the counter to operate properly.

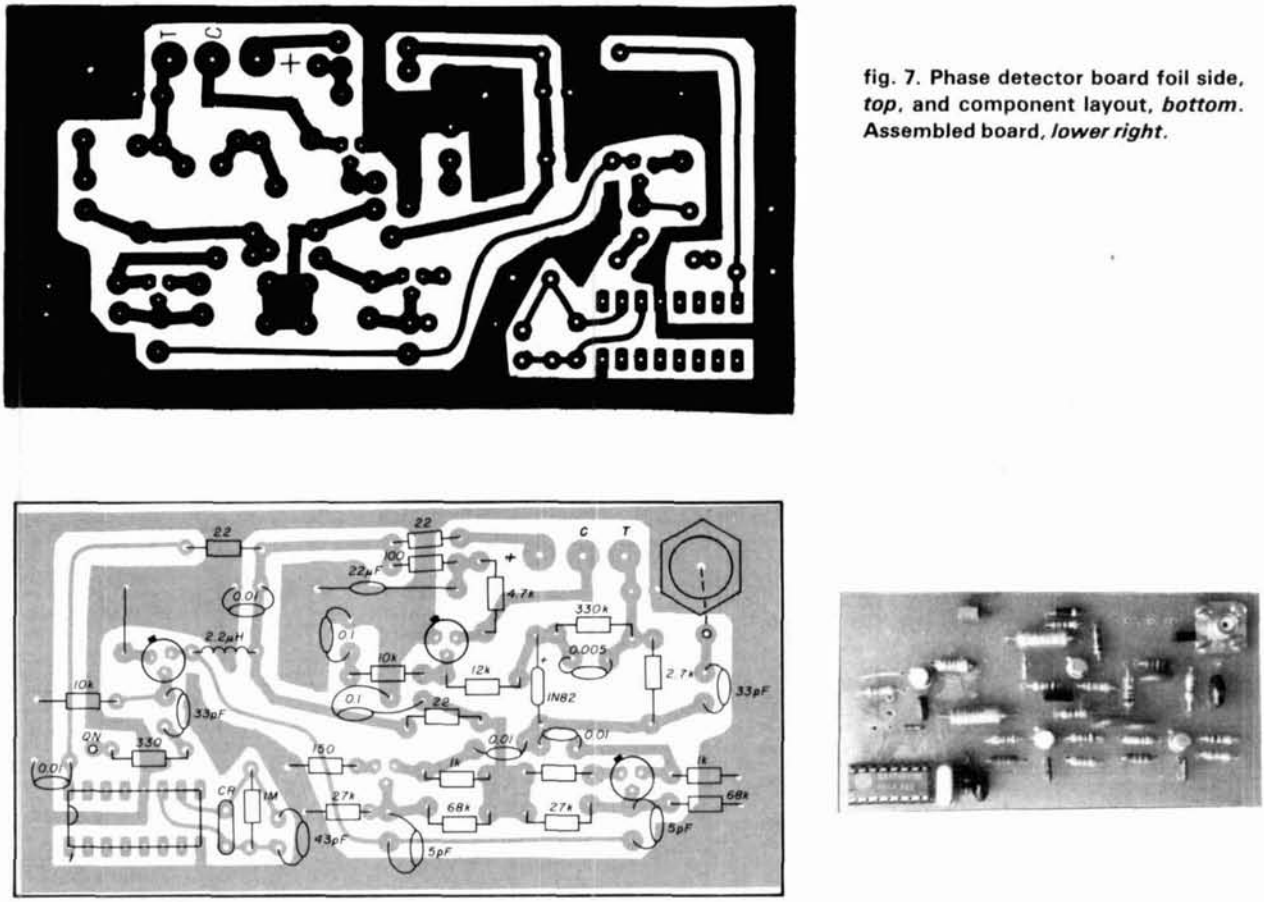


fig. 7. Phase detector board foil side, *top*, and component layout, *bottom*. Assembled board, *lower right*.

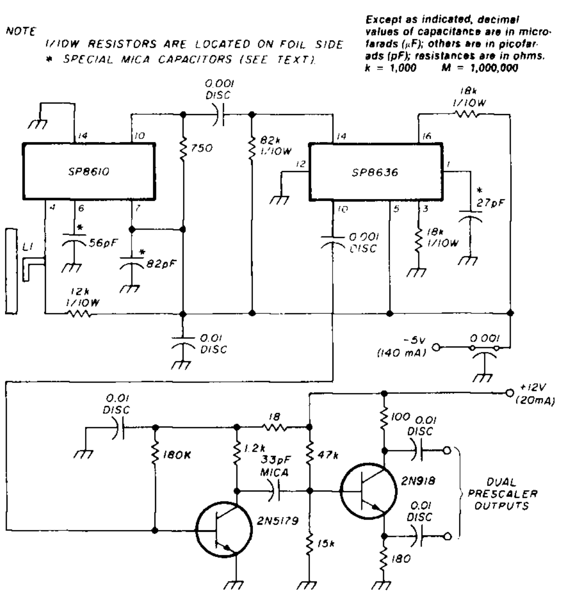


fig. 8. Prescaler schematic. An input signal is coupled to the SP8610 by a small L-shaped probe, which is soldered to pin 4. See text for correct coupling adjustment.

At 1200 MHz, a very small coupling capacitance is sufficient.

construction details

The task of duplicating the performance of the original uhf oscillator is relatively simple when PC boards designed specifically for this project are used. If you don't have the facilities to etch your own boards, they can be obtained from Rock Engineering Supply Company, Inc., 1769 Armitage Ct., Addison, Illinois 60101.

Construction sequence. For the most part, the uhf oscillator assembly is simple except that there is a certain sequence that makes the task easier if followed. I suggest that the feedthrough capacitors be mounted on the board first, followed by the DIP sockets, then all discrete parts not directly associated with the oscillator. **Fig. 9** is a drilling template to be used to locate the feedthrough holes, shoulder washers, and Teflon standoff. If the oscillator is to be used at its fundamental frequency, holes should be drilled for the SMA connector. The coupling loop dimensions and assembly are shown in **fig. 10** if an SMA fitting is not available, a BNC type may be substituted.

Connection is made to the rf ground-return pad of the varactor diode by inserting a 2-56 (M2) screw in hole **A**, using a fiber washer to insulate it from the ground plane on the component side of the board. This is the terminal used to bring the tuning and control voltage to the varactor diode.

Varactor diode. The varactor tuning diode should be mounted with special care. Locate it on the rf pad with the cathode side up and solder the anode to the pad. Use a toothpick or pointed object to hold the diode in place during the soldering operation. Apply the soldering iron to the pad, *not* the diode, and only long enough for the solder to flow. Then tin the diode cathode terminal using a fine soldering iron tip. Apply as little solder as possible.

Before proceeding further, cement the two phenolic shoulder washers in the base bypass pad holes with two-part epoxy cement. Use the quick-setting (5-minute) variety to avoid a 12-hour cure cycle.

Collector line. The collector line, L1, should be mounted next. Tin the bottom side of the line where contact will be made with the varactor diode. Insert the pointed end of L1 into the collector shoulder washer hole and solder the line to the varactor diode. Also, to take the stress off the varactor diode, a fiberglass shim should be cemented in place under the line near the rf choke. Trim the shim with a file so that it slides under the line without forcing, then ap-

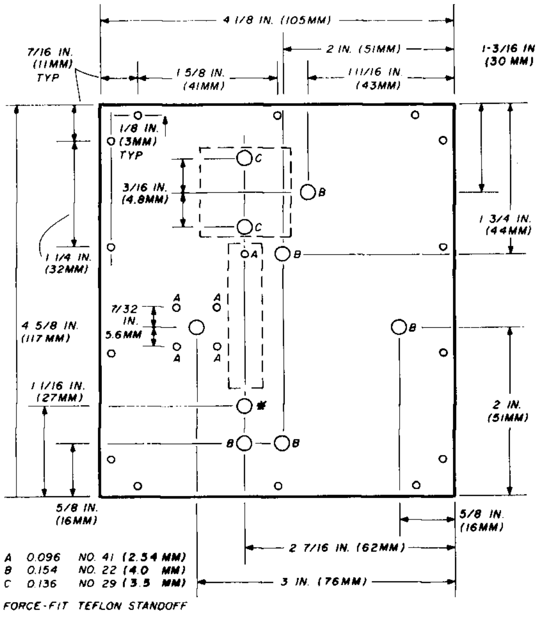


fig. 9. Top view of uhf oscillator board showing mounting-hole locations.

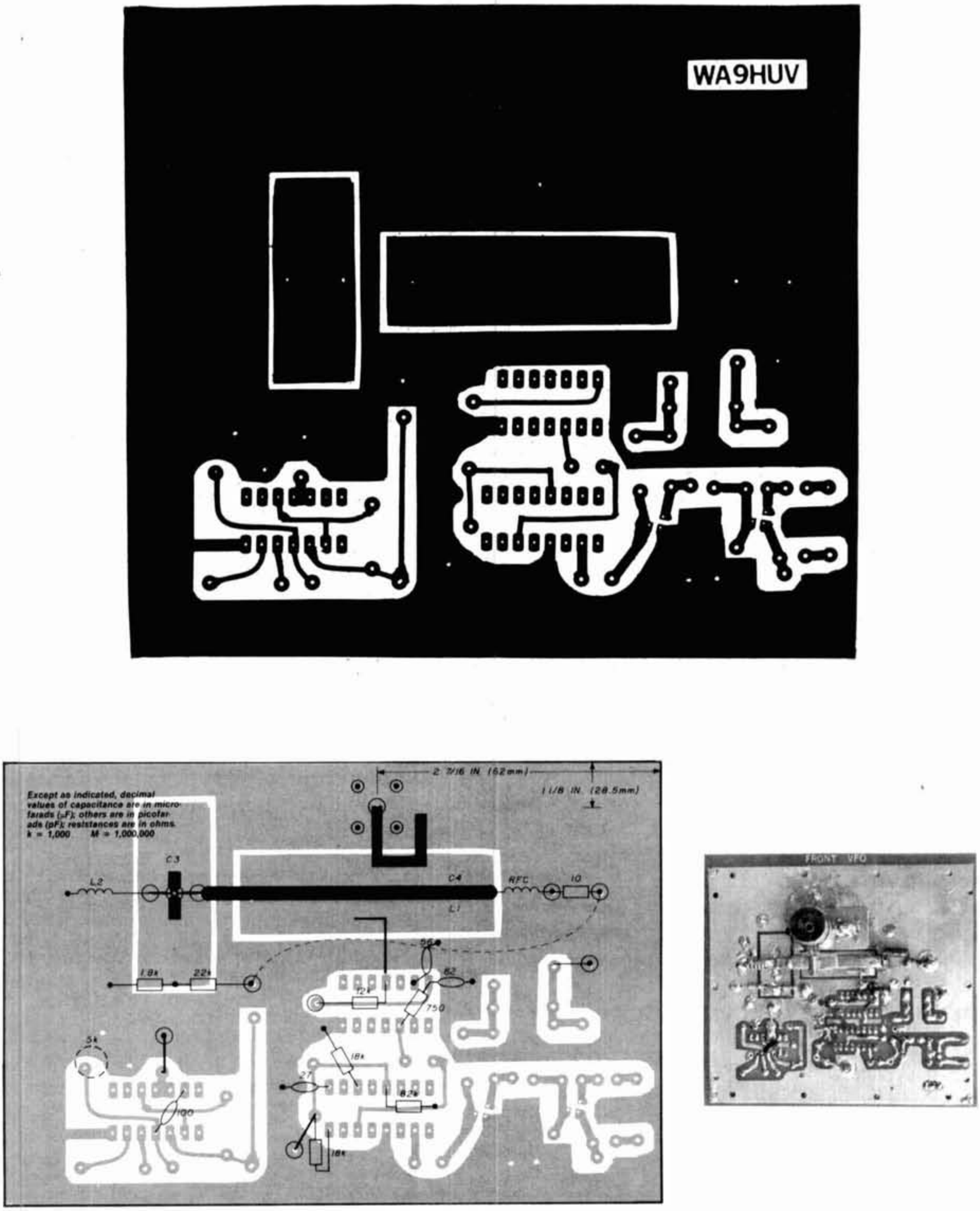


fig. 10. *Top*: Uhf oscillator board, front side. ***Bottom left***: Foil side of oscillator board showing parts placement. ***Bottom right***: Uhf oscillator assembly, top view.

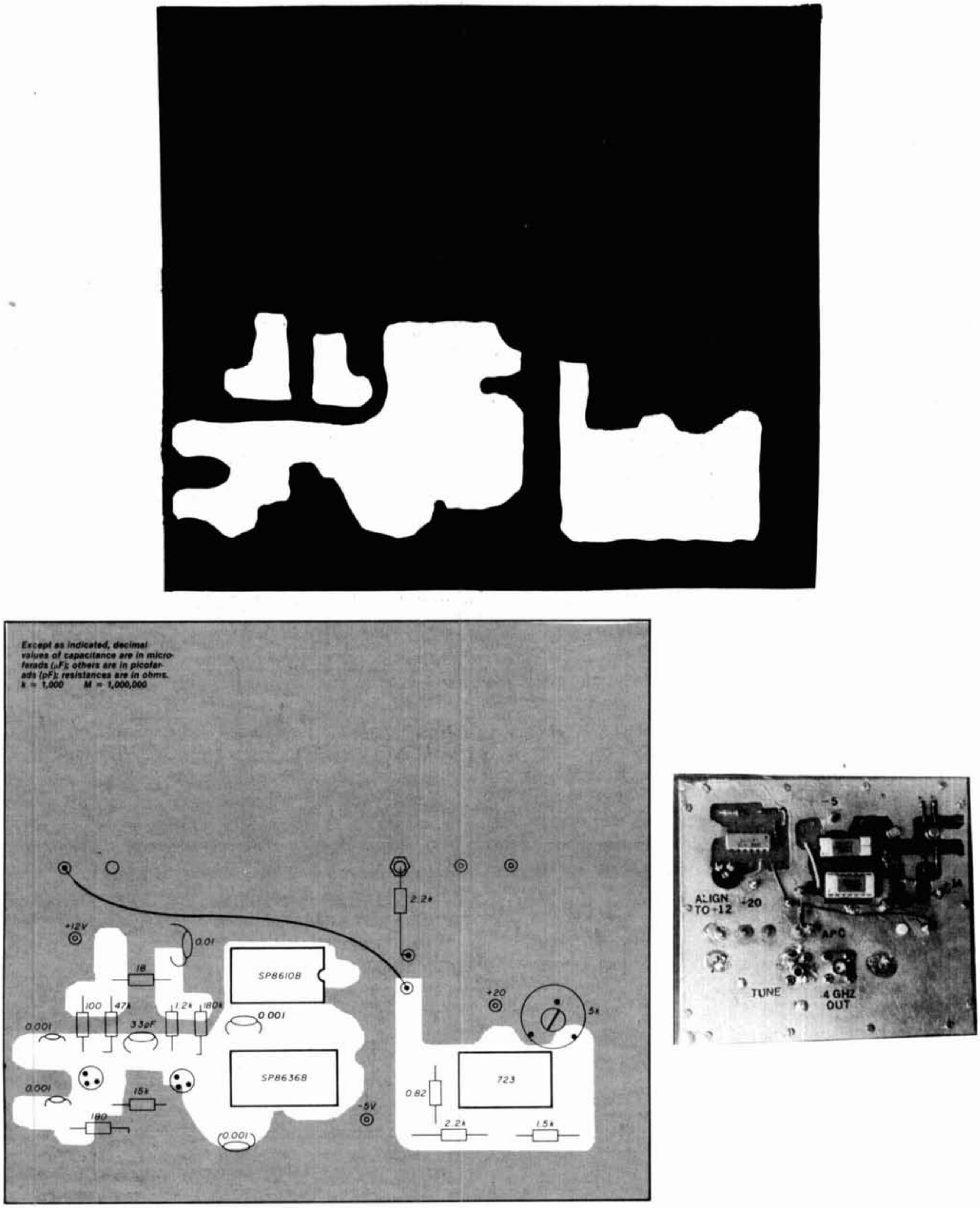

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 11. ***Top***: Uhf oscillator board, rear side. ***Bottom left***: Component-side of oscillator board showing parts placement. Voltage control is a 5k Piheri pot. ***Bottom right***: Uhf oscillator assembly, bottom view.

ply a small amount of epoxy cement and secure the assembly in place. Finally, apply a very small amount of epoxy cement into the collector shoulder washer hole to secure L1.

Emitter coil. The emitter coil should be mounted next, and epoxy cement should be applied to the shoulder washer hole to secure it in place. Mount the transistor on the base pad and solder the base leads to the pad. Solder the emitter and collector leads to the emitter coil and L1 respectively, as shown in **fig. 10**. Solder the feedback shim to the emitter end of L2 (not shown) and insulate the shim with Mylar tape. Space it about 1/8 inch (3 mm) above the collector line.

Before mounting the rf choke and the 10-ohm resistor, check out the 723 regulator and set its output voltage to + 12 volts by adjusting the trimpot.

There are five 1/10-watt resistors and three special mica capacitors that are soldered to the *foil* side of the board (see **fig. 10**). The parts layout on the component side of the uhf oscillator board is shown in **fig. 11**.

Connect a shielded wire from one of the buffered prescaler outputs to a frequency counter and confirm that the counter displays frequencies between ≈ 27-33 MHz as the tuning control is adjusted.

oscillator enclosure

The mechanical details of the aluminum shield cover that encloses the uhf oscillator are shown in **fig. 12**. The 2-56 (M) screws used to mount the shield cover on the board also interconnect the groundplane foils on opposite sides of the board. Since initial tests will be made without the enclosure, it will be necessary to insert the screws and temporarily secure them with nuts to simulate the grounding condition.

initial oscillator tests

The uhf oscillator should be checked out first, without the aid of the phase detector board. Temporarily connect a 10k ten-turn potentiometer between + 12 volts and ground and connect the arm of the pot to the varactor terminal. Use the regulated voltage from the 723 post regulator. Set the tuning voltage to about 5 volts and monitor the current from the 20-volt source with a milliammeter. When power is applied, the current should be approximately 25 mA. Gradually increase the feedback capacitance until the collector current is approximately 35 mA, but do not exceed 40 mA.

Finally, the phase detector board is integrated into the system as illustrated in **fig. 5**, and the PLL is then checked out.

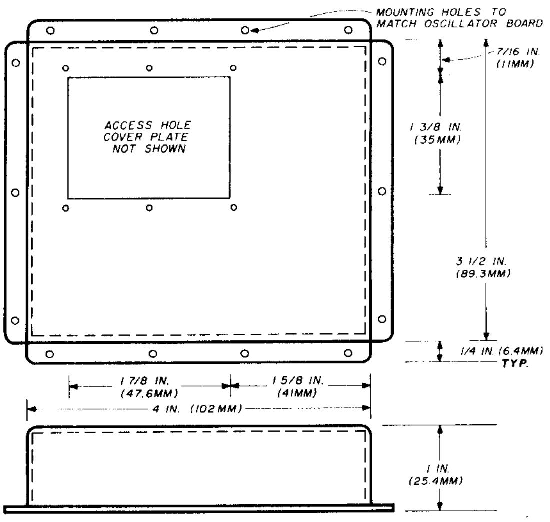


fig. 12. Oscillator enclosure. Material is 0.032-inch (0.8 mm) aluminum.

conclusion

The uhf oscillator described here has many potential applications, depending on your interests. In my case, the performance of an existing 1296 TV converter was considerably improved when the basic uhf oscillator operating in the PLL mode was substituted for the original crystal-oscillator-multiplier chain. A similar uhf oscillator equipped with a doubler circuit was used as the local oscillator in a converter originally designed for use at 2304 MHz. Excellent MDS and ITFS TV pictures were received. Note that the uhf oscillator is not recommended for use in a narrowband receiver intended for CW, am, or SSB service because of its relatively high phase noise.

I've also used the uhf oscillator with a tripler as the local oscillator in a TVRO receiver. In this case, the PLL was built with 20-MHz lock point spacing corresponding to the channel spacing of this class of service. In a future article I'll describe frequency multipliers designed for use with the uhf oscillator.

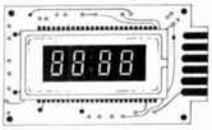
Some of the parts required to build this uhf oscillator probably won't be found in Amateur parts boxes. These include the prescalers, oscillator transistor, and the tuning varactor. I may be able to suggest sources for some of these parts or help you with other problems. In either case, please send an SASE with your inquiry.

reference

1. Norm Foot, WA9HUV, "High-Frequency Communications Receiver," *ham radio*, October, 1978, page 10.

ham radio

National Semiconductor Clock Modules



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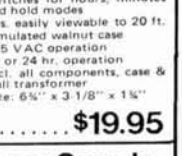

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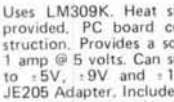

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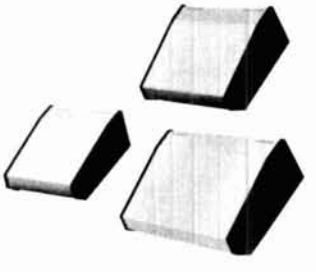

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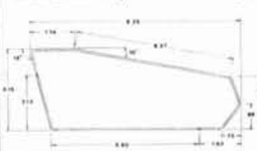

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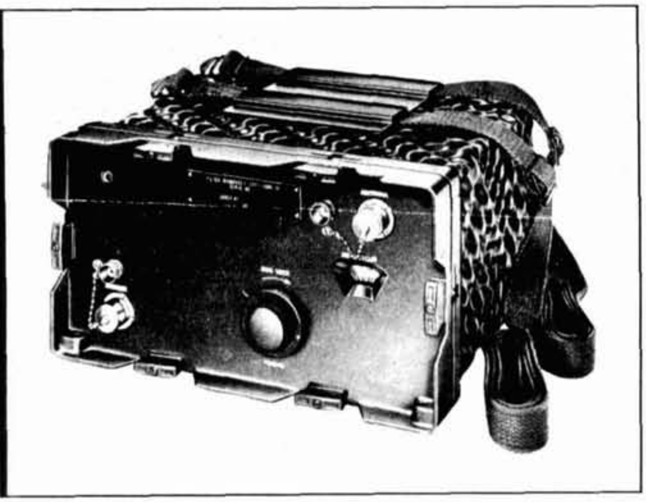

conversion versatility

using the F-237/GRC surplus cavity filter

Good news for
VHF/UHF experimenters —
this surplus filter
can be easily converted
for use on
6, 2, and 1-1/4 meters

In two recent articles,[1,2] I described the conversion of several obscure surplus cavity bandpass filters for use in the vhf and uhf Amateur bands. Since then I've found another very interesting surplus cavity bandpass filter* that I've converted for use in the 50-54 MHz, 144-148 MHz, and 220-225 MHz Amateur bands.

The theory and operation of resonant-cavity bandpass filters have been fully covered in the literature[3] and in my two previous articles. Therefore I'll go right into a description of this surplus "sleeper" and the conversions.

the F-237/GRC-10 bandpass filter

This filter was designed for use with the receiver section of Army radio set AN/GRC-10 and consists of three individual coaxial resonant re-entrant cavities connected in cascade, each tuned with its own variable capacitor ganged for single-dial control.

---

*Fair Radio Co., Post Office Box 1105, Lima, Ohio 45802

By William Tucker, W4FXE, 1965 South
Ocean Drive, 15-C, Hallandale, Florida 33009

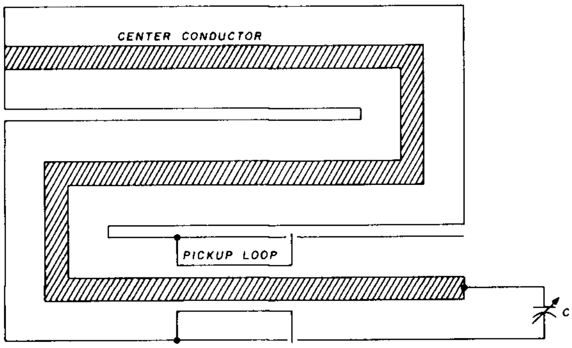

fig. 1. The F-237/GRC-10 bandpass filter showing the zig-zag configuration to compress 22 inches (55.9 cm) of coax cavity into a compact package. Note that the pickup loops are close to the open end.

Each cavity is about 20 inches (51 cm) long but compressed into a compact package by using a snake-like configuration as shown in **fig. 1**. The cavities are of sturdy copper, and the center conductor is silver plated for high conductivity.

Normally, rf pickup loops are located near the shorted high-current end of coaxial type re-entrant resonant cavities where the electromagnetic field is at a maximum. Note that in this cavity, however, the pickup loops are located closer to the open end, evidently to provide looser coupling. This will provide greater selectivity at the expense of a higher insertion loss, which becomes a little over 2 dB per cavity.

The three cavities are similar electrically and physically except that the input and output pickup loops L1 and L6, (**fig. 2**) are a little larger than the others. Also the coaxial cable connection to each cavity varies slightly.

Receiver and antenna jacks on the front panel are made to accommodate a type-C UG-573 connector, which is a jumbo type BNC that's not in general use. If you wish, an N type or uhf type socket can be used in its place by removing the existing socket. Some filing of the socket flange may be necessary to fit into the recessed opening on the front panel.

The F-237 has an input and output impedance of 50 ohms and covers 54-70.9 MHz with continuous tuning. The bandwidth at the 3-dB points is 250 kHz. The attenuation is 40 dB at 4.5 MHz. Insertion loss is 7 dB at resonance. The complete assembly in its cabinet weighs about 16 pounds (7.3 kg) and is approximately 6 × 11 × 11 inches (15 × 28 × 28 cm).

simple conversion to the 6-meter band

Fortunately, the three air-dielectric trimmers C1002-3-4, which are mounted directly on the three-gang variable capacitor C1001 A-B-C, **fig. 3**, have sufficient spare capacitance so they can be adjusted to cover the 50-54 MHz band. After adjustment, the range is 49.5-60 MHz.

Because of the high selectivity, the following procedure is suggested. Set the tuning dial at the lowest frequency position, 54 MHz, and feed a 53-MHz signal into the antenna terminal from any convenient source, such as a grid-dip meter or signal generator. Adjust the three trimmers for maximum output as measured at the receiver terminal using an rf meter or receiver S-meter. A simple rf meter can be made using a germanium diode such as the 1N34 in series with a microammeter.

Repeat the above procedure in small steps until 49.5 MHz is reached; the trimmers should now be at almost maximum capacitance with some to spare for final adjustment. If this filter is to be used with a receiver only, it can be inserted into the transmission line and, with a weak signal around 52 MHz, the filter tuning dial can be tuned for maximum output. The trimmers can then be repeaked for maximum output.

If the filter is to be used with a transmitter or transceiver, an SWR indicator should be used between transmitter and filter. The trimmers should be adjusted for minimum SWR at 52 MHz. The tuning dial can then be calibrated in any manner you choose.

lowering the insertion loss

For general Amateur use, 7 dB is quite a large bite to take out of the received or transmitted signal. The F-237 filter assembly can be modified to provide less insertion loss at the expense of a little selectivity by using only one or two of the original cavities instead of all three. Even with a single cavity, selectivity is adequate for most Amateur applications.

To lift out the cavity assembly and its ganged capacitors in one piece, remove all the screws from

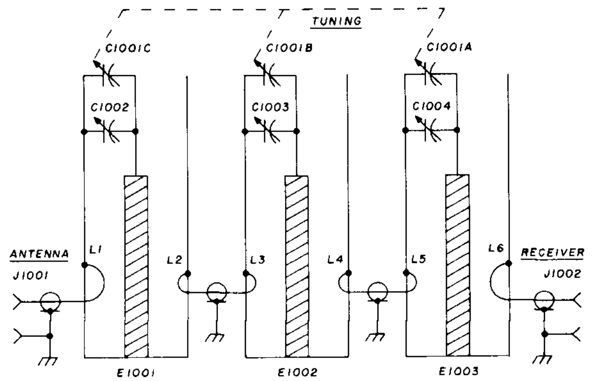

fig. 2. Schematic of the F-237/GRC-10 cavity. The antenna and receiver pickup loops are larger than those connecting the assembly.

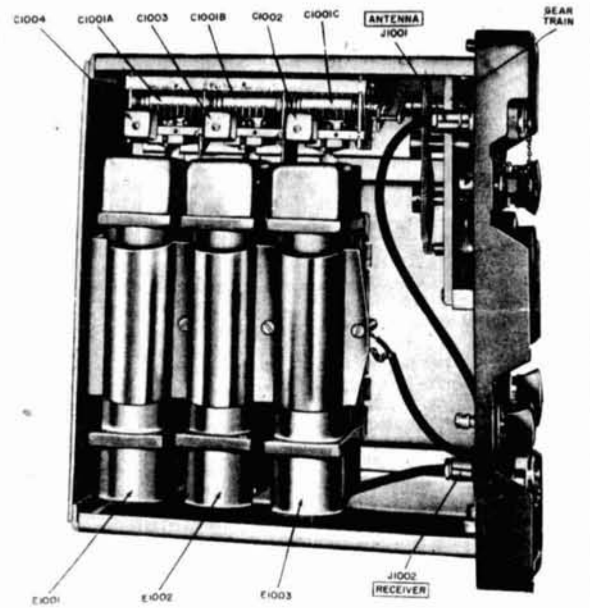

fig. 3. Interior of the F-237/GRC-10 cavity. The three air-dielectric trimmers have enough capacitance to allow coverage of 50-54 MHz.

the underside and unsolder the two coaxial cable leads leading to the front panel. To eliminate a cavity section, remove the Phillips-head screw and unsolder the ground strap. Unsolder the cavity center conductor from the variable-capacitor stator plates and the cavity will unplug from its adjacent cavity (**fig. 4**).

If only one section is to be used, any of the cavities will do. If two sections are to be used, then eliminate the center cavity and interconnect the remaining two with a short length of RG-58/U coaxial cable. This arrangement is necessary to ensure proper tracking. Adjustment follows the original procedure.

even less insertion loss

The insertion loss can be reduced to under 1 dB per cavity section by rearranging the cavity so that the pickup loops are placed in the high-current end of the cavity. This can be done by reversing the cavity sections as shown in **fig. 5**.

Unsolder the closed end plate at **A** and resolder it to the other end, **B**. Make certain that very good electrical contact is made between the center conductor and the housing at this high current end, **B**. Unsolder the ground strap and relocate as shown. Cut a short length of copper or brass rod and insert it into the center conductor at **A** so that it will reach the tuning-capacitor stator. Finally, unsolder the mounting bracket and replace it at the other end as shown.

Reassemble the cavities to the ganged capacitors and you now have a bandpass filter with an insertion loss of less than 1-dB per cavity section. The selectivity is still adequate even if you use only one cavity to do the job. The adjustment and tuning is as previously described.

for use with higher power

The F-237 bandpass filter is tuned to resonance by a three-gang variable capacitor of excellent quality with 0.06-inch (1.5-mm) spacing between plates. It should withstand power levels in the order of several hundred watts. The weak point in the filter is the very small air dielectric trimmers, which will probably arc over with rf power in excess of 30-40 watts. To overcome this limitation, the trimmers can be removed and replaced with the APC type of trimmer, 20 pF or more, and with a plate spacing of at least 0.03 inch (0.76 mm). The larger trimmer will also extend the low range a few MHz below 49.5 MHz.

conversion to the 2-meter band

This conversion can be made from either left-over cavities from the 50-54 MHz conversion or from another F-237. A length of 22 inches (56 cm) of coaxial re-entrant cavity is too long for 144-148 MHz and must be shortened to allow for variable capacitance loading.

Fig. 6 shows a convenient method of obtaining a workable length, while at the same time placing the pickup loops very close to the shorted high-current end of the cavity. In addition, the open end is terminated in a handy housing for the variable capacitor.

As shown in **fig. 7**, carefully eliminate the shaded portion with a sharp hacksaw; this will leave about 11 inches (28 cm) of cavity for the 2-meter band. File all rouugh edges to a flat and smooth finish and tin thoroughly at both ends for soldering. Unsolder the

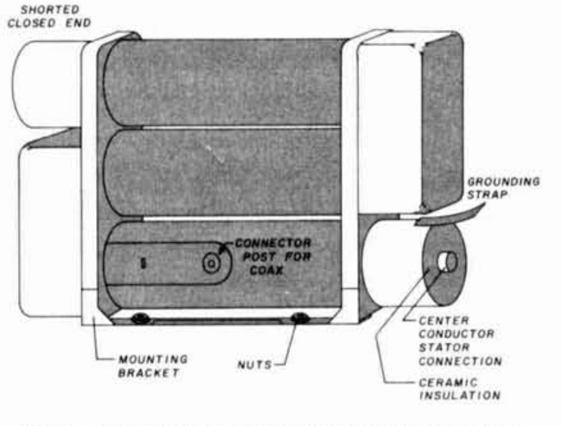

fig. 4. Unmodified cavity section detached from the F-237/GRC-10 assembly.

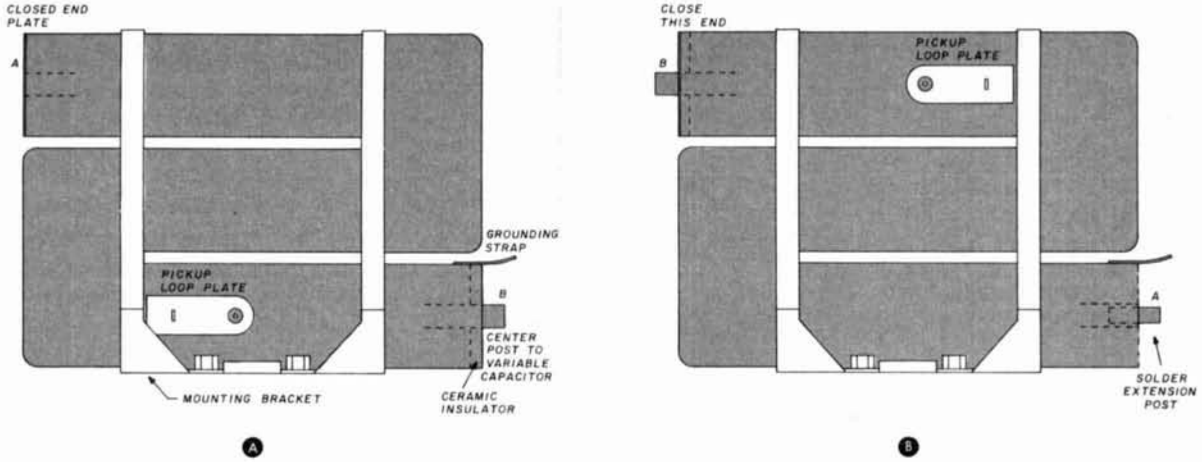

fig. 5. Reversal of cavity to place pickup loops in the high-current area for lower insertion loss. Sketches (A) and (B) show before and after mods.

right-angle portion of the inner conductor as shown.

The two pickup loops will now be visible and accessible from the short open end. Using a screwdriver, bend the center of each loop toward the housing away from the center conductor as shown by the dotted line in **fig. 8**. Try to make the loops as symmetrical as possible.

To close up the end near the pickup loops, unsolder the end plate on the cut-off portion or cut a piece of flashing copper to 1-1/2 inch (3.8 cm) diameter with a 1/4-inch (0.6-cm) opening in the center. Solder either one securely to ensure good electrical contact at this high-current area.

Select an APC air dielectric trimmer capacitor and install in the cubical housing as shown in **fig. 9**. A capacitance of about 25 pF with an air gap spacing of at least 0.03-inch (0.76-mm) should fit into the available space and provide adequate tuning range. Solder the stator plates to a heavy lead and attach to the center conductor. The rotor wiper arm should be soldered directly to the housing wall. Try to obtain an APC trimmer with a standard 1/4-inch (0.6-cm) shaft so

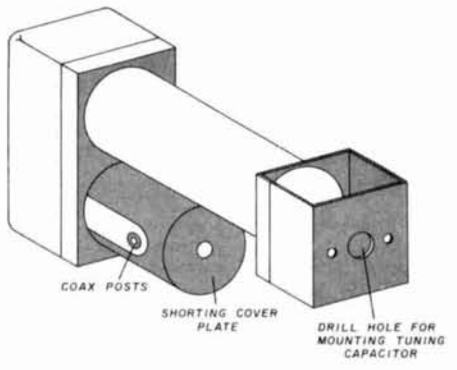

fig. 6. Modified cavity section for the 2-meter band. A workable length is obtained, and pickup loops are close to the shorted high-current end of the cavity.

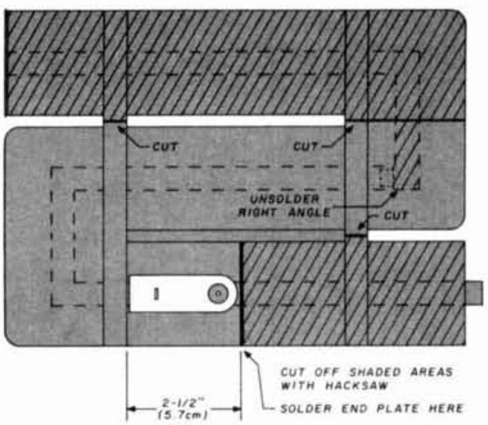

fig. 7. Shaded portions are removed with a hacksaw to leave about 11 inches (28 cm) of cavity for the 2-meter band.

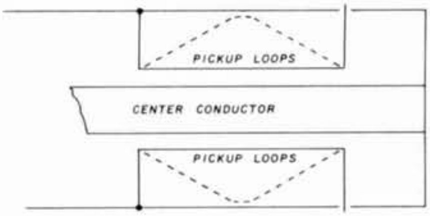

fig. 8. Pickup loops are bent as shown for the 2-meter conversion.

that a knob can be used instead of the inconvenient screwdriver adjustment.

To test the unit for frequency coverage, attach a 3/4-inch (1.9-cm) loop to either coaxial terminal and couple a grid-dip meter to it. A sharp dip will indicate resonance, which should occur about midrange with plenty of spare capacitance on either side of resonance. The open end of the cavity can then be closed with flashing copper or left open as you wish.

conversion to 220-225 MHz

This modification is identical to the 144-148-MHz conversion except for the tuning capacitor. At this frequency, even the minimum capacitance of the APC trimmer is too high; therefore, a simple very low capacitance trimmer can be built using two copper pennies. Solder one penny to the inner conductor and the other to a brass machine screw as shown in **fig. 10**. Solder a brass hex nut to the outside of the housing and use a second hex nut to lock in the frequency adjustment. A grid-dip meter can be used to check the frequency range, which should be between approximately 180-240 MHz.

an experimenter's delight

The several conversions discussed in this article are just a small sampling of what can be done with the F-237. One assembly will supply three cavities; one for each band, or all three for one band.

For those who wish to experiment, a length of cavity somewhat shorter than the 11 inches (28 cm) used for the 144-MHz band can be used with a 50-pF air trimmer to provide coverage of both the 144- and 220-MHz bands with one cavity. Also, by using a shorter length of about 3-5 inches (7.6-12.7 cm), this cavity section can be made to resonate in the 440-MHz band.

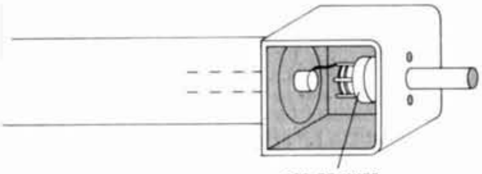

fig. 9. Installation of the APC trimmer capacitor for the 2-meter conversion.

The size of the pickup loops, which serve an important role in impedance matching and determining cavity selectivity, can be changed by unsoldering the elongated mounting strip for easy access. Also, for convenient cable connection, small sockets such as the BNC, F, or RCA type can be used as they are small enough to be mounted into the strip.

Another suggestion: You can attach three modified cavities, each for a different band, to the stators of the three-gang tuning capacitor. Separate sets of coaxial cables can be run to sets of separate terminals on the front panel, or a three-position switch can be used to select the cavity to be used. Depending on the length of each cavity, the individual capacitor sections can be used to tune the desired band. If the capacitance is too high, rotor plates can be easily removed to lower capacitance to fit the application. The main tuning dial can be calibrated with three separate scales, as required.

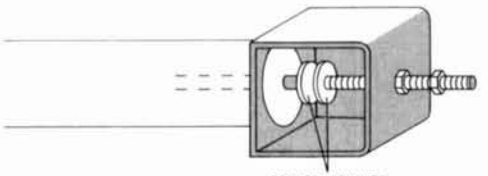

fig. 10. Air trimmer installation for the 22-MHz conversion using two copper pennies to replace the APC trimmer capacitor.

summary

With 66 inches (167.6 cm) of good-quality coaxial cavity available, a three-gang variable capacitor, three shielded miniature air dielectric trimmers, a precision tuning assembly, and a sturdy metal cabinet, vhf and uhf experimenters can really have a field day with the F-237/GRC-10.

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1. William Tucker, W4FXE, "How to Modify Surplus Cavity Filters for Operation on 144 MHz," *ham radio*, February, 1980, page 42.
2. William Tucker, W4FXE, "More Conversions of Surplus Cavity Bandpass Filters," *ham radio*, March, 1980, page 46.
3. William Tucker, W4FXE, "How to Modify Surplus Cavity Filters for Operation on 144 MHz," Bibliography, *ham radio*, February, 1980, page 46.

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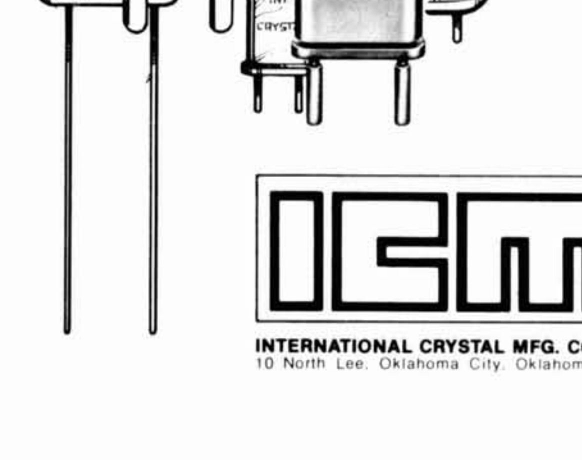

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Synthesized/PTO Frequency Control—A Drake exclusive: carefully engineered high-performance synthesizer, combined with the famous Drake PTO, provides smooth, linear tuning with 1 kHz dial and 100 Hz digital readout resolution. 500 kHz up/down range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

Advanced, High-Performance Receiver Design—The receiver section of the Drake TR7 is an advanced, up-conversion design. The first intermediate frequency of 48.05 MHz places the image frequency well outside the receiver input passband, and provides for true general coverage operation without i-f gaps or crossovers. In addition, the receiver section features a high-level double balanced mixer in the front end for superior spurious and dynamic range performance.

True Passband Tuning—The TR7 employs the famous Drake full passband tuning instead of the limited range "i-f shift" found in some other units. The Drake system allows the receiver passband to be varied from the top edge of one sideband, through center, to the bottom edge of the opposite sideband. In fact, the range is even wider to accommodate RTTY. This system greatly improves receiving performance in heavy QRM by allowing the operator to move interfering signals out of the passband, and it is so flexible that you can even transmit on one sideband and listen on the other.

Unique Independent Receiver Selectivity—Space is provided in the TR7 for up to 3 optional crystal filters. These filters are selected, along with the standard 2.3 kHz filter, by front panel pushbutton control, independent of the mode control. This permits the receive response to be optimized for various operating conditions in any operational situation. Optional filter bandwidths include 6 kHz for a-m, 1.8 kHz for narrow ssb or RTTY, and 500 Hz and 300 Hz for cw.

Broadband, Solid State Design—100% solid state throughout. All circuits are broadbanded, eliminating the need for tuning adjustments of any kind. Merely select the correct band, dial up the desired frequency, and you're ready to operate.

Rugged, Solid State Power Amplifier—The power amplifier is internally mounted, with nothing outboard subject to physical damage. A Drake designed custom heat sink makes this possible. The unique air ducting design of this heat sink allows an optional rear-mounted fan, the FA7, to provide continuous, full power transmit on SSTV/RTTY. The fan is not required for ssb/cw operation, since normal convection cooling allows continuous transmit in these modes.

Effective Noise Blanker—The optional NB7 Noise Blanker plugs into the TR7 to provide true impulse-type noise blanking performance. This unit is carefully designed to maximize both blanking and dynamic range in order to preserve the excellent strong-signal handling characteristics of the TR7.

* NOTE: Transmitter coverage for MARS, Government, and future WARC bands is available only in ranges authorized by the FCC, Military, or other government agency for a specific service. Proof of license for that service must be submitted to the R. L. Drake 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 Aux7 Range Program Board. Prices quoted from the factory. See Operator's Manual for details. (Not available for services requiring type acceptance.)

More Details? CHECK—OFF Page 126

TR7

ACCESSORIES

Model 1336	Drake TR7 General Coverage Digital R/O Transceiver
Model 1338	Drake RV7 Remote VFO
Model 1502	Drake PS7 120/240V Ac Supply for continuous duty operation (25 amps)
Model 1570	Drake PS75 120/240V Ac supply for intermittent duty (15 amps continuous, 25 amps intermittent)
Model 1553	Drake SP75 Speech Processor
Model 1230	Drake LA7 Line Amplifier
Model 1533	Drake CS7 Coax Switch
Model 7077	Drake Desk Microphone
Model 1520	Drake P75 Phone Patch
Model 1536	Drake Aux7 Range Program Board **
Model 1531	Drake MS7 Matching Speaker
Model 1537	Drake NB7 Noise Blanker
Model 1529	Drake FA7 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 TR7 Service Kit/Extender Board Set
Model 385-0004	Drake TR7 Service/Schematic Book

**Aux7 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 Aux7.

TR7 SPECIFICATIONS

GENERAL

Receive	
Without Aux7	1.5 to 30 MHz, continuous, no gaps.
With Aux7	Same, plus 0 to 1.5 MHz at reduced performance.
Transmit	
Without Aux7	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 Aux7*	Above ranges, plus any eight 500 kHz segments from 1.8 to 30 MHz.
Modes of Operation	Usb, Lsb, Cw, RTTY, A-m equiv. (A-3H).
Frequency Stability	Less than 1 kHz first hour. Less than 150 Hz per hour after 1 hour warm up. Less than 100 Hz for ± 10% line voltage change.
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 Freq.	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

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 4.4 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 spacings of 100 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 FA7 Fan—33%, 5 min. transmit, max. with 1529 FA7 Fan—100%.
Wattmeter Accuracy	± 5% @ 100 watts (50 ohm load).
Carrier Suppression	Greater than 50 dB.
Microphone Input	High Impedance.

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Yagi antennas: practical designs

Last in the Yagi design series, with emphasis on scaling and element taper

In all the previous articles of this series the specifications for a Yagi antenna have been stated only in terms of strictly cylindrical elements. Each element is characterized by an x coordinate or position along the boom, a physical length, LE, and a radius RO; each of these three quantities is expressed in terms of wavelengths, λ, at a central design frequency. Such specifications have led to a number of rather good antenna designs, and I shall shortly list a brief selection of such designs. However, when a real Yagi antenna is constructed it will rarely ever be convenient to adhere rigorously to the given cylindrical element design. To start, the element diameter is usually adjusted to fit a mechanical requirement (wind loading, etc.); moreover, the element itself is usually not a cylinder, but a series of telescoping tubes starting with a large-diameter section at the boom and tapering to a small-diameter section at the outer end of the element. In addition, the element is fastened to the boom with a clamping arrangement that may be a plate or angle bracket U-bolted to both boom and element. Some mechanical designs even put the element directly through the boom. Thus, the path from the cylindrical design to a practical antenna will involve three tasks: scaling the original design to an equivalent new design using a different (average) element radius, computing the potentially significant change in element length as a result of the chosen (telescoping) taper schedule, and making (usually minor) corrections to allow for the boom clamping system. Methods for carrying out each of these three tasks will be given following the next section on preferred antenna designs.

By James L. Lawson, W2PV, 2532 Troy Road,
Schenectady, New York 12309

preferred antenna designs

In this section I shall discuss one preferred design for a two-, three-, four-, five- or six-element Yagi antenna. Recall that simplistic Yagis[4] (element spacing uniform and all directors having a common length) are as good as any other design up to a boom length of one wavelength. It was shown that a good two-element beam would have a boom length of about 0.15λ; the exact length is not critical and is a compromise between better gain and lower efficiency and bandwidth. Best parasite element length is a compromise between better forward gain and lower F/B ratio. For a three-element beam it was shown that a boom length of about one-quarter wavelength produces a naturally high F/B and similarly for four-, five-, and six-element beams a boom length of about 3/4 wavelength gives a naturally good F/B ratio.

Table 1 shows the characteristics of these good Yagi designs. These particular antenna designs are not unique; for example, the boom length can be varied somewhat. Longer booms, in general, give larger forward gain, but the frequency for highest F/B ratio drops somewhat below the center of the band, where gain remains high.

A procedure has also been described that allows fine tuning or optimization to improve the F/B ratio;[5] this optimization procedure can be done for Yagi antennas having four or more elements. Optimization must be done for a specific end use. **Table 2** shows optimized six-element beams first for free-space use, *next for operation at 1.0λ over ground, and finally for* operation in a two-Yagi stack at heights of *0.60*λ and *1.5*λ. These parameters are mathematically correct. But note that approximations used in the model really do not justify complete confidence in the precise values in **table 2**. Nevertheless, I suspect that practical antennas constructed from this table (for use over ground) will exhibit superior properties to the (free-space) 6-element case shown in **table 1**.

scaling

Any of the Yagi antenna designs, such as those in **table 1**, can be scaled either to other center frequencies or to elements of different diameter at the same center frequency. Because all design parameters include *dimensions expressed in wavelengths at a central design frequency*, the design itself is independent of frequency scaling; therefore, the behavior of the antenna will not be affected by the choice of central design frequency. However, this is true only if the *design is truly unchanged; that is, all physical dimensions* (including element radii) are adjusted proportional to the desired wavelength.

table 1. Preferred Yagi antenna designs. All elements with radius, *RO*, of *0.0005260* (λ_0), length, *LE*, in (λ_0), and boom position, *X*, in (λ_0).

element	*X*	*LE*	*X*	*LE*	*X*	*LE*	*X*	*LE*	*X*	*LE*
R	0.000	0.49366	0.000	0.49801	0.000	0.49185	0.0000	0.49994	0.000	0.49528
DR	0.150	0.47050	0.150	0.48963	0.250	0.47900	0.1875	0.48040	0.150	0.48028
D1			0.300	0.46900	0.500	0.46319	0.3750	0.45232	0.300	0.44811
D2					0.750	0.46319	0.5625	0.45232	0.450	0.44811
D3							0.7500	0.45232	0.600	0.44811
D4									0.750	0.44811
number elements	2		3		4		5		6	
gain (dBi)	6.88		7.86		10.62		10.45		10.70	
F/B (dB)	7.94		23.60		41.62		32.27		52.71	

table 2. Optimized 6-element Yagi antenna, *RO* is *0.0005260* (λ_0), *LE* in (λ_0), and *X* in (λ_0).

	A		B		C	
element	*X*	*LE*	*X*	*LE*	*X*	*LE*
R	0.0000	0.49528	0.0000	0.49528	0.0000	0.49528
DR	0.1500	0.48071	0.1500	0.48028	0.1500	0.48157
D1	0.2992	0.44811	0.3039	0.44811	0.3029	0.44811
D2	0.4500	0.44811	0.4500	0.44811	0.4500	0.44811
D3	0.6000	0.44811	0.5959	0.44811	0.6395	0.44811
D4	0.7500	0.44811	0.7500	0.44811	0.7500	0.44811

Note:

A. Optimized in free space.

B. Optimized at *1.0*λ_0 over ground.

C. Optimized in a stack/ground at *0.6*λ_0 and *1.5*λ_0.

Experience has shown that *desired* element radii expressed in wavelengths is not constant; at low frequencies (long wavelengths) relatively thin elements are used, while at high frequencies relatively fat elements are normal. How, then, can a given design be altered to an equivalent design where element radii are changed? The clue is to make the impedance of the changed, or scaled, element exactly the same as the impedance of the original unscaled element at the central design frequency; in this way exactly the same element currents will flow, resulting in the same detailed antenna performance. Because the (radiation) resistance of the element is essentially unchanged, we need only to make the reactance invariant to scaling-element radius.

Recall[2] that element reactance, X, near resonance can be expressed as:

$$X = RQ(F/FR - FR/F) \qquad \textbf{(1)}$$

where R = the (radiation) resistance
Q = the effective Q
F = the frequency referred to central design frequency
FR = the element resonant frequency, also referred to central design frequency.

Recall also that RQ can be (rather accurately) empirically expressed as:

$$RQ = (215.15 \log K - 160) \qquad \textbf{(2)}$$

where $K \equiv 1/RO$
RO = the radius of the element expressed in wavelengths at $F = 1$, the central design frequency.

From **eqs. 1** and **2**:

$$X = (215.15 \log K - 160)\ (F/FR - FR/F) \qquad \textbf{(3)}$$

and at the central design frequency ($F = 1$):

$$X_{(F=1)} = (215.15 \log K - 160)\ (1/FR - FR) \qquad \textbf{(4)}$$

Thus, if we wish to scale the element radius from an original value to a new value, we must ensure that $X_{(F=1)}$ is unchanged. Note that $X_{(F=1)}$ contains two variables, (K and FR), which are a function of element radius RO. Recall[2] FR is calculated from the physical length of element LE and physical resonant length LER; both of these lengths are measured in wavelengths, λ_0, at $F = 1$:

$$FR = LER/LE \qquad \textbf{(5)}$$

Empirically,[2]

$$LER = [1 - (10.7575 \log K - 8)^{-1}]/2 \qquad \textbf{(6)}$$

Thus, from **eqs. 5** and **6**:

$$FR = [1 - (10.7575 \log K - 8)^{-1}]/(2LE) \qquad \textbf{(7)}$$

We now have the tools to convert a given antenna, such as one in **table 1**, to a new (scaled) antenna where the element radii are changed; the new scaled antenna will perform exactly in the same way as the original antenna at the central design frequency ($F = 1$). However, the frequency-swept behavior of the (scaled) antenna, while qualitatively similar to the original, will show a broader or narrower bandwidth, depending on the change in element Q (see **eq. 2**).

The procedure is simple. For any given original element (subscript 1) we are given LE_1 and RO_1. The new (scaled) (subscript 2) radius is designated as RO_2. Compute the new (scaled) element length, LE_2:

$$K_1 = 1/RO_1 \quad ; \quad K_2 = 1/RO_2 \qquad \textbf{(8)}$$

$$FR_1 = [1 - (10.7575 \log K_1 - 8)^{-1}]/(2LE_1) \qquad \textbf{(9)}$$

$$X_1 = (215.15 \log K_1 - 160)\ (1/FR_1 - FR_1) \qquad \textbf{(10)}$$

Having calculated reactance (at $F = 1$), compute the value of FR_2 that will give the same value of X with the new element radius, RO_2:

$$X_2 = X_1$$

$$(1/FR_2 - FR_2) = X_1/(215.15 \log K_2 - 160) \equiv A \qquad \textbf{(11)}$$

$$FR_2 = [-A + (A^2 + 4)^{1/2}]/2 \qquad \textbf{(12)}$$

$$LE_2 = [1 - (10.7575 \log K_2 - 8)^{-1}]/(2FR_2) \qquad \textbf{(13)}$$

It is simple and convenient to set up the entire procedure (**eqs. 8-13**) on a small programmable calculator.

An example illustrates the results. Consider the antenna design for the six-element antenna in **table 1**;

table 3. Six-element Yagi; element length, LE (λ_0).

	reflector			driver			director		
	1	2	3	1	2	3	1	2	3
LE (λ_0)	0.49528	0.49489	0.49465	0.48028	0.47876	0.47785	0.44811	0.44431	0.44204
FR	0.97252	0.97042	0.96917	1.00289	1.00311	1.00325	1.07489	1.08090	1.08451
X (ohms)	30.40800	30.40800	30.40800	− 3.14700	− 3.14700	− 3.14700	− 78.58200	− 78.85200	− 78.85200

Note:
Column 1 $R_0 = 0.0005260\ (\lambda_0)$, from **table 1**
Column 2 $R_0 = 0.0008\ (\lambda_0)$
Column 3 $R_0 = 0.0010\ (\lambda_0)$

this would be a reasonable design for a 14.2-MHz antenna where $\lambda_0 = 69.3$ *feet* (21.13 meters) and where an RO of 0.0005260 (λ_0) would correspond to an element physical diameter of 0.875 inch (2.22 cm). This would be a reasonable dimension for a mechanically adequate element. Now, suppose that we would like an equivalent antenna for 28 MHz, where RO probably should be increased. The results of **eqs. 8-13** are shown in **table 3**. Note that the (scaled) changed values for LE are not wholly intuitive, because two things happen simultaneously. As RO increases the Q decreases, requiring a greater spread in resonant frequencies of reflector and director; however, at the same time, the resonant *physical* length, LER, also changes. Note that, if one scales the actual physical dimensions of boom length up by a factor, S (from, say, a smaller high-frequency antenna model), and the element radius dimension is not also scaled up equivalently, it is wrong, conceptually, to scale element length by the same factor S. Moreover, it is also wrong, in this case, to scale down element resonant frequency by the same factor, S. The only correct way to scale an antenna element is to design it (length and radius) to give the same electrical reactance.

element taper corrections

To this point, antenna designs and all antenna calculations have been made for strictly cylindrical elements, and the results will apply directly to most high-frequency (small) Yagi antennas where the general practice is to use cylindrical elements. However, for frequencies less than about 30 MHz, mechanical considerations usually require that the elements consist of one or more telescoping sections of tubing. At the lower frequencies (say ≤7 MHz), the Yagi antenna becomes gigantic, and it is no small mechanical engineering task to construct even a good element. Small diameters favor smaller wind forces, but these diameters are insufficiently rugged for long elements. It is, therefore, a practice to make these large elements of several telescoping sections. The largest-diameter section is clamped to the boom, and succeeding monotonically smaller-diameter sections make up the outer portions of the element. The resulting element taper can introduce a significant change in the required element length.

It's important to understand how to relate the actual detailed taper schedule of an element (diameters and lengths of all sections) to the equivalent length of a cylindrical (untapered) element having the same average or mean diameter. Equivalence is intended to mean that the resonant frequency and the Q are the same for the actual tapered element as for the equivalent cylinder.

To start, I shall introduce the concepts of element pipe inductance and pipe capacitance. Consider a cylindrical element of length s and radius RO as shown in **fig. 1**. A length coordinate, x, is defined with the origin at the center of the element and a related (angle) coordinate, θ, where $\theta = \pi x/s$. Note that electrical excitation of this element in the neighborhood of the resonant frequency, f, will produce a current and voltage distribution:

$$I_\theta = I_0 \sin(2\pi ft) \cos(\theta) \qquad \textbf{(14)}$$

and

$$V_\theta = V_0 \cos(2\pi ft) \cos(\theta) \qquad \textbf{(15)}$$

The electrical driving-point impedance of the element consists of a resistance (which is directly related to far-field energy radiation) and, of course, a reactance.

All reactance effects, including resonant frequency and electrical Q, are caused by near-field (non-radiating) energy storage. Energy storage occurs in two ways: the magnetic flux surrounding the current distribution in **eq. 14** and the electrical field produced by the voltage distribution in **eq. 15**. Note that at certain instantaneous times $(t = n/2f)$, the current everywhere is zero, and all stored energy resides in the electrical field. Similarly, at certain other times $(t = n/2f + 1/4f)$ the electric field vanishes, and all stored energy resides in magnetic flux.

As time progresses the (constant) total stored energy transfers back and forth between magnetic and electrostatic fields. This transfer or exchange frequency is, of course, the element resonant frequency. As a result of this complete nonradiative energy transfer, the peak or maximum magnetic stored energy must exactly equal the peak electrostatic stored energy. Note also that the resonant or natural exchange frequency must decrease as the total stored energy is increased.

Now, consider the effect of inserting an infinitesimal length of pipe (of the same radius, RO) into the element of **fig. 1** at the center $(x = 0.)$ The original

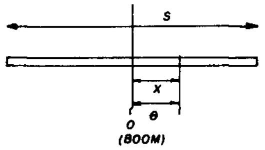

fig. 1. Coordinates of a single Yagi element. S = overall length, X is coordinate of length with zero at center (boom), and θ is corresponding angle $(\theta = \pi \cdot S/X)$.

(subscript 1) element driving-point reactance,[2] X, was shown to be:

$$X = (430.30 \log K - 320)\ (F/FR_1 - 1) \qquad \textbf{(16)}$$

where $K = \lambda/RO$

At the (original) resonant frequency, FR_1, the reactance vanishes; inserting an additional infinitesimal length of pipe, Δs, at $X = 0$ will change the resonant frequency to FR_2. At this new frequency the total reactance again vanishes. The added reactance due to the inserted pipe must be balanced by the original pipe reactance at the new frequency:

$$0 = (430.30 \log K - 320)(FR_2/FR_1 - 1) + 2\pi f \Delta L \qquad \textbf{(17)}$$

where f = actual (resonant) frequency
ΔL = increased inductance due to Δs.

The inserted pipe at $x = 0$ can produce only inductive effects (stored magnetic flux) since the electrical potential is strictly zero. Now, FR_2 is clearly related to FR_1 by the overall length(s) of the element:

$$FR_2/F_1 = s/(s + \Delta s) \qquad \textbf{(18)}$$

from which

$$\Delta L/\Delta s = (430.30 \log K - 320)/(S 2\pi f)$$

and

$$\Delta L/\Delta s = (430.30 \log K - 320)/(\pi c) \qquad \textbf{(19)}$$

where c is the velocity of light.

Thus, the addition of the small infinitesimal pipe section causes the element to behave just as though a pure series inductance were added. The effective inductance per unit length, which I designate by IND, is given by **eq. 19** and is easily expressed in conventional units as:

$$IND = (43.03 \log K - 32)(1.061 \times 10^{-8}) \quad \textit{henries/meter} \qquad \textbf{(20)}$$

From the simple model of a resonant circuit it is easy to relate the magnitude of voltage on the reactive components to magnitude of input current by:

$$|V_0| = |I_0(RQ)| \qquad \textbf{(21)}$$

with:

$$RQ = (215.15 \log K - 160) \qquad \textbf{(22)}$$

Now, consider extending the element in **fig. 1** by length Δs (of the same radius, RO) at its outer end ($x = s/2$). Here the current is zero so the small pipe increases only the electrostatic energy (capacitive effect). Since in this case **eq. 13** is still valid, the total increase in stored energy should be just the same as it was for insertion at $x = 0$. Therefore:

$$\Delta L(I^2)/2 = \Delta C(V^2)/2 \qquad \textbf{(23)}$$

where ΔC = the capacitance increase due to Δs at the element end.
ΔL = the increase in inductance due to Δs at the element center.

From **eqs. 21** and **23**:

$$\Delta C = \Delta L/(RQ)^2 \qquad \textbf{(24)}$$

Using **eqs. 22, 24**, and **19**:

$$\Delta s/\Delta C = (43.03 \log K - 32)(25\pi c/10) \qquad \textbf{(25)}$$

or in conventional units

$$\Delta s/\Delta C = 1/CAP \qquad \textbf{(26)}$$
$$= (43.03 \log K - 32)(2.356 \times 10^9)\ \textit{meters/farad}$$

where CAP = the capacitance per unit length.

Note that 1/CAP is directly related to IND, differing only in a constant multiplier.

Thus, we now can think of a cylindrical section of element pipe as contributing to element inductance (**eq. 20**) and element capacitance (**eq. 26**). Each contribution is a function of $K(\lambda/RO)$, and therefore RO, and each will depend on the current or voltage on the pipe section.

Let us now see what happens if a small section of pipe of length $\Delta B/2$ is first removed at a position x (or corresponding θ) and for symmetry also at $-x$ or $-\theta$ from the element shown in **fig. 1**. Now replace these removed sections with equal length sections ($\Delta B/2$) of larger radius RO. The overall length of the element remains s, but cylindrical "bumps" occur at X and $-X$. As a result of these bumps the stored energy of the system is changed and therefore the resonant frequency is changed. Designate the value of K for the original pipe as K_1 and for the short bumps as K_2. The contribution of the bump(s) to stored energy, W_2, will be

$$2W_2 = \Delta B\ [IND_2\ (I^2 \cos^2\theta) + CAP_2\ (V^2 \sin^2\theta)] \qquad \textbf{(27)}$$

The relationship of V at the end of the element to I at $x = 0$ is essentially unchanged from the original element, that is, $CAP_1 V^2 = IND_1 I^2$ (see **eq. 23**). Note also that (**eqs. 19** and **25**):

$$CAP_2/CAP_1 = IND_1/IND_2 \qquad \textbf{(28)}$$

so that **eq. 27** can be rewritten as

$$2W_2 = \Delta B\ [IND_2(I^2 \cos^2\theta) + (IND_1^2/IND_2)(I^2 \sin^2\theta)] \qquad \textbf{(29)}$$

Let us now find an equivalent length, $\Delta A/2$, of the original pipe which, when placed at the same positions as each of the bumps, contributes an equal stored energy.

$$2W_1 = \Delta A[IND_1 I^2(\cos^2\theta + \sin^2\theta)]$$
$$= \Delta A IND_1 I^2 = 2W_2 \qquad \textbf{(30)}$$

so that

$$\Delta A/\Delta B = (IND_2/IND_1)\cos^2\theta + (IND_1/IND_2)\sin^2\theta \qquad (31)$$

Now, for a longer section (longer bump) going from θ_1 to θ_2, the equivalent length of the original pipe can be easily calculated. Designate $IND_2/IND_1 \equiv m$, the length of the long bump as S_B, and the length of the original pipe, which gives equivalent stored energy, as S_A.

$$S_A/S_B = m\,\overline{\cos^2\theta} + (1/m)\,\overline{\sin^2\theta} \qquad (32)$$

The angular functions are to be averaged over the complete bump section. **Eq. 32** is easily integrated and averaged; the result is

$$S_A/S_B = (m + \frac{1}{m})/2 + (m - \frac{1}{m})F(\theta)/2 \qquad (33)$$

where

$$F(\theta) = (\sin 2\theta_2 - \sin 2\theta_1)/(2\theta_2 - 2\theta_1) \qquad (34)$$

with θ measured in radians.

We can now compute from a given element taper schedule (involving several sections with different pipe diameters) the equivalent lengths of sections of "standard" cylindrical pipe. The procedure is to first choose the "standard" cylinder that is expected to provide equivalent Q. This is, of course, the pipe size at the center of each half element; that is, the average or mean pipe size. Next, for each section of the tapered element, compute the starting θ_1 and ending θ_2. For each section compute m; it is easily derived from **eq. 20**, or

$$m = (43.03 \log K_2 - 32)/(43.03 \log K_1 - 32) \qquad (35)$$

From **eqs. 35** and **33** compute S_A/S_B, which, multiplied by the (tapered) section physical length, gives the equivalent section length of the standard pipe. Adding the lengths of all equivalent sections gives the overall length of the standard cylindrical element that should perform essentially the same as the chosen taper schedule.

Perhaps an example will illustrate the procedure. **Fig. 2** shows schematically a half element with five different sections whose physical diameters range from 1.250 inches (3.25 cm) at the boom ($x = 0$) to 0.500 inch (1.3 cm) at the outer end. Readers will recognize this taper schedule as one in common use (by Wilson) for a 14-MHz Yagi reflector antenna element. The middle pipe section, 7/8 inch (2.2 cm) in diameter, will represent the "standard" pipe. At a frequency of 14.2 MHz, $\lambda_0 = 831.76$ inches (21.13 meters), $RO = 0.0005260$, and $K_1 = 1901.17$. **Table 4** illustrates how to calculate the equivalent cylinder section lengths. For each section column 2 shows the actual physical length, S_B, column 3 shows pipe diameter, column 4 the K value, column 5 the value of m computed from **eq. 35**, column 6 values of θ_1, column 7 values of θ_2, column 8 values of $F(\theta)$ computed from **eq. 33**, and column 9 equivalent section lengths, S_A, also computed by **eq. 33**. Note that the overall actual length of the tapered half element is 215 inches (5.46 meters), whereas the overall length of the equivalent cylindrical standard 7/8 inch (2.2 cm) pipe is only 206.54 inches (5.25 meters). In other words, just due to the taper schedule alone the total (full length) tapered element must be made 16.9 inches (42.9 cm) longer than an equivalent cylinder! This taper correction is surprisingly large; it shows clearly that element length alone is a totally inadequate specification.

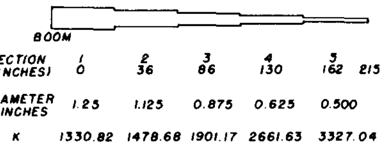

fig. 2. Diagram of a tapered half-element example.

The physical reason why the tapered element must be longer than an equivalent cylinder is that the inner (larger) sections have smaller inductance than a standard cylinder and therefore must be made longer; similarly, the outer (smaller) sections have smaller capacitance than the standard cylinder and must also be made longer. The taper correction will be quite

table 4. Equivalent length computations for element in fig. 2.

section	S_B (inches)	d (inches)	K	m	θ_1 degrees	θ_2 degrees	$F(\theta)$	S_A (inches)
1	36	1.250	1330.82	0.93890	0.000	15.070	0.95452	33.904
2	50	1.125	1478.68	0.95695	15.070	36.000	0.61449	48.696
3	44	0.875	1901.17	1.00000	36.000	54.419	-0.00718	44.000
4	32	0.625	2661.63	1.05764	54.419	67.814	-0.52851	31.102
5	53	0.500	3327.04	1.09586	67.814	90.000	-0.90300	48.835
	215							206.537

small if the taper is small, but quite significant if the taper is large.

In the derivation of taper correction calculations, I have assumed that radial "bumps" are treated as small perturbations on the strictly cylindrical case and that the current and voltage distributions are sinusoidal. Note that K values for the heavily tapered element of **fig. 2** differ from unity by only a few per cent; thus tthe calculation, even though made by a perturbation method, should be reasonably good. Moreover, the current distribution should still be reasonably sinusoidal over the tapered element. Nevertheless there may be some small inaccuracies in the overall calculation. It is important to note, however, that we are after a length *correction* of only a few per cent due to taper, and therefore some inaccuracy in the computation of the (small) correction is tolerable.

One further point merits elaboration. The procedure just outlined allows only a computation of cylinder equivalents from a given taper schedule; how may we compute a suitable taper schedule starting from a given cylinder? I have found that the simplest procedure is to initially specify all of the taper schedules from mechanical considerations, leaving as a variable only the length of the outermost section. Choose a guessed or estimated length for this section and compute the overall equivalent cylinder. It will generally miss the desired length by a differential length, Δ. One can now readjust the length of the outermost section by $-\Delta m$ and recalculate. One or two such iterations will bring the tapered element equivalent cylinder length into adequate agreement with the desired figure.

boom clamping correction

I now come to the subject of the boom-to-element mechanical clamping system and its effect on the element reactance and, hence, resonance. It is clear that a wide range of clamping systems are in common use; it is virtually impossible to make valid calculations for all varieties. Nevertheless there are two major kinds and it is helpful to understand them.

The first clamping system is simply to put the element directly through the (round) boom. In this construction a length of element equal to the complete boom diameter is replaced with the boom itself. Since this replacement occurs at a voltage node, we must determine the effective inductance of the replacement; once this is done it can be considered the first section of a tapered element from which an equivalent cylinder length can be calculated. I have not attempted a rigorous calculation of (boom) inductance; instead, I refer to the measurements of Viesbicke[9] in which his **fig. 10** shows that element length due to the presence of a (round) boom should be increased by about 0.7 the diameter of the boom. This is tantamount to saying that the inductance of the boom section of the element is very low compared with normal element inductance; physically this is an expected result. The low inductance, of course, is due to the blockage of magnetic flux by the boom.

The second clamping system is much more widely used since it permits easier element maintenance and replacement. In this sytem either a flat, metal, rectangular plate or an angle bracket is interposed between element and boom; two U-bolts fasten the boom to plate or bracket and two more U-bolts fasten the element to plate or bracket. The U-bolts may also use saddles or cradles, which are mechanically better and which further tend to separate boom and element. For this clamping system we wish to know the inductive effect of the boom itself and more importantly the inductive effect of the plate or bracket. I have found experimentally that for this clamping system the boom itself has remarkably little effect. Even though the (round) boom and (round) element are in physical contact, the element length should be increased by only 6 per cent of the boom diameter; this small correction rapidly disappears as the element is spaced away from the boom (even by a small amount). The reason this result is so different from the through-the-boom result is the relative ease with which the magnetic flux (which results from element current flow) can squeeze between boom and element, especially if there is any gap between them.

The correction in length due to the mounting plate or bracket is readily calculable. The method is to first calculate the equivalent radius of the element plus bracket (which produces the same inductance) and second to use this equivalent radius as the first (short) section of a taper design. The theory for equivalent radii of single and multiple parallel conductors is given by Mushiake and Uda.[10] In their notation the equivalent radius, ζ_ϵ, of a flat thin plate of total width, a, is simply:

$$\zeta_\epsilon = a/4 \qquad \textbf{(36)}$$

and that for a right-angled bracket of width a and b is given by a rather complicated expression, which depends only slowly on the ratio b/a. For ratios between 0.3 and 1.0 a good approximation (error <5 per cent) is:

$$\zeta_\epsilon \cong 0.2(a+b) \qquad \textbf{(37)}$$

Mushiake and Uda show that for two parallel conductors, it is possible to calculate the equivalent radius of the combination. If a_1 and a_2 are the lengths of the peripheries of the cross sections, ζ_1 and ζ_2 the equivalent radii of the two conductors, d_m the mean distance between them, and ζ_ϵ the equivalent radius of the combination of both conductors, then

$$\log \zeta_\epsilon = (a_1^2 \log \zeta_1 + a_2^2 \log \zeta_2 + 2a_1 a_2 \log d_m)/(a_1 + a_2)^2 \qquad (38)$$

Eqs. 36 *and* **38** permit a calculation of the equivalent radius of an element which is proximate to a plate; similarly, **eqs. 37** and **38** provide a way of calculating the equivalent radius of an element proximate to an angle bracket. To check this method of calculation, I have determined the experimental detuning effect of a plate just touching a 1 inch (2.54 cm) diameter element resonant at 46 MHz. **Table 5** shows both theoretical and experimental results for two different plates. These experiments were not particularly accurate because the resonant frequency is difficult to measure accurately; nevertheless the agreement of theory and experiment within estimated experimental accuracy is gratifying.

Note that element length corrections due to a proximate mounting plate or bracket can easily be as much as 10 percent of plate length. These corrections are not especially large in practice, but should be made wherever there is a relatively large boom-to-element clamping system.

scaling and taper example

It may be helpful to show how to specify a good three-element beam starting with the cylindrical design in **table 1**. I shall go through necessary scaling, then taper schedule calculations for element length(s), and finally apply reasonable boom clamping and boom corrections; this procedure is used to specify a 14.2-MHz beam, a 21.3-MHz beam and a 28.5-MHz beam.

First I choose an average cylinder size that is sufficiently strong. I shall assume that the final element is made of aluminum tubing such as 6061-T6 with seamless 0.059 inch (1.5 mm) wall thickness. For all three bands I choose a cylinder size of 0.875 inch (2.2 cm) OD, although for 28.5 MHz a slightly smaller size is probably permissible. Second, I choose a convenient taper schedule which is easily made from standard 12-foot (3.7-meter) lengths, leaving the length of the outermost section to be adjusted for correct overall length. The sections of seamless tubing (except the last section) are slit back about 3 inches (7.6 cm) at the outer ends (I use one slit only), and a common stainless steel hose clamp fastens sections together. Tubing overlap of about 8 inches (20 cm) gives good joint strength. For 14.2 MHz, the second section is a full 12-foot (3.7-meter) section, over which is slid the shorter first section; this procedure gives added (central) strength and improves the ease of clamping with U-bolts and saddles. For 21.3 and 28.5 MHz this extra inner section is unnecessary.

table 5. Increase in resonant frequency due to a proximate plate. Element radius is 0.50 inch (1.3 cm) and length produces resonance at 46 MHz.

	plate dimensions		change in resonant frequency	
plate	length (inches)	width (inches)	per cent theory	per cent expected
1	4.5	3.625	0.304	0.325 ± 0.1
2	6.0	4.000	0.530	0.521 ± 0.1

Table 6 shows the specifications for these tapered half elements where x_1 and x_2 represent the start (inner) and end (outer) positions (in inches) of each section. Note that the tubing requirements for all three elements are shown in 12-foot (3.7-meter) lengths.

table 6. Taper schedules (half elements), for the 3-element Yagi.

section	*d* (inches)	x_1 (inches)	x_2 (inches)	3-elements tubing lengths (12 feet)
		14.2 MHz		
1	1.125	0	24	1
2	1.000	24	72	3
3	0.875	72	136	3
4	0.750	136	176	2
5	0.625	176	<215	2
		21.3 MHz		
1	0.875	0	72	3
2	0.750	72	112	2
3	0.625	112	< 145	2
		38.5 MHz		
1	0.875	0	72	3
2	0.750	72	< 105	2

Third, for these three cases it is necessary to scale the original design of **table 1** to use the desired average cylinder size. **Table 7** shows the scaled cylinder lengths (in λ_0 for all three beams using scaling techniques discussed previously.

We are now ready to compute the effect of taper schedule. For the 14.2-MHz element(s), **table 8** shows the flow of calculations; x_1 and x_2 (inches) show the start and finish of each section. First, a trial guess at the overall reflector length is made; I guessed 212 inches (5.38 meters) in this case. For each section K, m, $F(\theta)$ and S_A (in inches) are calculated by the previously described technique. Note that the sum of all cylinder equivalents S_A is 207.63 inches (5.27 meters); what was desired was 207.11 inches

(5.26 meters). This was a lucky guess; however, a small correction should be made to section 5. This correction is *m* times the needed cylinder correction. Next in **table 8** is shown a second reflector calculation after the correction is made; note that the new cylinder equivalent is exactly what was desired. Thus the overall length of the half element (last x_2) is 211.45 inches (5.37 meters).

By using the correction procedure, the next calculation derives the overall length of the driven element and a small iteration sets it (last x_2) at 208.0 inches (5.28 meters). The same procedure is used for the director, whose overall length (last x_2) is 198.83 inches (5.05 meters). **Table 9** shows exactly the same calculation procedure for the 21.3-MHz beam elements, and **table 10** shows the results for the 28.5-MHz beam elements.

We are now ready for the final small boom and boom clamp corrections. For this purpose I assumed the elements are U-bolted with saddles to flat plates, which in turn are U-bolted with saddles to the boom. Boom diameters are assumed to be 3 inches (7.6 cm) OD (14 MHz) and 2 inches (5.1 cm) OD (21 and 28 MHz). Full plate dimensions are assumed to be 6 inches (15.2 cm) wide and 8 inches (20.3 cm) long (14 MHz); 5 inches (12.7 cm) wide and 6 inches (15.2 cm) long (21 MHz); and 4 inches (10.2 cm) wide 4 inches (10.2 cm) long (28 MHz). These plates reduce central pipe inductance and thus cause an electrical shortening of the half element. This shortening is easy to calculate by techniques previously described. It amounts to about 0.66 inch (1.7 cm) (14 MHz); 0.44 inch (1.1 cm) (21 MHz); and 0.24 inch (0.6 meters) (28 MHz).

table 7. Scaling computations (3-element beam of table 1).

Freq. (MHz)	λ_0 (inches)	*d* (inches)	*K*	*RO* (λ_0)	*R*	*DR*	*D*
14.2	831.76	0.875	190.17	0.0005260	0.49801	0.48963	0.46900
21.3	555.81	0.875	1270.42	0.0007871	0.49790	0.48916	0.46765
28.56	414.42	0.875	947.25	0.001056	0.49769	0.48819	0.46490

table 8. Taper calculations at 14.2 MHz.

	SEC.	*d* (inches)	x_1 (inches)	x_2 (inches)	*K*	*M*	*F(θ)*	S_A (inches)
R_{TRIAL}	1	1.125	0.	24.	1478.68	0.95695	0.97905	22.990
λ_0 = 831.76 inches	2	1.000	24.	72.	1663.51	0.97713	0.74164	47.189
CYLINDER	3	0.875	72.	136.	1901.17	1.00000	0.02854	64.000
RO = 0.0005260 (λ_0)	4	0.750	136.	176.	2218.03	1.02641	−0.66514	39.320
LE = 0.49801 (λ_0)	5	0.625	176.	212.	2661.63	1.05764	−0.95324	34.133
HALF LENGTH 207.11 inches								207.632
R	1	1.125	0.	24.	1478.68	0.95695	0.97894	22.989
	2	1.000	24.	72.	1663.52	0.97713	0.74038	47.190
	3	0.875	72.	136.	1901.17	1.00000	0.02467	64.000
	4	0.750	136.	176.	2218.03	1.02641	−0.66945	39.316
LE = 0.49801 (λ_0)	5	0.625	176.	211.45	2661.63	1.05764	−0.95540	33.609
HALF LENGTH 207.11 inches								207.104
x_2 *LAST* = 211.45 inches								
DR	1	1.125	0.	24.	1478.68	0.95695	0.97820	22.990
	2	1.000	24.	72.	1663.52	0.97713	0.73174	47.200
	3	0.875	72.	136.	1901.17	1.00000	−0.00145	64.000
	4	0.750	136.	176.	2218.03	1.02641	−0.69796	39.286
LE = 0.48963 (λ_0)	5	0.625	176.	207.8	2661.63	1.05764	−0.96192	30.135
HALF LENGTH 203.63 inches								203.611
x_2 *LAST* = 208.0 inches								
D	1	1.125	0.	24.	1478.68	0.95695	0.97619	22.992
	2	1.000	24.	72.	1663.52	0.97713	0.70843	47.226
	3	0.875	72.	136.	1901.17	1.00000	−0.06997	64.000
	4	0.750	136	176.	2218.03	1.02641	−0.76731	39.214
LE = 0.46900 (λ_0)	5	0.625	176.	198.75	2661.63	1.05764	−0.97859	21.538
HALF LENGTH 195.00 inches								194.97
x_2 *LAST* = 198.83 inches								

table 9. Taper calculations at 21.3 MHz.

	SEC.	d (inches)	x_1 (inches)	x_2 (inches)	K	M	$F(\theta)$	S_A (inches)
R_{TRIAL}	1	0.875	0.	72.	1270.42	1.00000	0.61831	72.000
λ_0 = 555.81 inches	2	0.750	72.	112.	1482.16	1.02836	−0.45812	39.503
CYLINDER	3	0.625	112.	140.	1778.59	1.06191	−0.93549	26.476
RO = 0.0007871 (λ_0)								137.979
LE = 0.49783 (λ_0)								
HALF LENGTH 138.35 inches								
R	1	0.875	0.	72.	1270.41	1.00000	0.62020	72.000
	2	0.750	72.	112.	1482.16	1.02836	−0.45319	39.509
LE = 0.49783 (λ_0)	3	0.625	112.	140.4	1778.59	1.06191	−0.93404	26.857
HALF LENGTH 138.35 inches								138.366
x_2 $LAST$ = 140.2 inches								
DR	1	0.875	0.	72.	1270.42	1.00000	0.60720	72.000
	2	0.750	72.	112.	1482.16	1.02836	−0.48664	39.471
LE = 0.48884 (λ_0)	3	0.625	112.	137.7	1778.50	1.06191	−0.94368	24.289
HALF LENGTH 135.65 inches								135.760
x_2 $LAST$ = 137.6 inches								
D	1	0.875	0.	72.	1270.42	1.00000	0.57571	72.000
	2	0.750	72.	112.	1482.16	1.02836	−0.56284	39.386
LE = 0.46675 (λ_0)	3	0.625	112.	131.65	1778.59	1.06191	−0.96375	18.547
HALF LENGTH 129.71 inches								129.933
x_2 $LAST$ = 131.4 inches								

table 10. Taper calculations at 28.5 MHz.

	SEC.	d (inches)	x_1 (inches)	x_2 (inches)	K	M	$F(\theta)$	S_A (inches)
R λ_0 = 414.42 inches	1	0.875	0	72	947.25	1.00000	0.37839	72.000
CYLINDER	2	0.750	72	104	1105.12	1.02988	−0.85138	31.209
RO = 0.001056 (λ_0)								103.209
LE = 0.49769 (λ_0)								
HALF LENGTH 103.13 inches								
DR	1	0.875	0	72	947.25	1.00000	0.35882	72.000
	2	0.750	72	101.89	1105.12	1.02998	−0.86433	29.140
LE = 0.48819 (λ_0)								101.140
HALF LENGTH 101.16 inches								
x_2 $LAST$ = 101.90 inches								
D	1	0.875	0	72	947.25	1.00000	0.31133	72.000
	2	0.750	72	97.08	1105.12	1.02998	−0.89377	24.429
LE = 0.46490 (λ_0)								96.429
HALF LENGTH 96.33 inches								
x_2 $LAST$ = 96.97 inches								

Thus, to compensate for the boom clamp, each half element should be lengthened by an equivalent amount; it should be further lengthened by the empirical 1/16 boom radius described previously.

With all these corrections, the overall physical length of each half element is shown in **table 11**. None of these taper schedules is severe; therefore, the actual element lengths are not a great deal longer than the cylinder lengths shown in **tables 8**, **9**, and **10**; nevertheless the differences are there. Also shown in **table 11** is the boom position x_B for each element expressed both in λ_0 and in inches. Although I have not tested any of these particular three-element beams experimentally, I am confident that their performance will be excellent and, moreover, they all should be easy to construct.

summary

Let me now summarize briefly the results of the entire Yagi design series.

1. A computational methodology was developed and validated[2,3] that allows the important Yagi antenna

table 11. Overall element half lengths (in inches) and boom positions (in λ_0 and inches); 3 element beams with taper schedules of tables 8, 9 and 10.

(MHz) freq	element	initial taper (inches)	clamp (inches)	boom (inches)	final length (inches)	boom location, x_B (λ_0)	(inches)
14.2	R	211.45	0.66	0.09	212.20	0	0
	DR	208.00	0.66	0.09	208.75	0.15	124.7
	D	198.83	0.66	0.09	199.58	0.30	249.5
21.3	R	140.2	0.44	0.06	140.70	0	0
	DR	137.6	0.44	0.06	138.10	0.15	83.4
	D	131.4	0.44	0.06	131.90	0.30	166.75
28.5	R	103.91	0.24	0.06	104.21	0	0
	DR	101.90	0.24	0.06	102.30	0.15	62.2
	D	96.97	0.24	0.06	97.27	0.30	124.3

properties to be computed. Such computations produce results which are judged accurate to a few per cent; such an accuracy probably exceeds the accuracy of state-of-the-art experimental techniques.

2. Computations have been made throughout the series which have led to many new insights to Yagi antenna behavior. Among them are:

a. Simplistic designs (all elements spaced equally along the boom, and all directors of equal length) are as good as any other design for the same boom length as long as the boom is shorter than one wavelength.[4]

b. Yagi forward gain basically depends only on boom length (in λ); it is essentially independent of number of elements as long as element spacing along the boom is not too large.[4] Conceptually, the boom can be considered an aperture illuminated in a quasi-uniform way by the discrete elements. The illumination produces a diffraction pattern (the radiated antenna pattern) whose details are controlled by the precise illumination schedule.

c.Yagi F/B ratio is (naturally) best when the diffraction pattern has a null in the back direction. This occurs approximately when the boom length is an odd multiple of $\lambda/4$.

d. A procedure exits whereby a Yagi antenna having four or more elements and roughly favorable boom length can be fine tuned by slight changes in element positions on the boom to give an indefinitely high F/B ratio; this astronomical F/B (that is, >120 dB) exists only at a single frequency. It occurs due to vectorial cancellation of individual element contributions and is equivalent in concept to a notch frequency filter which is carefully adjusted to give an exceptionally deep notch.[5]

e. Yagis, quads and quagis all behave alike qualitatively. Conceptually a quad can (if properly adjusted) have a somewhat higher gain (a fraction of one dB) than a single Yagi; for horizontal polarization the increased gain comes about from slightly increased vertical directivity. This conceptual advantage may be eroded in practice by the difficulty of experimental quad adjustment compared with the accurate construction of a Yagi to a valid computed design.[6]

f. The gain and impedance of any equilateral quad loop is *strictly independent* of the position of the feed point.

g. Ground effects are extremely important and lead directly to preferred antenna heights (1 to 2λ) with corresponding preferred radiation elevation angles.[7]

h. Stacking (horizontally polarized) Yagis vertically over ground is very effective if the top Yagi is sufficiently high (1 to 3λ). Stacking does result in significant mutual coupling effects, which can degrade normally expected performance, especially F/B ratio.[8]

i. A new method is suggested for raising the radiation acceptance angle for stacked beams. This method uses phase reversal for one of two antennas in a stack; the apparent advantage is the retention of stack gain at the higher angles.[8]

j. Fine tuning, or beam optimization, for high F/B ratio depends on the ultimate end use. Designs are different for free-space conditions, a single Yagi antenna over ground, and Yagi antennas to be used in a stack.[8]

3. Practical computation procedures are provided in this article for *scaling* a given design to use elements of different radii, for *length corrections* due to element taper schedule, and for length corrections due to mechanical boom-to-element clamps.

4. The entire series provides a way for anyone to make a Yagi antenna system having high computed *performance, starting from his own computed designs,* or starting from designs which have been suggested in this series. Moreover, it is also shown in this article how to make a Yagi antenna which will accurately emulate the performance of any existing Yagi design; the performance will be just as good (or just as bad) as the emulated design.

final comments

In the development and exposition of this series of related articles, which I found both technically challenging *and requiring considerably more effort than* originally anticipated, I have attempted to proceed from basic electromagentic theory to a model of a Yagi antenna system which could ultimately be used in a practical way. All of the required steps and tools have been described. However, along the way I have noticed a number of areas in which further work by interested people could be very helpful. Among these are the following:

1. Valid theoretical treatment of mutual impedance where element length is not $\lambda/2$, and where the current distribution is not sinusoidal but consistent with the element function and environment. A particularly difficult question exists with regard to the imaginary part of this impedance at small distances.

2. Valid theoretical treatment of the screening effect of closely adjacent dipoles on the electric field normally present at a given dipole.

3. Valid theoretical treatment of the mutual coupling between quad loops, especially including the imaginary component of coupling at all loop distances.

4. Valid theoretical treatment of the reactance of a full quad loop as a function of its length (perimeter) in the neighborhood of λ.

None of these tasks is easy. All require good physics followed by tractable mathematics. Moreover, even if "solutions" are claimed, they must be viewed with some suspicion until *accurate* experimental results confirm their validity.

In addition to these theoretical tasks, it would be extremely helpful if good experiments could be made in one or more of the following areas:

1. Experiments on model Yagi antennas, similar to those reported by NBS[9], but carried out with improved instrumentation and especially improved control of the physical environment. Such experiments could be exceedingly useful in attempting to validate not only the models I have used, but improved models which I am sure will occur in the future.

2. Find a way to better characterize real (rough, contoured, or both) ground sites. Such characterization *should also include the electromagnetic properties of* ground. The objective of such work is to provide valid models for a wide spectrum of real-world sites; the use of these models should lead to better understanding of ground effects and perhaps methods for minimizing ground problems.

3. From (flat) ground sites at several magnetic latitudes measure the (statistical) arrival angles of incoming signals. Such measurements should be made at a number of widely separated useful frequencies; at each frequency the results should be correlated with the measured state of the ionosphere. These measurements should be made, not only over a yearly cycle, but over at least one complete solar cycle. Only in this way will a real understanding of the relevant behavior be reached. The end result of this understanding is, of course, to allow specifications for needed incoming arrival angles and hence specifications for optimum antenna height(s) and stacking arrangements.

It is clear that all of these suggestions require an uncommon competence and dedication, as well as the development of sophisticated experimental instrumentation. They also require a great deal of effort.

In the meantime I am convinced that the tools now available will not only permit the design of improved antenna systems, but in many aspects also permit a practical design that is unlikely, even in principle, to be significantly improved.

It is my wish that many readers will construct these superior Yagi antenna systems, make meaningful measurements of their properties, and report results accurately in the literature.

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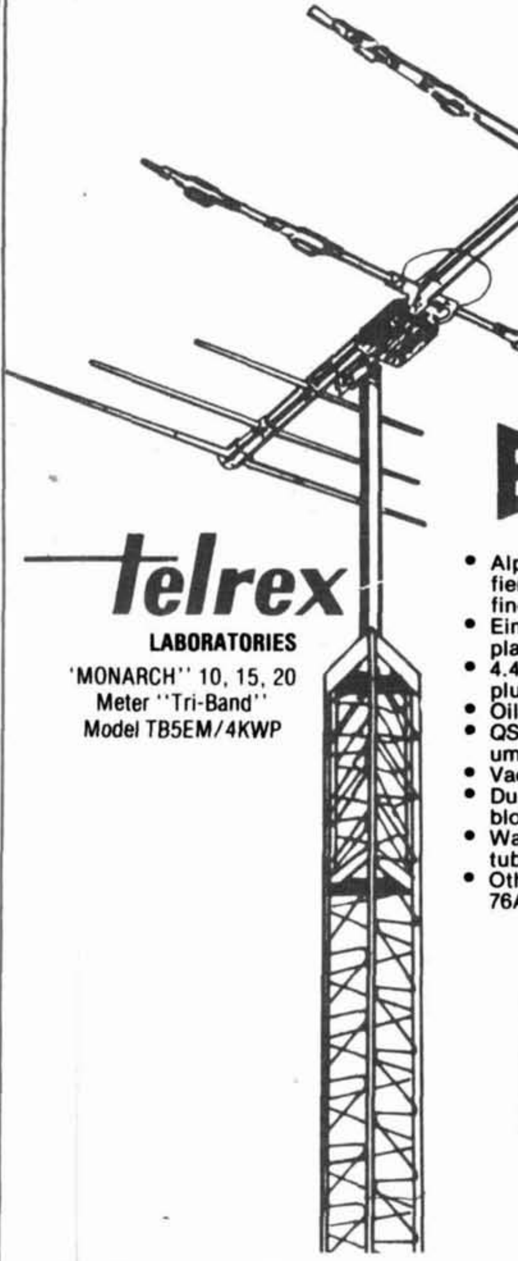

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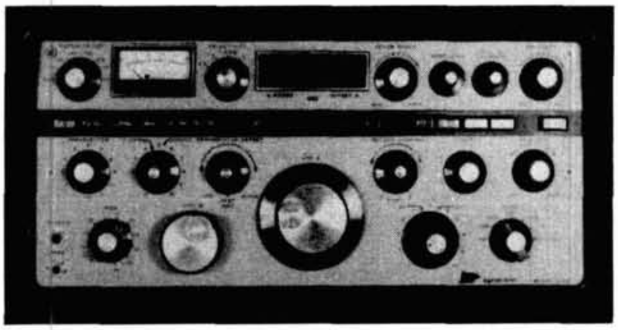

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In my mobile a TS-120S drives a modified HA-14 amplifier. The TS-120S is powered from the standard 55-ampere automotive system. To power the HA-14 linear, I use a three-phase alternator-powered supply.

high-voltage mobile supply

A three-phase Leece-Neville alternator is used as a primary source, which I bought for $10. It has a rating of 7 volts at 60 amperes. The alternator circuit is shown in **fig. 1**.

I mounted the alternator on the car and used a belt drive from the crankshaft pulley on the engine. (It takes a tight belt to prevent slippage under maximum load.)

The high-voltage supply (**fig. 2**) is a three-phase delta configuration with voltage from each phase applied to a full-wave voltage doubler.

The outputs of each voltage doubler are connected in series to obtain 2400-2600 Vdc. I used three surplus transformers with 12-volt primaries and 170-volt secondaries. Other transformers can be used, and if the turns ratio is correct, voltage doublers aren't necessary. Regular 12-volt, 60-Hz filament transformers with a 220-volt winding can be used.

Applying 25-30 volts to a transformer rated at 12 volts can be alarming but because of the alternator output frequency, the impedance is acceptable, and the transformers will work well without any heating.

Although the alternator is rated at 7 volts output, the output voltage is dependent on the regulator. The regulator (**fig. 3**) sets the alternator output at 25-30 volts. The alternator works quite efficiently at the elevated output voltage. I've had no problems while running it this way. The field current must be taken into account, however, and the 5.6-ohm resistor (**fig. 1**) limits it to a safe value. I've test-loaded this power system at approximately 2400-2600 watts with no problems.

regulator

The regulator is a modified version of a circuit published several years ago. No battery is needed in this power system. Several regulator designs were tried and worked well; this is the one I like best. The alternator will usually self-excite when turned on, but if not a momentary push button switch will do it (S2, **fig. 1**).

This power system has been trouble-free and very dependable. Since the high power drain doesn't affect the automotive power system or its battery, rundown battery problems don't exist. I can run full power with this mobile setup for hours on end with no overheating or other problems. The limitation, of course, is that the engine must run at idle rpm or more to operate the linear.

installation

I mounted the power supply and linear amplifier in the car trunk and the regulator under the hood away from engine heat. If the antenna is bumper mounted, it *must* be well grounded. While transmitting with this high power, allow no one to touch the antenna — severe burns will result. Even the outside of the car can give rf burns.

While pulling into my drive one night I was surprised to see what I thought was lightning on a clear night.

By Don Winfield, K5DUT, 6080 Anahuac Avenue, Fort Worth, Texas 76114

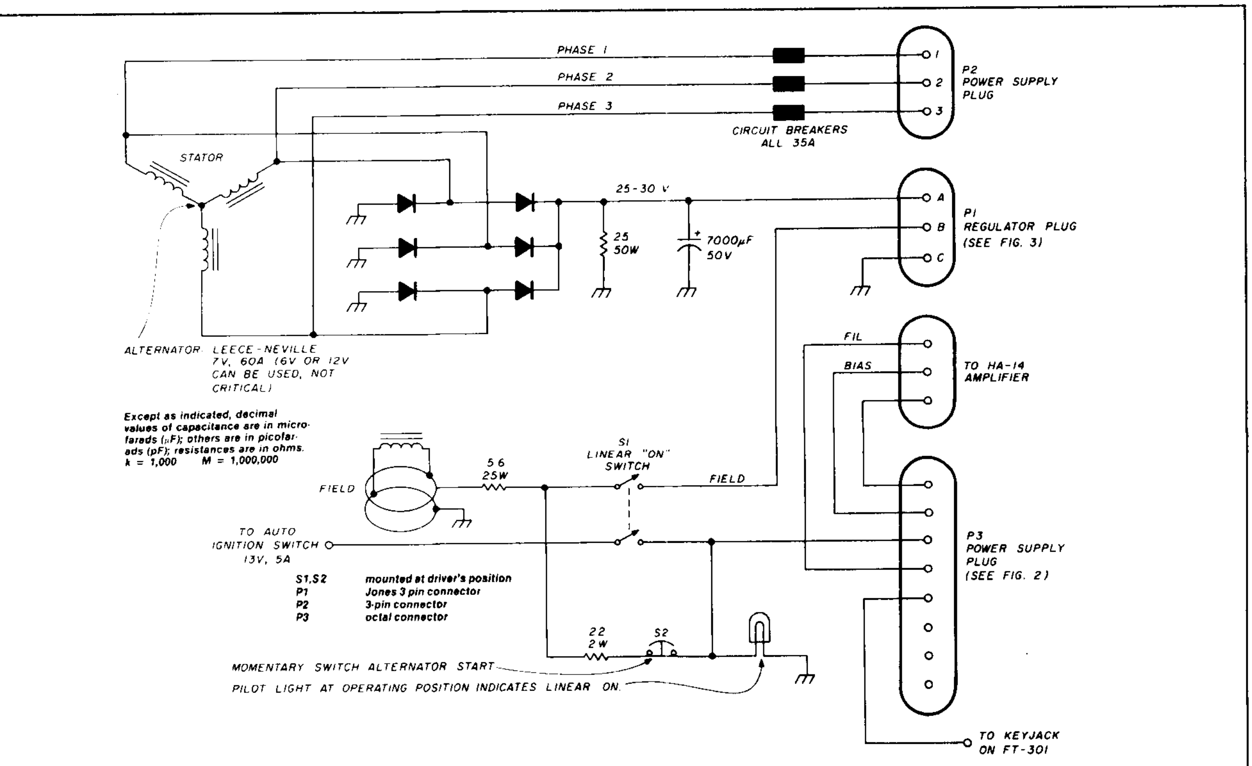


fig. 1. Primary power source for the mobile linear amplifier uses a three-phase Leece-Neville alternator. Voltage from each phase is applied to a fullwave voltage doubler.

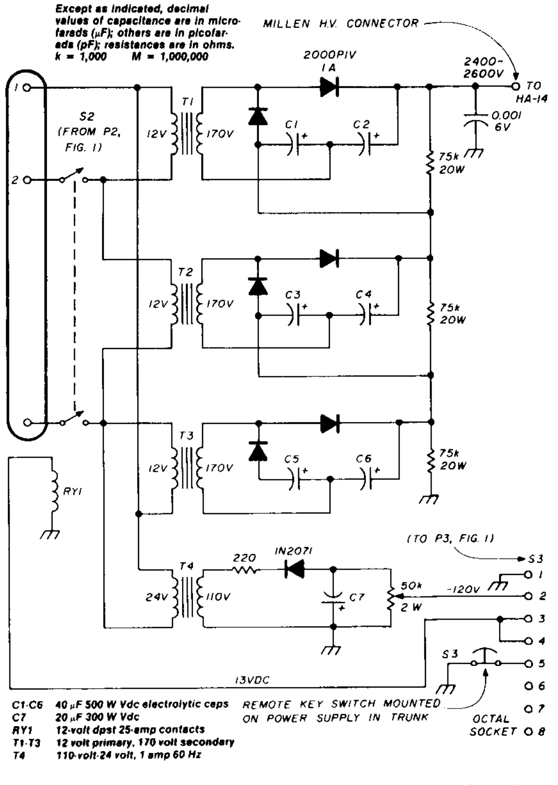
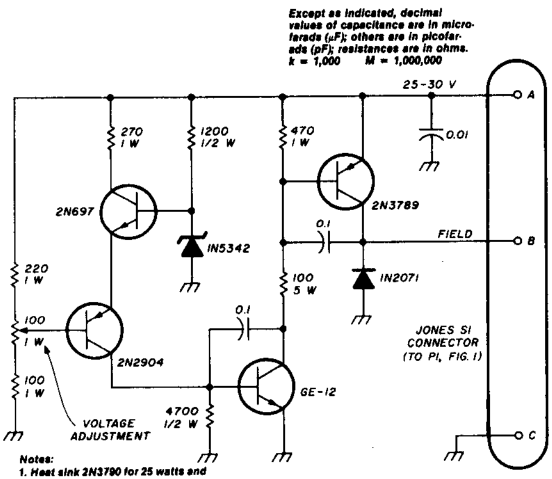


fig. 2. High-voltage supply. Outputs of each voltage doubler are connected in series to provide 2400-2600 Vdc for the mobile final amplifier. The system has been test loaded at 2400-2600 watts.



Notes:
1. Heat sink 2N3790 for 25 watts and GE-12 for 5 watts.
2. Set voltage pot for 2500-2600 Vdc output from power supply.
3. Transistor type is not critical. Similar types can be substituted.

fig. 3. Power-supply regulator, which sets alternator output at 25-30 volts. The 5.6-ohm resistor in series with the alternator field (fig. 1) limits field current to a safe value.

The top of the antenna was touching low tree limbs as I transmitted, and the damp limbs drew arcs from the antenna, with one of the limbs smoldering and on *fire*. I've since learned to shut down when under trees with low limbs.

results

After everything is in place and working, what kind of results can be expected from a kW in the car?

DX stations such as D4, ZS, EL2, 6W8, XT2, H44, VR8, and TR8 have been worked with 5-9 or better reports from the 20-meter mobile. I've enjoyed many contacts with DX friends such as ZS6DN, F3EG, and VR3AR while driving to and from work. In my case, that's a 35-minute trip on the interstate usually with light traffic. Just right for a little mobile DXing.

During peak band conditons, reports are routinely received from both coasts of 30-40 dB over S9 and occasionally "pegging the S meter." Numerous comments such as, "You're too strong to be a mobile," have occurred. I usually honk the horn to convince the doubters.*

Other bands are worked also, and, what with the excellent conditions during the fall of 1979, the 10- and 15-meter band propagation was so good that the mobile was just as good as a fixed station. Many DX stations were worked on first call on these bands in pile-ups during this time. During the winter months, 75 meter DX is worked routinely into most areas of the world. I use CW from the mobile also. A memory keyer is a great help.

The biggest limitation to DX work from a mobile is the ability to receive. On today's crowded bands, with the nondirectional vertical, interference is a problem, as is noise while operating mobile in populated areas. Noise blankers help a great deal. The most common problem with the mobile occurs when a CQ is called. The average ham expects a mobile not to be too strong, and when he hears one calling CQ and answers him, he finds it hard to believe that the mobile can't copy his signal on a simple antenna.

I've enjoyed this mobile for about 1½ years and can recommend mobile DXing as another means of enjoying ham radio. For a mobile station to be able to jump into a huge pileup on a rare station on 20 meters and come up with a contact is something that apparently never ceases to amaze the Big Guns at their multikilowatt stations with huge antennas scraping the clouds.

I'll be glad to help in planning your super mobile DX station on the receipt of a large, self-addressed stamped envelope.

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*Using Morse code, of course. **Editor.**

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SPECIFICATIONS:

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	Less than 50 MV to 500 MHz
Resolution:	0.1 Hz (10 MHz range)
	1.0 Hz (60 MHz range)
	10.0 Hz (600 MHz range)
Display:	9 digits 0.4" LED
Time base:	Standard-10.000 mHz, 1.0 ppm 20-40°C.
	Optional Micro-power oven-0.1 ppm 20-40°C
Power:	8-15 VAC @ 250 ma

7 DIGITS 525 MHz $99.95 WIRED

SPECIFICATIONS:

Range:	20 Hz to 525 MHz
Sensitivity:	Less than 50 MV to 150 MHz
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Resolution:	1.0 Hz (5 MHz range)
	10.0 Hz (50 MHz range)
	100.0 Hz (500 MHz range)
Display:	7 digits 0.4" LED
Time base:	1.0 ppm TCXO 20-40°C
Power:	12 VAC @ 250 ma

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SPECIFICATIONS:

Range:	1 MHz to 500 MHz
Sensitivity:	Less than 25 MV
Resolution:	100 Hz (slow gate)
	1.0 KHz (fast gate)
Display:	7 digits, 0.4" LED
Time base:	2.0 ppm 20-40°C
Power:	5 VDC @ 200 ma

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	Less than 150 mv to 600 MHz
Resolution:	1.0 Hz (60 MHz range)
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Display:	8 digits 0.4" LED
Time base:	2.0 ppm 20-40°C
Power:	110 VAC or 12 VDC

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PRICES:

CT-50 wired, 1 year warranty	$159.95
CT-50 Kit, 90 day parts warranty	119.95
RA-1, receiver adapter kit	14.95
RA-1 wired and pre-programmed (send copy of receiver schematic)	29.95

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PRICES:

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DM-700 Kit, 90 day parts warranty	79.95
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BP-3, Nicad pack +AC adapter/charger	19.95
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The DM-700 offers professional quality performance at a hobbyist price. Features include; 26 different ranges and 5 functions, all arranged in a convenient, easy to use format. Measurements are displayed on a large 3½ digit, ½ inch LED readout with automatic decimal placement, automatic polarity, overrange indication and overload protection up to 1250 volts on all ranges, making it virtually goof-proof! The DM-700 looks great, a handsome, jet black, rugged ABS case with convenient retractable tilt bail makes it an ideal addition to any shop.

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amplitude compandored sideband

Narrowband techniques for vhf mobile communications

It is obvious to most observers in larger metropolitan areas (New York, Los Angeles, Chicago, San Francisco) that saturation is beginning to occur on the 2-meter band. Even with the extra megahertz provided by the added repeater sub-band, with a total possible repeater population of 60 or so machines above 146 MHz, and 20 or so in the 144-145 MHz region, there are times when a ham population of 10,000 or more in such regions taxes these systems to their limit. Timers of 60, 40, or even 30 seconds are *not* really the answer.

If hams have been experiencing a problem, consider the plight of commercial users of vhf/uhf. It has been impossible for some time to obtain vhf licenses in many areas, and uhf channels are in short supply as well. Common carrier multiplexing schemes and/or 900-MHz channels have been proposed, but individual vhf/uhf or semi-shared channels have many advantages to the ultimate user, not the least of which is long-term cost.

Sideband use on vhf has long been used by Radio Amateurs (and the military). With the recent introduction of multi-mode 2-meter rigs, a surge in interest and activity has been sparked using this mode. Below fm threshold, SSB provides distinct advantages in sensitivity and range capabilities. Unfortunately, many of the convenience featues of fm operation do not work with our current sideband transceivers, and the signal-to-noise ratio on stronger signals, as well as the audio bandwidth and quality, do not match the better fm rigs.

amplitude compandored sideband

Recent developments promise to change the situation. In his report to the FCC after an extensive two-year research program into narrowband techniques for vhf land mobile,[1] Dr. Bruce Lusignan of Stanford University's Satellite Planning Center has come to some very interesting conclusions. By modulating a standard single-sideband transceiver with specially processed audio and processing the recovered audio through a similar system on the receive end, equal or even better performance can be obtained than when using NBFM. Because less than one-fifth the spectrum is required for equivalent channel-to-channel protection, five times as many stations can occupy the same spectrum space.

ACSB, or amplitude compandored sideband, combines several common techniques especially tailored for SSB. The system, developed by Dr. Lusignan in conjunction with Dr. Fred Cleveland of the University of the Pacific and VBC, Incorporated, features 4:1 amplitude compandoring, a pilot subcarrier system, and 12-dB/octave pre-emphasis/de-emphasis. The resultant ACSB system provides:

1. 50-70 dB adjacent channel protection using 5-kHz channels (as opposed to 20-25 kHz spacing for fm).

2. 10-dB power advantage due to both processing and bandwidth.

3. Automatic frequency locking and carrier identification.

4. Very rapid AGC (20 Hz) to greatly reduce mobile flutter.

5. A degree of quieting performance that, combined with its greater sensitivity, equals or exceeds normal fm.

6. Extended, reliable range by a factor of two, up to

By James Eagleson, WB6JNN, 280 Manfre Road, Watsonville, California 95076

about 25 miles (40 km), limited to a factor of 1.5 times only by earth curvature beyond this distance.

Furthermore, noise during fading is much less distracting (and less tiring as a result). This feature is the result of compandor characteristics, which reduce both noise and signal at poor signal-to-noise ratios rather than producing the noise bursts common to fm. Unlike normal sideband, ACSB provides a 5-dB capture effect that is several dB better than fm's normal 6-8 dB capability.

description of a typical system

The microphone audio is first passed through a preamplifier to bring it up to the proper level for the compressor circuitry. It is then passed through the first of two 2:1 compressors so that the normally desired 40-dB dynamic range of speech is compressed into 20 dB.

This compressed audio is then mixed with a 2850-Hz pilot tone set −7 dB below peak audio output. Both signals are then passed through a second 2:1 compressor, which compresses the 20-dB dynamic range of the first compressor into a 10-dB dynamic range. As one might expect, the pilot tone will be reduced during voice peaks by the amount of gain reduction produced by the audio peaks. This works out to about 10-dB reduction of the pilot on voice peaks, or 17 dB below peak reference level. Obviously, if we were to monitor this signal at this point, very compressed audio with a high pitched tone would be heard.

The final processing technique is to pre-emphasize the speech at a rate of 12 dB per octave. This is done to equalize the inherent differences in power levels in human speech, which tends to be concentrated in the low frequency areas.

The processed signal is transmitted on an otherwise standard single-sideband transmitter. It is also received on a standard single-sideband receiver using its normal AGC techniques (perhaps modified slightly to complement ASCB characteristics).

The received signal is passed through an AGC-controlled audio stage, which is controlled by a detector tuned to the pilot tone frequency. Its time constant is set very fast so that up to a 20 Hz-per-second change in input signal will be kept nearly constant in output level. Additionally, as a reduction in pilot level will cause an *increase* in output level, the suppression of the pilot in the second transmit amplitude compandor will be translated by the pilot AGC system into expansion at the same rate. Thus, a strong signal with no modulation will be quieted by the presence of the pilot signal. As the pilot is at its peak when there is no modulation, maximum quieting will occur.

After processing by the pilot-derived AGC, the leveled, expanded signal is again passed through a 2:1 expander. The pilot-derived expansion restores the 20-dB dynamic range from the transmitted 10-dB dynamic range signal. The second expander restores the original 40 dB dynamic range from the 20-dB pilot-derived expansion. The resulting audio is then processed through a 12-dB per octave de-emphasis filter to restore the original frequency response.

ACSB, then compresses 40 dB of speech information into a dynamic range of 10-dB, transmits it, then restores the 40-dB dynamic range at the receiving end of the system. This means that a signal-to-noise ratio of just over 10-dB is all that is required for an effective restored dynamic range (signal dynamics *and* signal-to-noise) of 40 dB. Additionally, a 2:1 quieting curve is established due to the constant presence of the pilot tone in the AGC and pilot expander receiver circuits.

A close look at the dynamics of ACSB will show that a carrier-to-noise ratio of only 5 dB will provide the equivalent of 20 dB quieting (to use fm terminology). Indeed, a 10-dB carrier-to-noise will give almost the full 40 dB dynamics and signal-to-noise we started with, except for the addition of a few dB of noise due to proximity to the noise floor. This certainly explains the weak signal superiority of this mode.

Dr. Lusignan estimates that ACSB has about a 15-dB advantage over normal SSB (he assumes 20-22 dB signal-to-noise is required for "high intelligibility" . . . all consonants audible). It also has a bandwidth advantage over fm, giving less interference from impulse noise (ignoring fm limiting) and higher signal-to-noise for a given power level at the receiver. This combination provides the measured 10-dB advantage of ACSB over 5-kHz deviation fm.

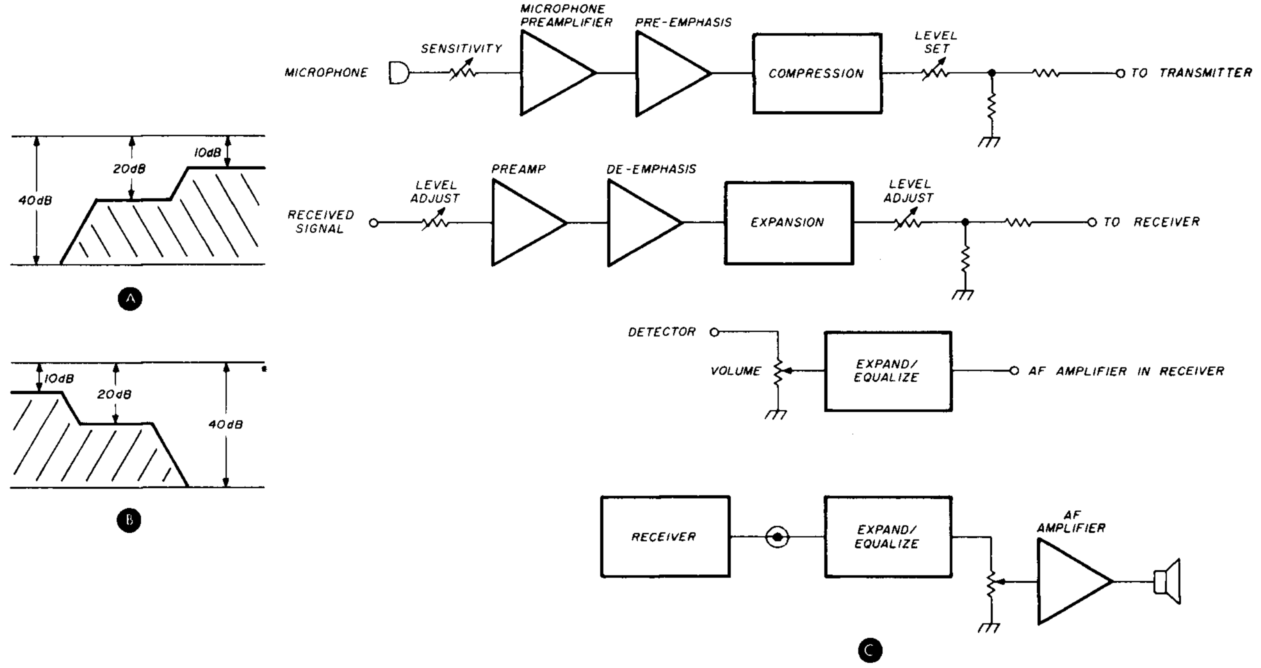
ACSB and NBFM comparison

Dr. Lusignan's report to the FCC[1] compares ACSB with NBFM as follows:

Signal to noise. ACSB shows a 10-dB advantage over fm at equal peak power levels.

Power required. ACSB requires 1/10th the power of fm for equal signal to noise. Additionally, ACSB requires 1/3 to 1/2 the average power of fm when the transmitters have equal peak output power, since the unmodulated output of ACSB is 7 dB less than its peak output.

Range. ACSB provides a reliable range equal to twice the fm range at distances up to 25 miles (40 km). Beyond 25 miles (40 km) this is reduced to 1.5 times due to the earth's curvature, which then becomes the limiting factor.



Characteristics of the Amplitude Compandored Sideband (ACSB) system proposed by Dr. Bruce Lusignan of Stanford University's Satellite Planning Center. *A* and *B* show respectively the 4:1 and 1:4 compression and expansion levels of the system. The block diagram, *C*, is a suggested starting point for experimenters for an ACSB Level 1 system.

During 8-watt PEP tests, simultaneously transmitting ACSB and fm combined on a common transmitting antenna and receiving on an ACSB and fm receiver fed from a common receiving antenna, the fm signal was lost in the south San Jose, California, area, while the ACSB signal was lost near Gilroy, California — some 16 miles (26 km) and 35 miles (56 km) from the Stanford transmitting site respectively.

Fading/multipath noise bursts (kerchunking). Field tests and bench tests show ACSB burst noise is 10 dB less than fm burst noise. Additionally, ACSB should be less prone to multipath distortions due to its narrower bandwidth and lack of sensitivity to phase relationships.

Message completion. ACSB is 3-5 times more reliable at a 9-mile (15-km) range than fm at equal power levels. ACSB at this range gives an 85 per cent completion rate compared with fm's 20 per cent rate.

Co-channel protection. On-channel rejection is 2-3 dB better with ACSB than with fm. Capture ratio for ACSB is about 5 dB compared to 7-8 dB for fm.

Adjacent-channel rejection. At 5-kHz spacings, ACSB provides 50-70 dB rejection of adjacent channel interference (depending on linearity and frequency stability). Fm at 25-kHz channel spacing yields 65-75 dB; at 20-kHz spacing it yields 55-65 dB.

According to the report,[1] the protection of 50 dB is sufficient, because other factors (intermodulation, co-channel interference) become equally problematical beyond this point.

"In typical applications the probability of loss from adjacent channel transmissions compared with 50 dB isolation is negligible compared with . . . shadowing or co-channel transmissions. Increasing . . . from 50-70 dB would not result in a noticeable change in the probability of successful transmissions."

Stability requirements. The ACSB system developed by VBC, Incorporated, for this study will automatically lock signals that are ±800 Hz from the center of the channel. At 160 MHz this is not outside normal stability for current fm equipment.

Digital transmissions. ACSB can handle up to 4 Kb/second in the main 2-kHz audio channel as well as about 20 b/second superimposed on the pilot carrier.*

Doppler shift in mobile service. The AFC circuit will control Doppler shifts normally encountered at all frequencies through 900 MHz (±800 Hz).

Fm/ACSB shared channels. It is possible to use ACSB and fm from a common repeater site providing the two channels are separated by 12.5 kHz. That is,

---

*Experiments with wider audio bandwidths (up to 3 kHz) are in progress. This should increase digital rates as well as improve audio fidelity.

an fm repeater could also provide two ACSB channels each 3 kHz wide centered 15 kHz away without interference from the ACSB channels to the main channel. (This might be a solution to the 15-kHz split situation on 2 meters between 146-148 MHz, for example.)

hardware

Commercially available LSI chips that perform all ACSB functions should be available in one to two years, depending on FCC action, market acceptance, and other normal factors relating to volume and production. In the meantime, experiments with ACSB Level 1 is within easy reach of the experimentally inclined ham. The Signetics NE 570/571 Compandor IC is available from Jameco Electronics, 1021 Howard Ave., San Carlos, California 94070. Their price is $4.95 (1980 catalog), but they also have a $10.00 minimum.

The NE570, an LM324 op amp, and an rf-tight box will allow everything necessary for 2:1 compandoring with pre-emphasis/de-emphasis. My own experimentation shows a marked improvement on all but the weakest signals (signals under 4-5 dB signal-to-noise ratio show no apparent improvement, even though background noise with no signal *will* be improved). The block diagram on the preceding page is recommended as a starting point.

conclusion

Out here in the west we like to talk about the wide open spaces. Well, you can still drive to those wide open spaces without too much effort. In the crowded city, however (and we *do* have some crowded cities), one soon learns that it is best to give one's neighbor plenty of elbow room whenever possible. On vhf, ACSB promises a good way to do just that.

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Simple and stable 1152-MHz multiplier chain for Amateur microwave bands

There is an apparent abundance of commercially built high-frequency and vhf equipment available, little of which is adaptable for use above 1 GHz. Purchased equipment may be used to provide a 1296-MHz station (generally a varactor tripler driven by a 432-MHz transmitter, and a relatively high-noise-figure receiving converter with no rf preamplification). It's virtually impossible to purchase any station equipment specifically designed for weak-signal communications above 1296 MHz. Thus far only the most intrepid experimenters have ventured above 1296 MHz, generally hand-in-hand with a master machinist and expensive power tools (lathes and the like).

All is not lost, however. Because of two interesting factors, building a microwave station is now possible for most experimenters willing to spend a few evenings etching PC boards and soldering components. That's right — no more machinists, at least not for 1296-MHz and 2304-MHz equipment.

frequency relationships

The first factor to help resolve the microwave dilemma lies in the arithmetic of our microwave bands. Within all our bands above 1300 MHz is at least one frequency that is a multiple of that "magic number" — 1152 MHz. Even 1296 MHz is related to 1152 MHz. The former frequency was originally selected for weak-signal work because it is the third harmonic of 432 MHz and therefore can be obtained by tripling. A difference frequency of 144 MHz exists between 1296 and 1152 MHz, which becomes the receiving i-f. Note also that 1152 MHz is the eighth harmonic of 144 MHz. The relationships between the 1152-MHz magic number and weak-signal frequencies in our uhf and microwave bands are listed in **table 1** and graphically illustrated in **fig. 1**.

low-order frequency multiplication

Another interesting mathematical feature is that the frequency of 1152 MHz can itself be generated by a chain of low-order (and therefore relatively good efficiency) multipliers. This chain, made up only of frequency doublers and/or frequency triplers, allows filtering to reduce undesired (spurious) signals at the multiplier-chain output. Many writers have insisted that starting frequencies be in the range of about 50-100 MHz to avoid producing undesired harmonics in the 144-MHz and/or 432-MHz bands. This requires an overtone crystal. As the frequency of such crystals is notoriously difficult to pull, a variable-crystal-frequency source was developed that allows use of

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crystals operating in the fundamental mode, below about 20 MHz.

One optimum chain (shown by the heavy-bordered boxes in **fig. 2**) thus starts at 16 MHz, triples to 48 MHz, doubles to 96 MHz, doubles a second time to 192 MHz, doubles a third time to 384 MHz, then triples to 1152 MHz. The use of this chain requires that the unwanted third harmonic of 48 MHz be very greatly attenuated. If present, the third harmonic will fall into the low end of the 2-meter band (at the weak-signal EME portion around 144.000 MHz). Radiation of any significant amount of energy at that frequency will tend to irritate neighboring 2-meter CW operators. In a vhf-contest environment, the third or ninth harmonics may very well QRM your own 2-meter or 70-cm station. These two undesired harmonics, however, appear to be the only problem harmonics. The ability to suppress undesired harmonics is enhanced by proper partitioning of the multiplier chain. The basic-frequency (for example 16-MHz) oscillator and only a few of the total number of multipli-

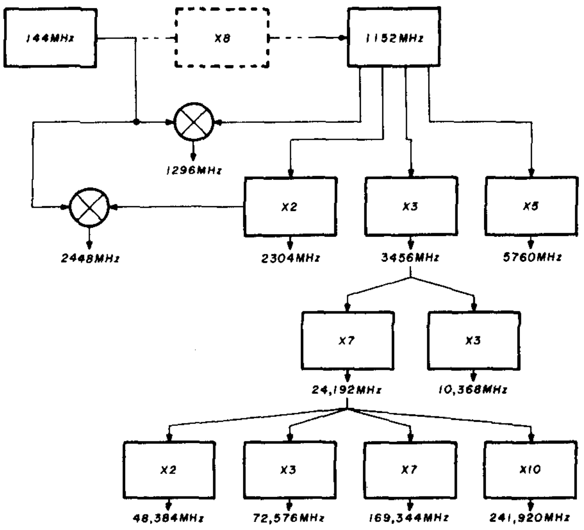


fig. 1. Relationship between 1152 MHz and desirable frequencies for highly stable, weak signals in the Amateur bands above 1 GHz.

table 1. Relationship between "magic number" 1152 MHz and weak-signal frequencies in the Amateur uhf and microwave bands.

band (MHz)	desirable frequency (MHz)	*by mixer*	*by multiplier*
1240-1300	1296	1152 + 144	432 × 3 or 108 × 2 × 2 × 3
2300-2450	2304 2448	(1152 × 2) + 144	1152 × 2 or 102 × 2 × 2 × 3 × 2
3300-3500	3456		1152 × 3
5650-5925	5760		1152 × 5
10,000-10,500	10,368		1152 × 9 = 1152 × 3 × 3 = 3456 × 3
24,000-24,250	24,192		1152 × 21 = 1152 × 3 × 7 = 3456 × 7
48,000-50,000	48,384		1152 × 42 = 1152 × 3 × 7 × 2 = 3456 × 7 × 2 = 24,192 × 2
71,000-76,000	72,576		1152 × 63 = 1152 × 3 × 7 × 3 = 3456 × 21 = 3456 × 7 × 3 = 10,368 × 7 = 24,192 × 3
165,000-170,000	169,344		1152 × 147 = 1152 × 3 × 7 × 7 = 3456 × 49 = 24,192 × 7
240,000-250,000	241,920		1152 × 210 = 3456 × 70 = 48,384 × 5

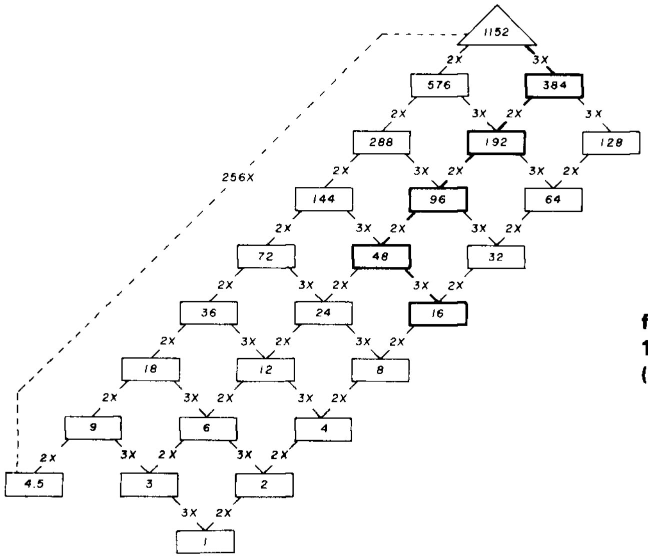


fig. 2. Frequency-multiplication tree for generating a 1152-MHz signal using only low-order multiplication (doublers and triplers) from a low-frequency source.

ers are packaged in a low-frequency building block. The remainder of the multipliers are packaged in a separate, second building block. The low-frequency block output may then be made to have very low levels of signals at undesired frequencies.

The second important factor is the present-day ability to generate the desired 1152 MHz signal in a practical manner from a lower frequency driving signal. In this regard, great thanks should be given to Paul Shuch, N6TX, for his design of a PC board 96-1152 MHz multiplier unit.[1] This microstrip unit, for which a printed circuit board and set of tuning capacitors are available from N6TX, was apparently designed to replace a multiplier chain[2] using a packaged oscillator, at 96 MHz, driving a pair of 2N5179 transistor frequency doublers to 384 MHz; a pair of 2N3866 power amplifiers, providing several hundred milliwatts at 384 MHz;[3] and a step-recovery-diode tripler to provide about 5 milliwatts at 1152 MHz.[4] Having built three such frequency-multiplier chains, I must concur with the general undesirability of vhf multipliers using step-recovery diodes.

The replacement of the entire 96-1152 MHz chain with three stages of transistor multipliers (using the Motorola MRF 901) results in a great saving of time, labor, and parts cost. I've built several of the 1152-MHz sources (described later in this article) as well as a 1296-MHz solid-state transmitter, based on the microstrip multiplier of reference 1, and have selected that basic design for the 96-1152 MHz portion of this common microwave system. While some may desire to be purists and design *all* their equipment themselves, I believe that judicious use of the contributions of others often makes for the best (and the most rapid) attainment of the end goal: to get as many stations on the microwave bands as quickly and inexpensively as possible.

96-MHz VXS

As mentioned, the N6TX unit was designed for use with a fifth-overtone oscillator, which is replaced with the variable-crystal-frequency source (VXS) shown in the block diagram of **fig. 3**. I've arbitrarily chosen a tuning range, at 2304 MHz, of 2303.928-2304.086

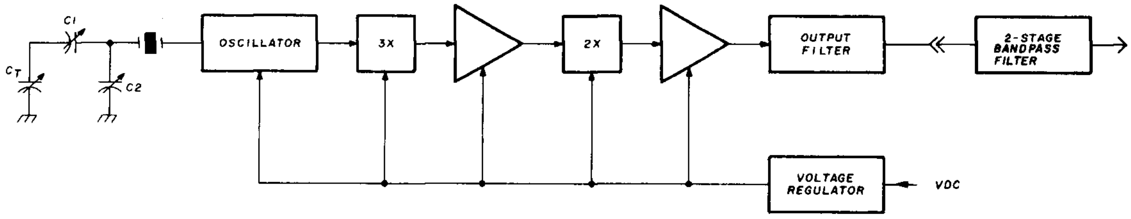
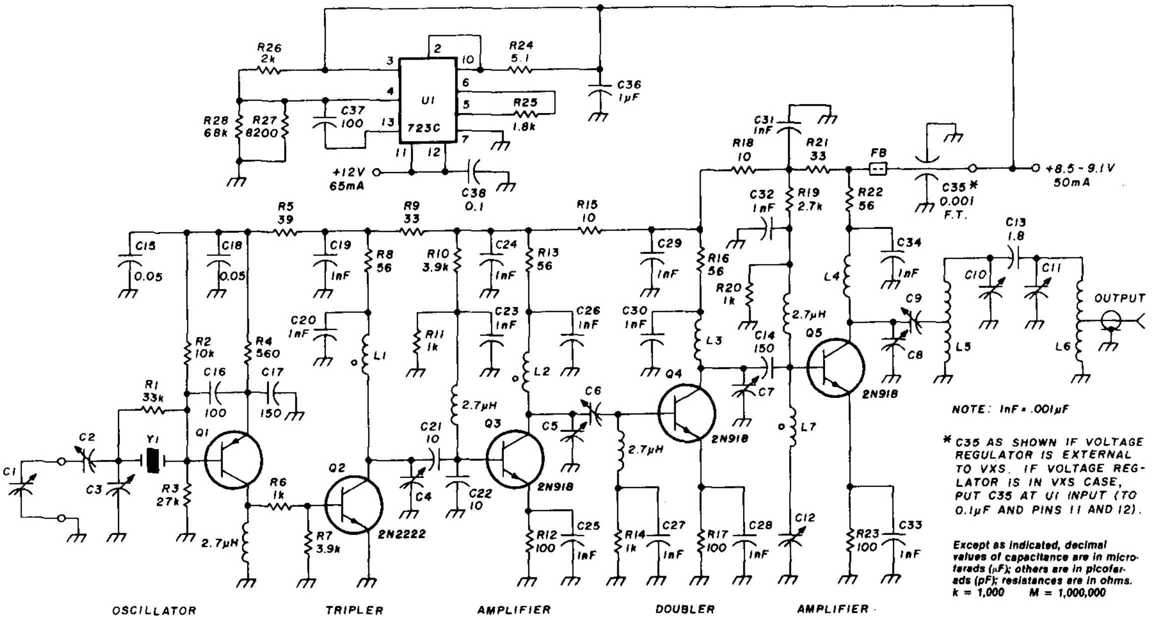


fig. 3. Block diagram of the VXS/BPF system. Output is at 96 MHz using a 16-MHz overtone crystal oscillator.



C1	1-10 pF tuning
C2	10-60 pF trimmer
C3	5-50 pF trimmer
C4,5,6,7, 8,9,10, 11,12	5-25 pF trimmer
L1,L2	6 turns No. 28 (0.3 mm) on T37-10 core
L3,L4	4 turns No. 22 (0.6 mm) on T37-12 core
L5,L6	8 turns No. 20 (0.8 mm) 1/4" (6.4 mm) 5/8" (16 mm) long
L7	8 turns No. 28 (0.3 mm) on T37-10 core
Q1	2N2907, 2N3906 PNP
Q2	2N222, 2N918, 2N2857 or 2N5179 NPN
Q3,Q4,Q5	2N918, 2N2857 or 2N5179
U1	LM723 (DIP)
Y	16,001 kHz Crystal (HC-6U holder)

fig. 4. Schematic of the 96-MHz variable-crystal-frequency (VXS) source. Circuit is built on a single-sided PC board.

MHz, corresponding to an oscillator frequency range of 15.9995-16.0006 MHz (therefore, an 1100-Hz range at 16 MHz gives a 158.4-kHz range, when multiplied 144 times, to frequencies around 2304 MHz). This requires that the crystal frequency be pulled about 0.0066 per cent, which certainly can be achieved with almost any fundamental crystal.

The crystal frequency was chosen as 16,001 kHz with 20 pF parallel capacitance, and thus is slightly higher than the nominal 16,000-kHz frequency. By paralleling the crystal with a bit more capacitance, provided by the main tuning capacitor C1 and its series and shunt band-setting capacitors C2 and C3, the desired frequency range can be realized. The schematic of the 96-MHz variable-crystal-frequency source is shown in **fig. 4**, the PC-board layout is shown in **fig. 6**, and the parts placement in **fig. 7**.

PNP transistor Q1 is the crystal-controlled oscillator, driving a frequency tripler, Q2. Transistor Q3 is a 48-MHz buffer, Doubler Q4 and a tuned buffer, Q5, at 96 MHz, follow. The 96-MHz output filter is a double-tuned bandpass configuration. An additional double-tuned bandpass filter (**fig. 5**) may be used at the high-frequency multiplier block or placed in a separate shielded box outboard of the source and multiplier blocks to provide an additional 20 dB suppression of the undesired signals provided by the VXS. The tuning range, with the components listed, is sufficient to allow the VXS to be used with crystals between 15-18 MHz. In the first case (15-MHz crystal) the final multiplier output is 1080 MHz, which is used for doubling to 2160 MHz. This frequency is used for local-oscillator output in 2304-MHz receiver converters with a 144-MHz i-f. The 18-MHz crystal produces a final multiplier output of 1296 MHz for use in exciters in the 23-cm band.

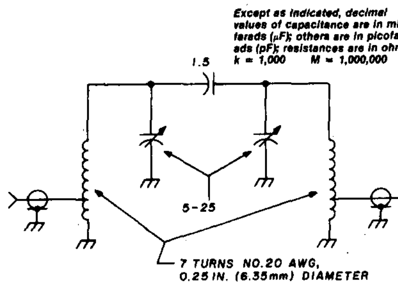


fig. 5. An auxiliary outboard bandpass filter that may be used at the high-frequency multiplier block to provide an additional 20 dB suppression of unwanted signals from the VXS.

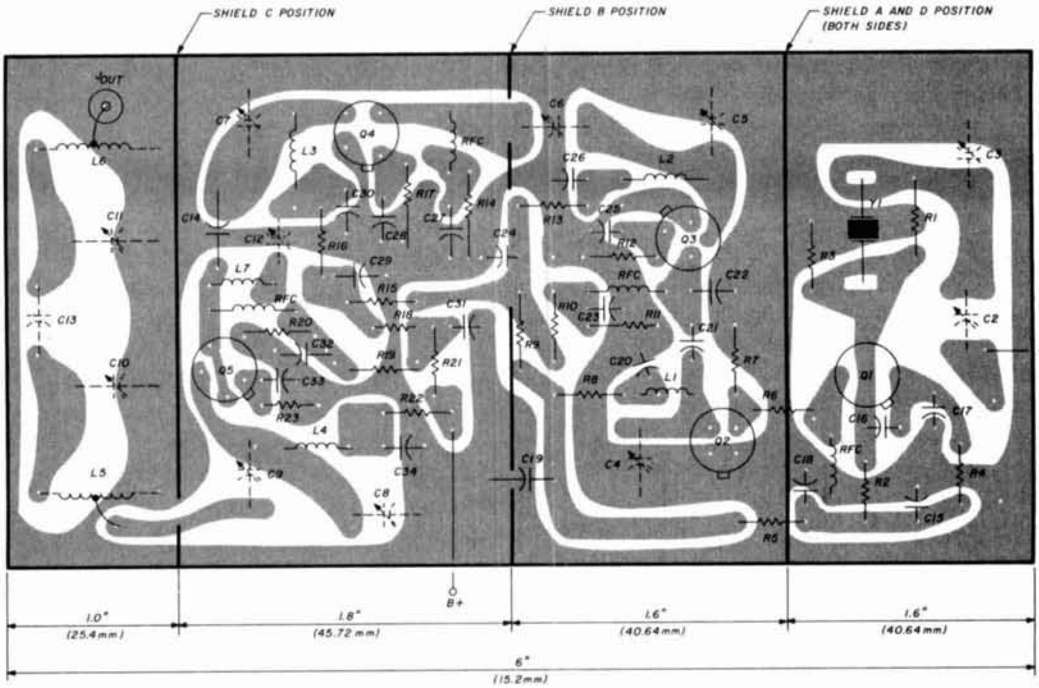


fig. 6. PC-board layout for the VXS-96 microwave signal source.

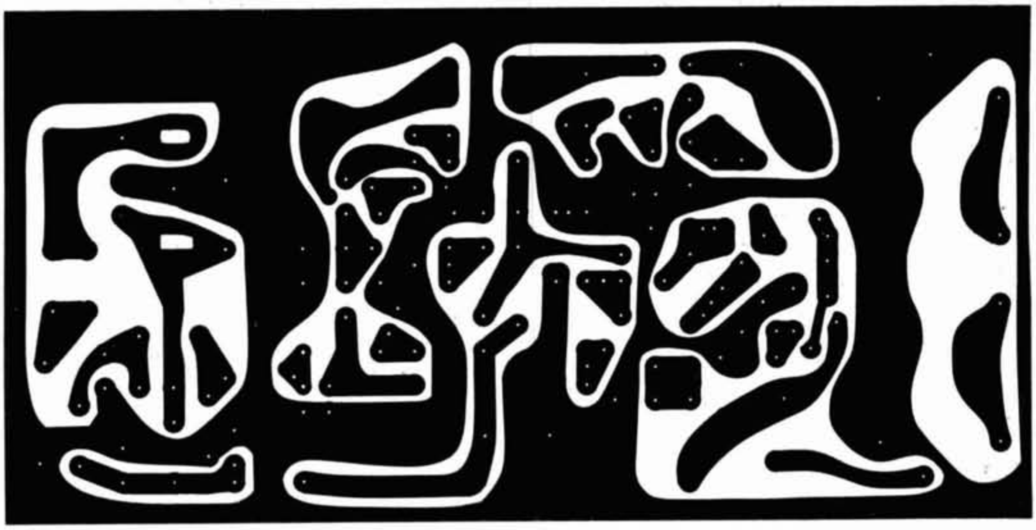


fig. 7. Component side of the VXS-96 board (copper side). Mount all variable caps on this side; all other components are mounted on the reverse side.

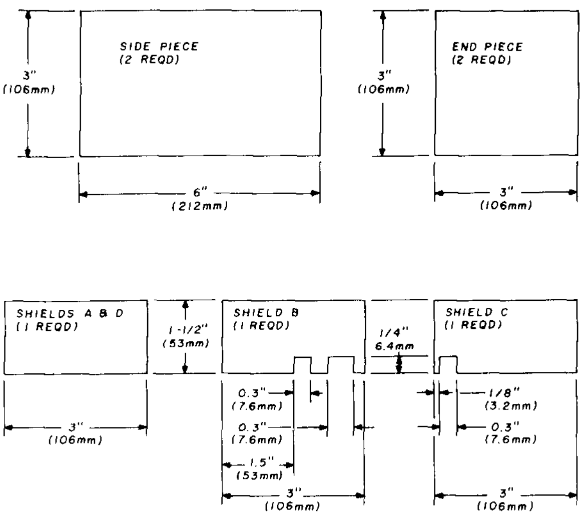


fig. 8. Shield, side, and end pieces for the VXS constructed from double-clad PC-board stock. Covers (top and bottom) are 3¼ × 6¼ inch (8.25 × 15.9 cm) PC-board pieces.

some other uses

The output of the VXS can be:

1. Set to 116 MHz (by using a 19.334-MHz crystal) for use as a 2-meter local oscillator.
2. Used with a frequency doubler to generate a 192-MHz signal for use as a 220-MHz local oscillator.
3. Set by a 16.834-MHz crystal to provide a 101-MHz signal for input to a cascaded pair of frequency doublers to generate a 404-MHz local-oscillator signal for use in 70-cm equipment. (See **fig. 9**.)

In the VXS schematic of **fig. 4**, both crystal leads are above ground in the circuit. This might be a problem if crystal switching is desired. For higher stability the crystal will be placed in a thermally isolated environment (such as a crystal oven positioned above the PC board or in a block of styrofoam).

shielding considerations

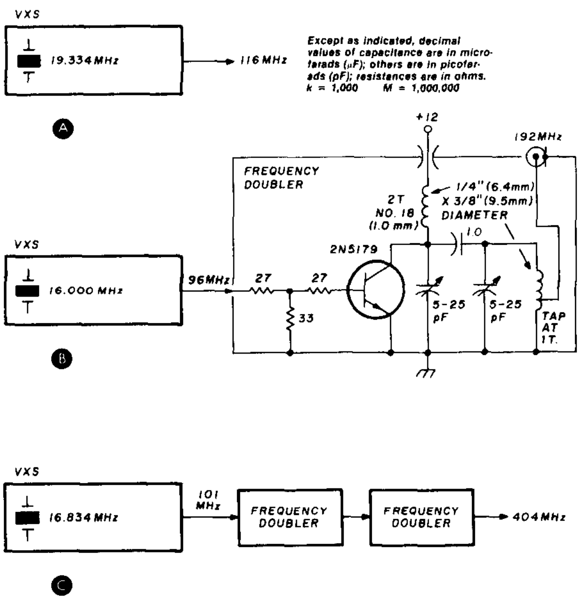
Note, in **fig. 8**, that pieces of double-clad PC board form three shield partitions, A, B, and C, directly soldered to the copper-clad side of the PC board. A similar partition, D, is soldered to a PC-board case built around the entire board above shield A (between the oscillator and the multiplier stages) for added attenuation of oscillator harmonics. The oscillator is enclosed in a shielded compartment separated from the tripler-buffer area, which is separated from the doubler-buffer area. The output filter is in its own compartment, shielded from all oscillator, frequency multiplier, and buffer stages.

The VXS circuit also includes a high degree of power-supply decoupling. An IC voltage regulator, U1, provides a constant voltage to the circuit; this is necessary not only to prevent oscillator frequency changes with varied input voltage (in my case, from the battery in my automobile during mobile operation from any convenient mountaintop), but also to keep all transistors operating at fixed biased points, which causes the transistor input and output impedances to be stabilized. This stabilization of device impedances prevents changes in tuning with changing input voltage and contributes to the overall spectral purity of the VXS output signal. Note the use of a BNC connector for the rf output of the source, and the use of a feedthrough capacitor to bring the voltage into the VXS enclosure. Both components are used to maintain the shielding integrity and provide minimum amplitude of undesired signals.

Also note that the voltage regulator IC, U1, and the associated resistors, R24-R28, and capacitors C36-C38 are mounted on a wire-wrap 18-pin IC socket, with the end pins on either side extending full length and soldered to the inside of the case. The remaining 14 pins are bent at right angles, close to the bottom of the socket; the regulator-circuit resistors and capacitors are soldered between the bent pins. See **fig. 10**.

spectrum analysis

The VXS is aligned by using any of the well-known



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

fig. 9. Other uses for the VXS. A shows a 116-MHz local oscillator for a 2-meter converter with a 28-MHz i-f. A frequency doubler for generating a 192-MHz local-oscillator signal for a 220-MHz converter, using a 28-MHz i-f, is shown in B. A pair of cascaded frequency doublers, C, generate a 404-MHz local-oscillator signal for a 432-MHz converter with a 28-MHz i-f.

tuning procedures including: a) monitoring the emitter or collector current of the stage following the stage you're tuning for an increase in current, and b) using a test receiver, grid-dip meter and so forth. If a spectrum analyzer is available (and its use is highly desirable although not mandatory) an output signal spectrum similar to that shown in **fig. 11** may be obtained. In **fig. 11**, spectrum (A) is for the basic circuit, built on the circuit board, but without the output filter (L5, L6, C10, C11, and C13). Note that the second harmonic is at a level of only −14 dBc (dB below the desired carrier, at 96 MHz). Adding the output filter, but without shielding, typically provides the (B) spectrum, wherein the greatest-amplitude undesired signal is still the second harmonic, now suppressed to a level of −40 dBc. Adding the shields and a shielded box (**fig. 8**) results in the (C) spectrum (shown in solid lines in **fig. 11**). With the shields and shield box, the greatest-amplitude undesired signals are those spaced above and below the desired signal by the fundamental frequency; for example, at 80 and 112 MHz.

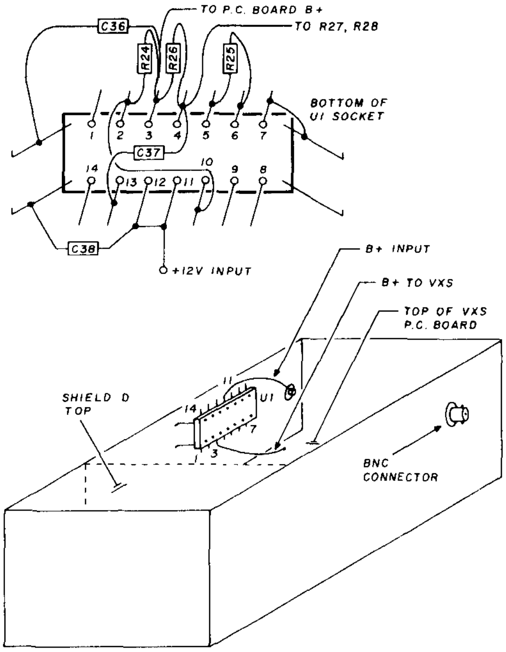


fig. 10. Method of mounting the U1 (18-pin) socket on the VXS case wall with four end pins used for support. The remaining 14 pins are bent outward and hold R24-R26; C36-C38.

With the use of the outboard additional filter (labeled BPF-96) the only signals found, up to 1500 MHz, are as shown in spectrum (D):

frequency (MHz)	16-MHz oscillator harmonic	96-MHz output harmonic	attenuation (dBc)
80	5		−77
112	7		−79
192	12	2	−70
288	18	3	−70
480	30	5	−74

Minor signals occur at the 65th, 66th, and 67th harmonics of the crystal frequency (16.001 MHz), with respective amplitudes of −73, −75, and −76 dBc.

Even with the additional two-section BPF-96 filter, the desired 96-MHz output has a level of 16 dBm (40 milliwatts). Because a significantly lower level, on the order of 0 dBm (1 milliwatt), is required for driving the first doubler in the high-frequency multiplier circuit, additional bandpass filters, or a lowpass filter having a cutoff frequency on the order of 150 MHz, could be easily used. Note that the presence of the second and third harmonic of the desired output signal is not particularly troublesome, since these frequencies will be generated in subsequent multiplier circuitry anyway.

To achieve the required *Q*, the on-board double-tuned bandpass filters use air-wound rather than

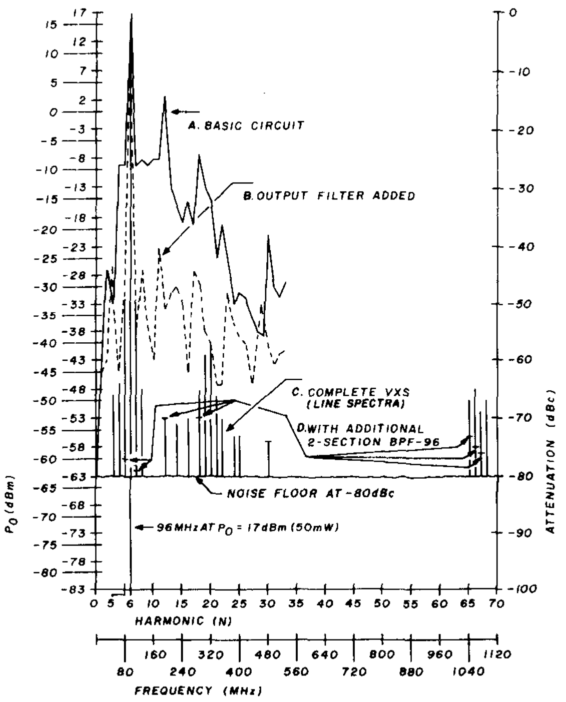


fig. 11. Output of the VXS as seen on a spectrum analyzer. Attenuation of undesirable signals is shown as a function of frequency for three different configurations. Note the effect provided by adding the bandpass filter.

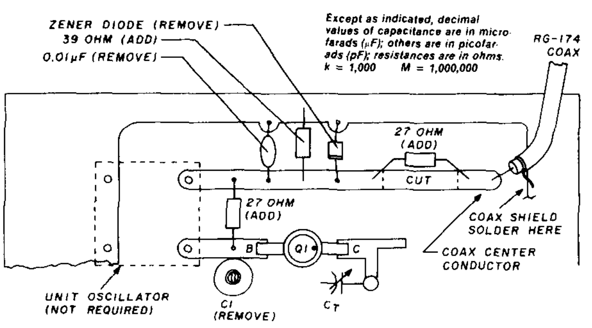


fig. 12. Modification of the N6TX multiplier board for use with an external signal source.

lower-*Q* toroidal inductors. It is probable, because of the relatively high insertion loss of the bandpass filter sections, that the filters are not completely optimized. However, the ability to provide easily tuned filters using low-cost components was deemed more important than squeezing out an additional few dB of harmonic rejection. Whether or not additional filtering is used, at least 10 dB of attenuation (a T-pad with 22-ohm series arms and a 33-ohm shunt arm) is used at the N6TX high-frequency multiplier board input to ensure that a relatively constant output terminating impedance appears, as well as to reduce the drive level. (I've burned out several MRF 901s but haven't harmed any 2N5179s, in the first doubler stage, with only 6 dB attenuation.)

I prefer the 2N5179 in this stage with an increase in the tuning capacitance of the 192-MHz circuit; this is especially advantageous because the 2N5179 is not only less expensive but is also more readily available than the MRF 901. Of course, any change in terminating impedance can detune the filter or filters and reduce the ultimate suppression of undesired harmonics. Similarly, the ultimate suppression of harmonics of the 1152-MHz signal is a function of the suppression provided by the N6TX circuit and any additional filtering applied thereafter. See **fig. 13**.

construction

After building the basic PC board of **fig. 6** and drilling all component mounting holes, mount the crystal socket on the non-copper side of the board with 4-40 (M3) by 0.37 (9.5 mm) screw, lockwasher and nut. If a crystal oven is to be used, don't mount the crystal socket but wire the oven crystal leads to the appropriate PC-board pads after assembling the board and mounting the crystal oven on it. Before mounting the components, solder the two box sides, cut as shown in **fig. 8**, to the long edges of the PC board. About 1-1/2 inches (38 mm) of the sides extend above and below the plane of the circuit board. A hole for the feedthrough capacitor and for the BNC connector can be drilled in the appropriate side, either before or after soldering.

Solder shield A between the two sides and also to the copper-clad PC-board side. The counterpart of shield A (shield D in **fig. 8**) is positioned against the non-copper side of the PC board and soldered to the pair of opposed box sides. At this side, all components should be mounted to the PC board. Variable capacitors C2-C12 are soldered to the copper pattern on the bottom of the board, while all remaining components are mounted from the top (non-copper bearing) surface of the board.

After installing all components, carefully mount shield B then shield C before soldering the end pieces between the two sides and to the ends of the PC board. A hole may be drilled in the end piece, at the oscillator end of the board, for tuning capacitor C1. However, if the VXS is to be used as a fixed-frequency source, in which capacitor C1 is merely adjusted to set the output to a particular frequency and not to be continuously tuned (as in setting a local-oscillator frequency in a receiver), then capacitors C1 and C2 are dispensed with; frequency is adjusted with C3. Note that output filter inductors L5 and L6 and the 48-MHz trap inductance L7 are also mounted beneath the PC board. The voltage regulator IC socket, with its components, can now be mounted by soldering to one copper side piece, as shown in **fig. 10**.

tune up

Tack solder the top cover to all four sides, but don't completely solder. Install the crystal in its socket and apply at least +12 but less than +20 volts to the B+ feedthrough. Note the voltage at regulator pin 3 (which will be pin 4 of the socket, since pin 1 is attached to ground). The regulator output voltage should be between +8.5 and +9.1 volts dc. Total

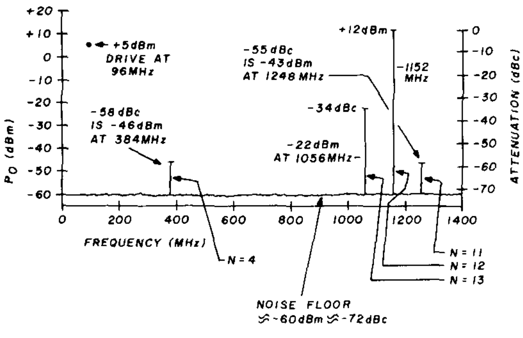


fig. 13. Output spectra of the N6TX multiplier block using the VXS/BPF circuits.

current into the box will be no more than about 75 milliamperes and will probably be considerably less at this time. The base lead of Q2 can be monitored for a 16-MHz signal, indicating that the oscillator is working. Monitor the base lead of Q3 with a 48-MHz rf indicator and tune C4 for maximum rf voltage. Shift the rf indicator to the base lead of Q4 and tune C5 and C6 for maximum voltage at 48 MHz. Retune the indicator to 96 MHz and monitor the base of Q5; tune C7, then C6 and C5, for maximum voltage.

Move the monitor to the tap of filter coil L5 and tune C8 and C9 for maximum voltage. Now connect the monitor to the output connector and tune C10, C11 for maximum output. Then retune C9, C8 for maximum 96 MHz signal. Note that a commercial fm receiver, with carrier-strength meter, may be used for the 96 MHz monitor indicator.

After tuning the bandpass filter for maximum 96-MHz signal, reset the tuning monitor to 48 MHz and adjust C12 for minimum 48-MHz signal. The outboard filter can now be tuned, if used, for maximum 96-MHz signal. As indicated previously, if you can beg or borrow a spectrum analyzer, set the analyzer to display the spectrum from at least 15 MHz to at least 150 MHz (and preferably to at least 500 MHz). Finely adjust C4-C11 several times in sequence for best suppression of undesired harmonics while maintaining the desired 96-MHz signal at a reasonable maximum.

Capacitors C6 and C9, especially, are used to adjust the symmetry of the amplitudes of the undesired fifth and seventh harmonics of the crystal oscillator next to the desired sixth-harmonic signal at 96 MHz. Capacitor C12 has some effect on the tuning of C7. Furthermore, if you use a spectrum analyzer, the 68-k resistor in the voltage regulator circuit may be replaced with a 25-k pot in series with a 56-k fixed resistor, and the pot will vary the circuit voltage. Varying the regulated voltage will often allow you to find a specific voltage at which maximum harmonic suppression is achieved, although power output will change [but, as previously mentioned, it isn't particularly important so long as at least 20 milliwatts (+ 13 dBm) are available at the attenuator input to be added to the N6TX multiplier].

multiplier modifications

The N6TX multiplier board (**fig. 12**) is modified by removing the 9.1-volt zener, the 0.01-μF capacitor in parallel with the zener, and the 180-ohm resistor to the zener (not shown). A 27-ohm, 1/8-watt resistor is soldered from the base lead of the first multiplier transistor to the circuit trace that was the unit oscillator B+ line. A 39-ohm resistor is soldered from the B+ trace to ground, and one end of another 27-ohm resistor is also soldered to the B+ trace. The other end of the second 27-ohm resistor is soldered to the outer conductor of a piece of RG-174 coaxial cable, whose shield is soldered to multiplier ground.

A coaxial cable is connected from the input of the outboard bandpass filter, if used, to the BNC connector on the VXS. If transistor Q1 of the multiplier is a 2N5179 transistor, tuning capacitor CT, on the collector side, should be increased from 1 to 5 pF. The original C1 capacitor (at the first doubler input and unit oscillator output) is no longer needed.

The multiplier should be tuned in the same manner as specified by N6TX in his article. I've found that the tripler input and three output filter capacitors should be the suggested Triko 202-08M, although the pair of 384 MHz tuning capacitors may have to be increased to 2-10 pF, to adequately tune the modified multiplier board. **Fig. 13** illustrates the output spectra of the modified multiplier block when driven with the VXS and 96-MHz outboard bandpass filter.

Some uses of the VXS and multiplier blocks are shown in **figs. 14** through **17**. In **fig. 14**, one possible way that high transmitting power may be eventually economically realized within the next several years in the 2300-2450 MHz band will probably be by use of microwave oven magnetrons (a magnetron being es-

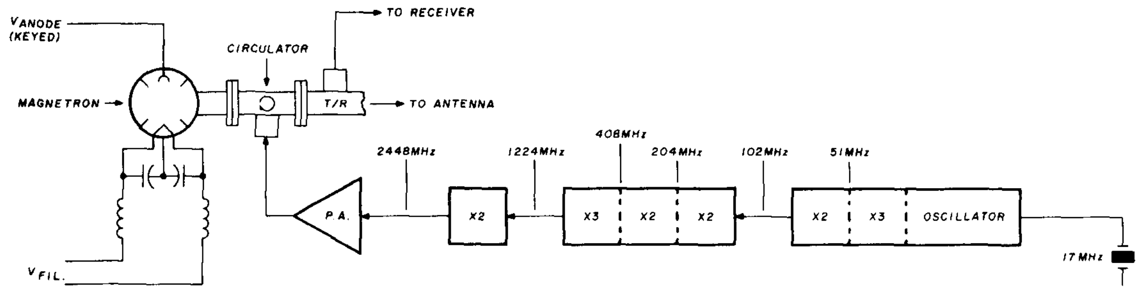
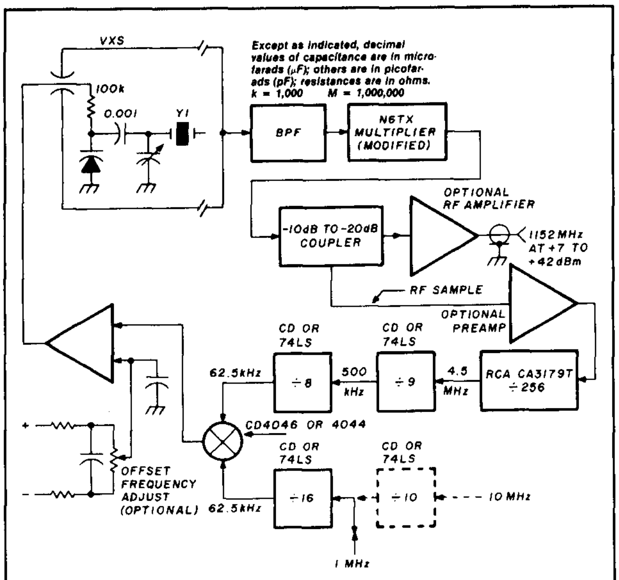


fig. 14. Possible transmitter for crystal-controlled injection locking of a microwave magnetron producing 600 watts output at 2448 MHz.



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

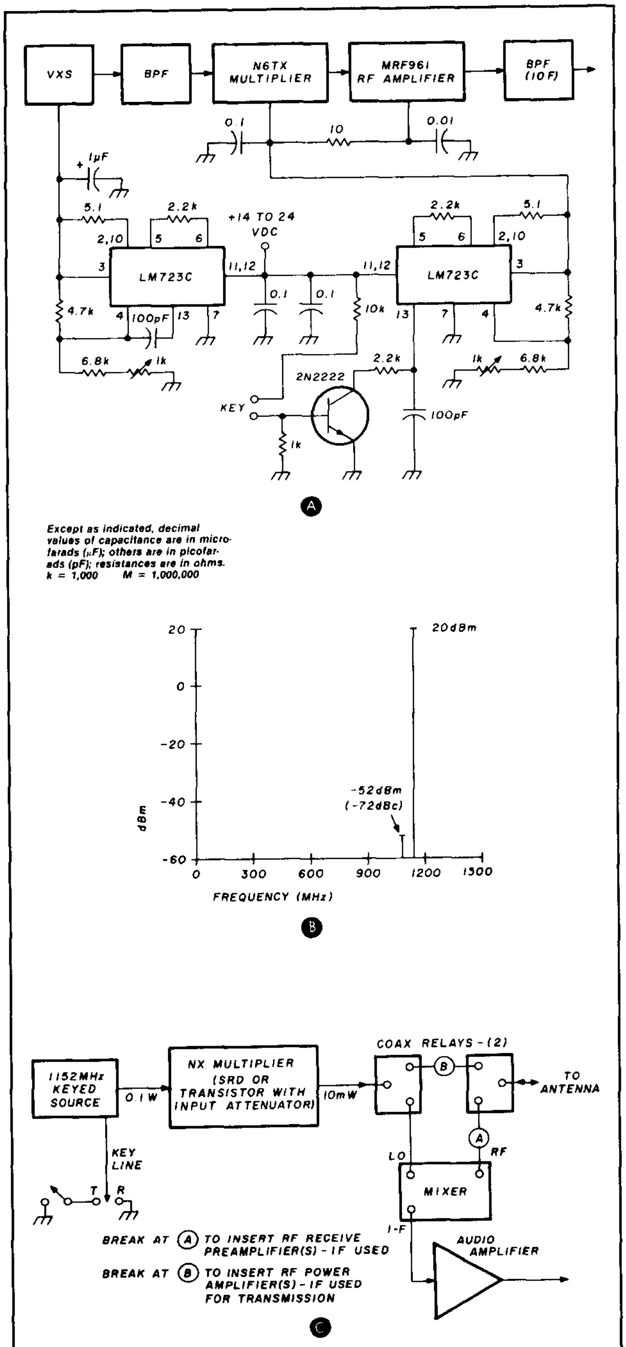
fig. 15. Block diagram of a phase-locked 1152-MHz subsystem that will provide a highly stable signal for multiplication into the 13-cm, 9-cm, or 5-cm bands.

sentially a diode tube in which oscillations occur at microwave frequencies because of the finite time required for electrons to travel or drift between the tube elements). Available magnetrons, which cost about as much as a vhf power tube of the 4CX250 type, provide up to 600 watts of output power but are normally pretuned at the factory for oscillation at about 2450 MHz.

The tuning adjustment is not normally accessible (apparently being inside the vacuum envelope of the tube), but some tuning can apparently be accomplished by varying the tube anode current. Many operators interested in magnetron use have concluded, although none (to my knowledge) have yet proved, that it should be possible to reduce the magnetron frequency to be just within the upper edge of the 2300-2450 MHz band. Advantageously, another multiple of 144 MHz is present at 2448 MHz, which is also a 144-MHz i-f above 2304 MHz, itself a second harmonic of 1152 MHz. It may well be possible, using equipment as shown in **fig. 14**, to injection-lock a 600-watt output magnetron with less that 10 watts of power from a very-high-frequency-stability source, whereby the magnetron assumes the same stability as its locking source. The 10-watt power level is obtainable, now, with fully-transistorized amplifiers.

Probably the greatest obstacle in achieving high power in the 13-cm band is the requirement for a 600-watt circulator. I know of no such unit commercially available, although the technology appears to exist. I am confident, however, that some experimenter will eventually design, or design around, a circulator for this frequency and power level, allowing an injection-locked, high-power source to be realized. **Fig. 15** is a phase-locked 1152-MHz subsystem that will provide a highly stable signal for multiplication into any of the 13-cm, 9-cm, or 5-cm bands.

Fig. 16A is a keyed 1152-MHz source having about



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

BREAK AT Ⓐ TO INSERT RF RECEIVE PREAMPLIFIER(S) - IF USED
BREAK AT Ⓑ TO INSERT RF POWER AMPLIFIER(S) - IF USED FOR TRANSMISSION

fig. 16. Keyed 1152-MHz source providing about 0.1 watt output, A. Output spectrum, B, shows only one spur at slightly more than −70 dBc. A direct-conversion transceiver using the 1152-MHz source is shown in C.

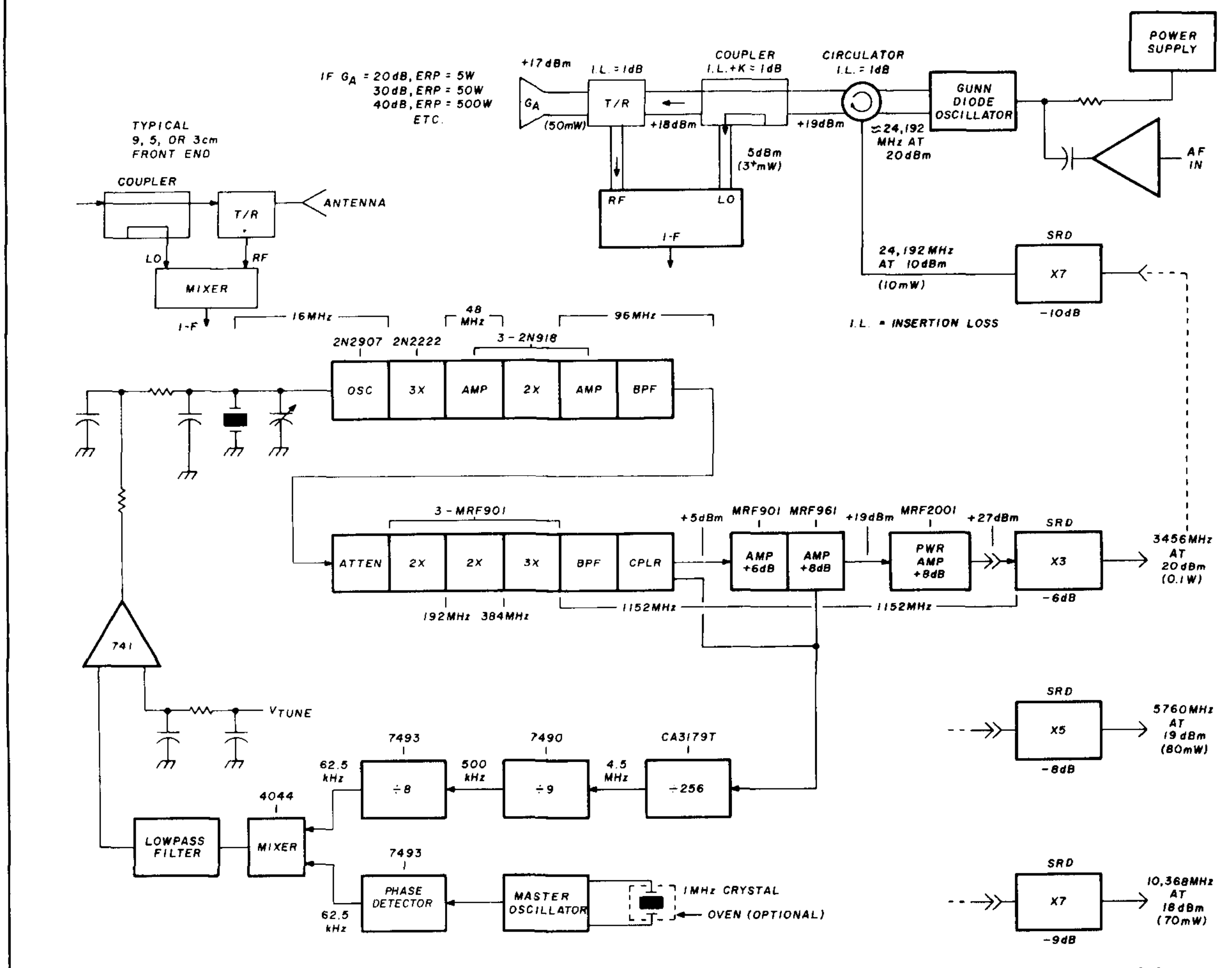


fig. 17. Some other microwave configurations based on the multiplication scheme described in this article.

1/10th watt output, while **fig. 16B** shows its output spectrum (only a single spurious output at slightly more than 70 dB below the carrier). **Fig. 16C** shows a direct-conversion transceiver using the source of **fig. 16A**. **Fig. 17** shows other microwave source configurations, all based upon multiplication of the 1152-MHz signal.

summary

All of our microwave bands have one frequency that's related to 1152 MHz. By building a power source at 1152 MHz, multiplication to the microwave bands becomes possible. A relatively simple, yet stable, 1152-MHz chain is necessary; one such chain is described. The power amplifier, producing 100 milliwatts at 1152 MHz, is an adaptation of a circuit designed by Dick Frey, WA2AAU. Simple frequency doublers and receiving mixers for 2304 MHz have been described in many articles (check your *ham radio* and *QST* indexes). Thus it's possible to find easily built components for 2304 MHz right now. Higher-frequency blocks and subsystems are being worked on, and further results, from this writer or others, should be forthcoming.

acknowledgments

I would like to thank Dick Frey, the other half of the present Mt. Greylock microwave gang, for his help and encouragement; all the local microwave people for their interest; and my four-year old son, Jeremy, and nine-year-old daughter, Alyssa, for helping to mount parts onto PC boards and for tuning and measuring.

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2. H. Paul Shuch, N6TX, "UHF Local Oscillator Chain for the Purist," *ham radio*, July, 1979, page 27.
3. H. Paul Shuch, WA6UAM, "Easy to Build SSB Transceiver for 1296 MHz," *ham radio*, September, 1974, page 8.
4. H. Paul Shuch, N6TX, "Improved Grounding for the 1296-MHz Microstrip Filter," *ham radio*, August, 1978, page 60.

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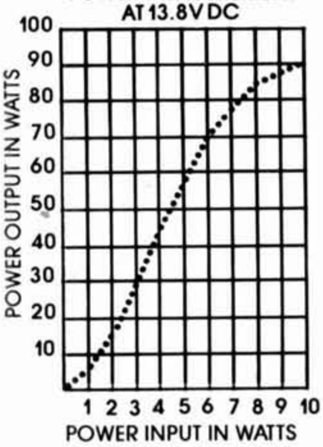

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- Two VFOs.
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- Automatic scan of memories and of 440-450 MHz band (in 25-kHz steps). Locks on busy channel and resumes when signal disappears. HOLD or mic PTT button cancels scan.
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the weekender

Inrush current protection for the SB-220 linear

Do you have adequate surge protection for your SB-220? If you own this fine piece of gear or similar equipment without the benefit of built-in surge protection, this article should be placed at the top of your project list. For about $10 in parts and six hours of bench work, you can breathe easy when you push the power switch. I call it the $10 insurance policy.

The subject of surge protection has been addressed by many in the past few years. In my opinion, one of the better articles was written by K. M. Gleszer, W1KAY, entitled "Upgrading Your SB-220 Linear Amplifier," which appeared in *QST*, February, 1979. Specific solutions were offered for operation with 117-Vac for filament inrush current, diode-transient and voltage-equalization protection, plus other items. But conspicuous by its absence was a scheme for diode inrush current protection. This protection is easily obtained with the simple circuit described here.

One other area where I'd suggest a change is the time-delay relay. The time-delay function is automatic with a standard relay coil and a current-limiting resistor. Therefore the high cost, plus purchase time and final alteration, of a time-delay relay can be avoided.

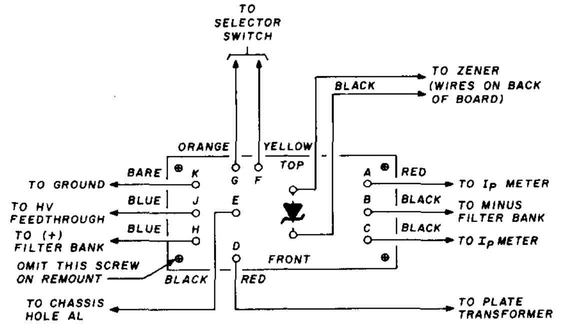


fig. 1. SB-220 rectifier board.

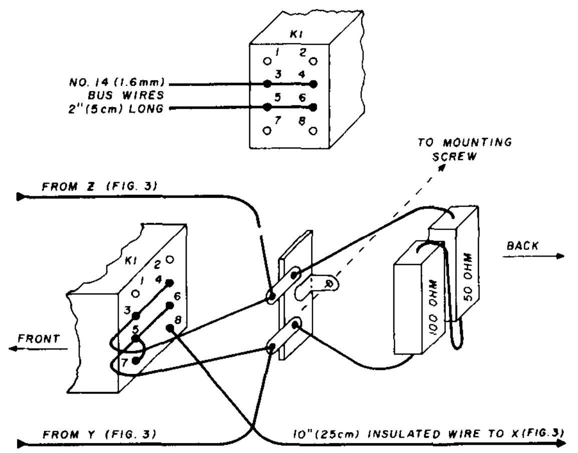


fig. 2. Relay connections for the surge-protection circuit.

By F. T. Marcellino, W3BYM, 13806 Parkland Drive, Rockville, Maryland 20853

The mods I've installed are not unfamiliar, as they've appeared in several 1970-series of the *Radio Amateur's Handbook*. However, I've described the procedures in a detailed order using short, sometimes elementary, phrases for clarification. I'm a stickler for the smallest detail, so you needn't bother with assumptions.

With the mods installed, the following benefits will be added to your SB-220:

1. Rectifier transient surge protection.
2. Rectifier reverse voltage equalization.
3. Rectifier inrush current protection.
4. Inrush current protection for the 3-500Z filaments.

This procedure is divided into two parts: rectifier protection and surge protection. You can elect to cancel one, but because the amplifier must be uncaged for installation of either, it seems wise to include both.

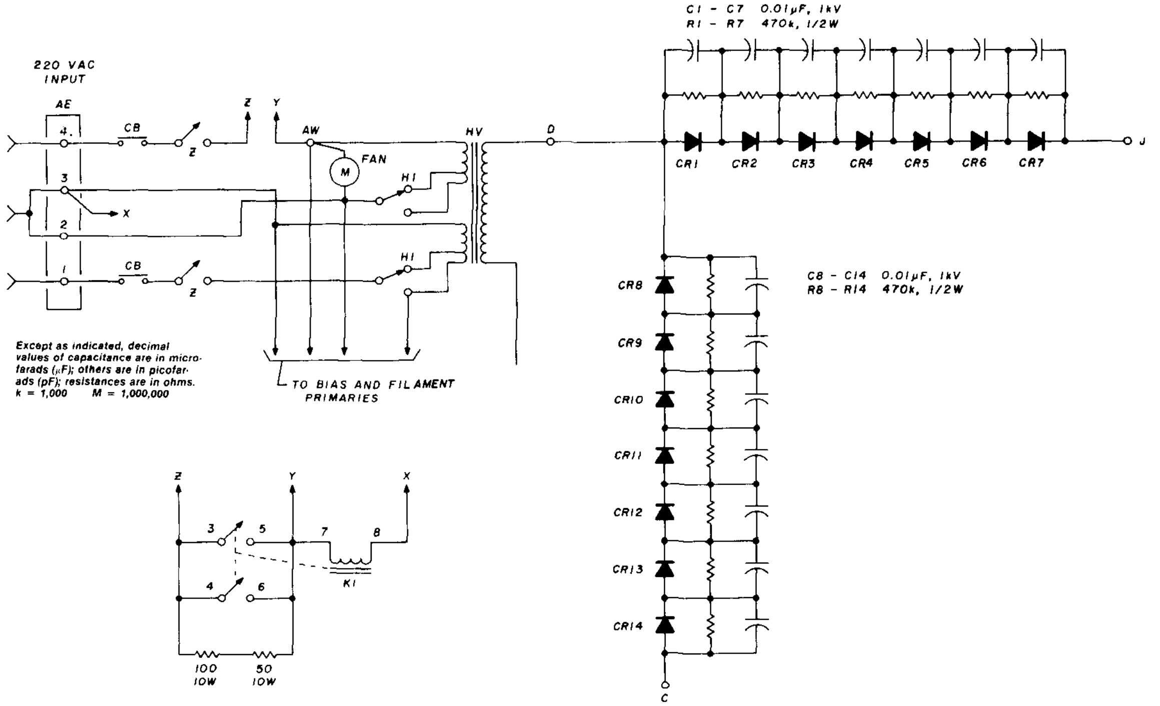
The fourteen original diodes in the SB-220 were not replaced with higher PIV units. This action is not necessary unless you break some during disassembly. These diodes are rated for 1 ampere average forward current at a PIV of 600 volts. The ratings are adequate for this application, and, combined with the modification, they will have a long life.

The nominal delay was selected as 5 seconds. This time can be altered by varying the total limiting resistance. A resistance of 200 ohms caused a long delay, and the resistors dissipated much power. At the opposite extreme, 100 ohms provided insufficient delay. Therefore, a satisfactory value of 150 ohms was selected. Note that the time delay and resistance values were selected using a line voltage of 220 Vac. I intended to operate this linear only on the higher line voltage for increased efficiency.

rectifier protection

1. Remove amplifier case, top shield cover, and right-side shield.
2. Remove the four rectifier board hold-down screws.
3. Make a wiring map of all twelve wires connected to the rectifier board and identify by color designator (**fig. 1**).
4. Unsolder all twelve wires at the board end, then remove diodes.
5. Wick twelve wire pads and all diode holes. Remove flux.
6. Drill out all *diode* holes using a No. 47 (2 mm) drill bit from the pad side of the board (assuming all boards are the same).
7. Using a No. 15 (4.5 mm) drill bit, deburr the new holes from the component side. Do not deburr the pad side.
8. Install resistors (470 k ½w) from the pad side, then install diodes and capacitors (0.01 at 1 kV) from the component side. Next:
   - a. Solder each pad with its three wires.
   - b. Clip component pigtails as you go.
   - c. Clean board to remove flux.
   - d. Ohmmeter check—note highs will be 470 k.
9. Connect board to SB-220 using the following sequence:
   - a. Solder red wire to hole D.
   - b. Solder blue wires at holes H and J.
   - c. Mount board using three screws—omit lower LH.
   - d. Solder bare wire at hole K.
   - e. Solder black wire at hole E.
   - f. Solder black wires to holes and pads for the zener. Observe proper polarity.
   - g. Solder orange wire to hole G.
   - h. Solder yellow wire to hole F.
   - i. Solder red small wire to hole A.
   - j. Solder black wire (minus filter bank) to hole B.
   - k. Solder black wire (I_p meter) to hole C.

This completes the rectifier-board wiring. Dress all wires at right angles away from the board, then



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 3. SB-220 surge modifications.

10. Reinstall right-side shield.

11. Oil felt pads on fan motor while top cover is off.

12. Install top shield cover.

13. Test the amplifier using a dummy load.

14. If OK, proceed to the next section.

surge protection

1. Solder No. 14 (1.6 mm) bus wire 2 inches (5 cm) long to pins 3 and 4 of relay K1 (**fig. 2**).

2. Solder No. 14 (1.6 mm) bus wire 2 inches (5 cm) long to pins 5 and 6 of relay K1.

3. Bend the two wires and solder to a two-lug tie strip.

4. Connect pin 5 to 7 using No. 20 (0.8 mm) bare wire.

5. Connect a black insulated wire (rated for 220 Vac, 10 amperes) about 10 inches (25 cm) long to K1 pin 8.

6. Stack the two current-limiting resistors (100 and 50 ohms) and connect in series. Solder this pair to the lower holes in the tie strip.

7. Mount the completed surge-protection into the SB-220 using the center ground lug on the tie strip and the existing chassis screw located about 2 inches (51 mm) forward of terminal strip AE. The relay case should rest against the chassis, being supported by the bus wires.

8. Connect the 10-inch (25-cm) black insulated wire (trim as required) from relay K1 pin 8 to terminal 2/3 on terminal strip AE of the linear.

9. Remove existing black jumper wire between power switch Z and front standoff AW.

10. Connect Z to pins 3 and 4 of K1 using the tie strip. Use insulated wire with (220 Vac, 10-ampere rating).

11. Connect Y from standoff AW to pins 5 and 6 using the tie strip. Use insulated wire with 220-Vac, 10-amp rating.

12. This completes the surge relay installation.

From the Heathkit manual, these codes are used:
AE 110/220 Vac input terminal strip.
AW front-mounted standoff tie point.
AL front corner hole.
Z power switch.

operation

Checkout of the surge protection circuit can be monitored each time the linear is fired up, assuming the filter capacitors have discharged to a low level. Place the selector switch in the HV position, while the mode switch can be in either the CW/TUNE or SSB position. After the power switch is pushed, there will be a time period of a few seconds of dead silence. This delay time is controlled by the value of the limiting resistors. During this period the plate voltage meter can be observed to slowly increase from zero to about 1500 Vdc. Additionally, the meter illumination lamps will *slowly* energize to about half brilliance. Since the 3-500Z filaments are in parallel with these lamps, they will be responding in the same way. If in doubt, turn off your room lights while energizing the linear and peer down through the case top.

The cooling fan will be turning very slowly while gradually building up speed. Therefore there will be no noise from this source during the initial few seconds.

After the five-second surge-delay period, adequate voltage will be available for surge relay K1 to pull in. During a brief interval K1 contacts will close and hold, thus shorting the limiting resistors and applying full line voltage to the transformers. Instantly the plate voltage will increase from 1500 Vdc to its normal maximum value. The 3-500Z filaments will glow with their normal brilliance, and the cooling fan will attain maximum speed. Don't be alarmed when you hear a brief buzzing sound as the relay closes. This sound is caused by K1 contacts bouncing (as all mechanical relays do) combined with slight inductive arcing.

Although this article is written specifically for the SB-220, other similar equipment could be surge protected using these mods.

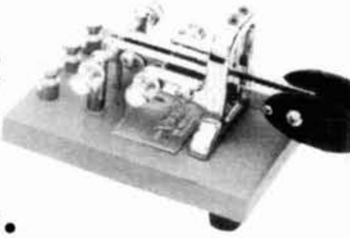
For additional information on rectifier diode protection I suggest the April, 1980, edition of *Worldradio*, which has a fine article written by Joe Carr, K4IPV.

Once you've installed the mods as shown in **fig. 3**, you can place the problem of surge protection on the shelf for a well-deserved rest. I've used these circuits on two other homebrew linear amplifiers with total success. In addition I've used them on power supplies for several transmitters using the lower line voltage. The only difference is the selection of the limiting resistance for a satisfactory delay period.

Note: K1 is a dpdt relay, 5000-ohm coil, 120 Vac. Contacts are rated at 10A, 125 Vac. Dimensions: 1-5/8 x 1 x 3/4 inches (41 x 25.4 x 19 mm).

ham radio

The Iambic Keyer Paddle.



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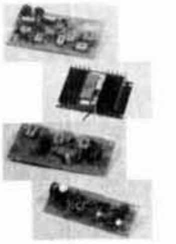

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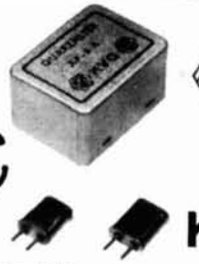

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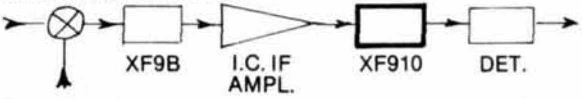

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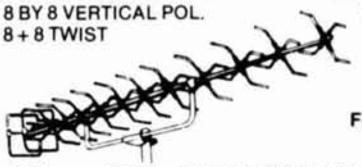

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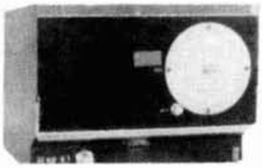
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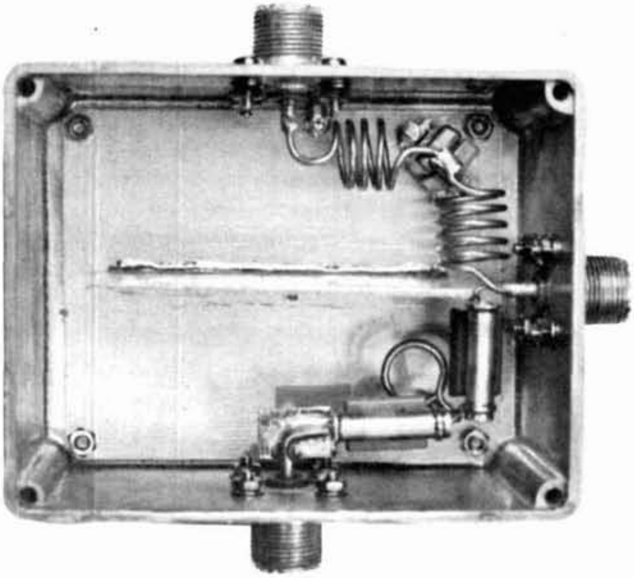
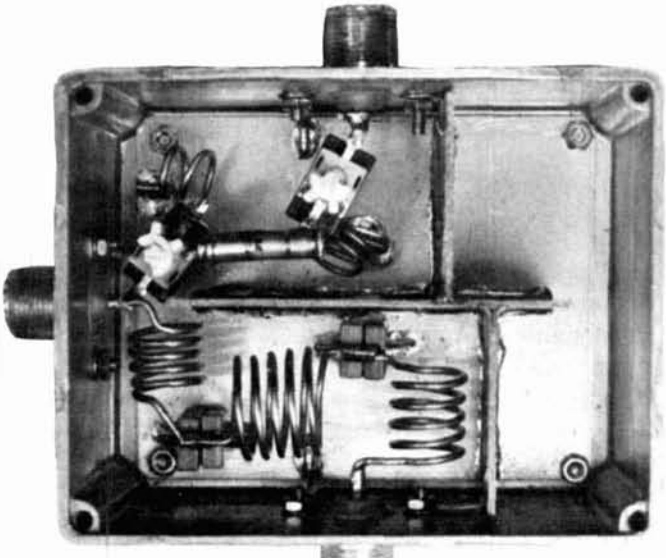
transceiver diplexer:

an alternative to relays

Frequency-selective filters allow vhf and hf antennas to share a common feedline





Diplexer consists of a matched pair of highpass and lowpass filters, which allow a vhf and high-frequency antenna to share a common feedline. The filter at the antenna end has three sections (*top*); that at the station end has five sections (*bottom*).

In many cases it's desirable to reduce the number *of feedlines between the ham station and the antennas.* One of the more important reasons is the price of high-quality coax cable. It's easy to spend as much money on transmission lines as on a small commercially manufactured 2-meter Yagi antenna. A second reason may be the need to tidy up your installation to please neighbors. If antenna restrictions exist in your area, and you're trying to avoid detection, the presence of several coax cables can be too much to hide.

One of the more popular ways of making the best use of feedlines is to use switching relays at the

By Terry A. Conboy, N6RY, 2631 S.W. Orchard Hill Place, Lake Oswego, Oregon 97034

antennas to select the desired antenna. Several systems to accomplish this are available commercially, and homebrewing such an arrangement is not technically difficult.

There are disadvantages to such schemes. What happens when you're chasing a rare station and still want to listen to the local DX repeater on 2 meters? If you have only one feedline, this can be inconvenient. Care must be taken to avoid transmitting on the wrong-frequency antenna to prevent possible damage to both transmitter and antenna.

enter the diplexer

An alternative to relays is frequency selective networks to select the proper antenna automatically. The networks can also allow simultaneous combination of more than one transceiver on the same coax cable.

These networks are called *diplexers*, since they allow two transmitters (or receivers) to use the same feedline at the same time. They differ from *duplexers*, as used in repeaters. Duplexers permit simultaneous operation of one transmitter and one receiver on a common antenna.

Although possible, it would be difficult to construct networks that would permit several different high-frequency antennas to share the same feedline. Relays are probably best used for this purpose. Because 144 MHz and 220 MHz are commonly used for local communications, I designed a simple network to permit either of these vhf bands to coexist with high-frequency signals on one coax cable. I did not include the 50-MHz band because this design would have required more complex networks. (The 420-MHz band will pass through the filters, but the impedance match is marginal.)

I used two networks. The one at the station end (**fig. 1**) allows both the high-frequency and vhf rig to access the coax simultaneously. The network at the antenna end (**fig. 2**) does the same for the high-frequency and vhf antennas. Each network consists of a mated pair of highpass and lowpass filters to accomplish the separation and combination of the two different frequencies.

Some disadvantages occur in the use of filters to perform these functions. A small amount of loss is added to the system. This is minimal, however. Also the impedance presented to the transceivers is modified. By proper filter design this mismatch can be kept to a minimum.

One added benefit of the filters should be noted: Lowpass filters are in the circuit to the high-frequency antenna, so some reduction in harmonic radiation is evident, which may reduce TVI to the point that an additional filter isn't needed.

designing the filters

The highpass and lowpass filters are simple Chebychev units that can be designed from tables of normalized filter prototypes or by calculating normalized inductor and capacitor values. I found it easier, however, to use the network design programs available on the engineering computer at my place of employment.

Reflection coefficient. To minimize the amount of mismatch introduced by the filters, I designed them to have a maximum reflection coefficient of 0.065. Since two filters are in tandem, the worst-case reflection coefficient with a 50-ohm load could be twice this amount, or 0.13, which corresponds to a maximum SWR of 1.3. The worst-case situation at the transmitter for a load with a 2-to-1 SWR would be SWR of 2.7. Because of the designs I used, the frequencies of worst match don't coincide, and such a degradation is unlikely. The match may also be better at some frequencies because of the small variations in the impedance transformation through the filters.

Cutoff frequencies. I set the filter cutoff frequencies about 7 per cent above and below the required maximum and minimum frequencies to avoid the loss appearing near the filter corners caused by the finite *Q*'s of the inductors. The resulting cutoff frequencies were 32 MHz for the lowpass filters and 135 MHz for the highpass filters.

Isolation. The number of filter sections is governed by the isolation required between high-frequency and vhf equipment. Isolation at the transceivers must be much greater than at the antennas. For protection against receiver overload, at least 50 dB isolation was desired between the high-frequency transmitter and the vhf receiver. Such isolation reduces 1000 watts to 10 milliwatts at the receiver front end. Because of the wide frequency separation, no undesirable intermodulation occurs in the vhf receiver.

The isolation required between vhf transmitter and high-frequency receiver is usually not as great, because most stations use much lower power on vhf than on hf. Even so, I designed the filters to be symmetrical, which should give the same isolation in both directions.

At the antennas, I set the isolation at 30 dB. This isolation should prevent high-frequency-antenna radiation from causing any significant reduction in the front-to-back ratio of a directional vhf antenna.

To obtain the desired isolation, I made the networks at the station end with five sections each and those at the antennas with only three sections each.

After I designed the filters I increased the reac-

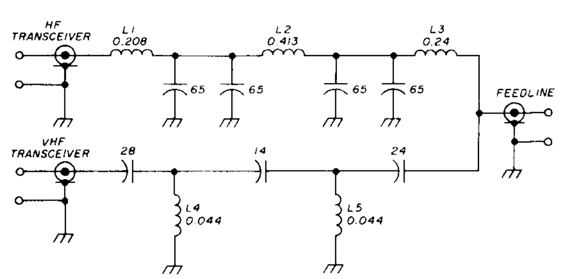


fig. 1. The five-section filters used at the transceivers. Inductor data is given in table 1.

tances of the components at the common port by the same ratio to compensate for the shunting effect of the other filter. I did this with an interactive network analysis program. To make the impedance match as good as for the highpass or lowpass filter alone, I increased the end inductor of the five-section lowpass filter by 15 per cent and decreased capacitor of the five-section highpass filter by the same amount. For the three-section filter, the change of the end components was 30 per cent.

I made allowances for the parasitic capacitances of the inductors to ground in the lowpass sections, which add in parallel with the shunt capacitors. I made allowance of 3 or 4 pF in the capacitors I used. I added small metal tabs about 0.4 inch (1 cm) square to the highpass filters. This restored symmetry to the highpass sections and improved the match at 220 MHz. The final design of the filters appears in **Figs. 1** and **2**.

construction

The filters were built in cast aluminum boxes and a piece of unetched copper-clad PC board was attached to the inside of the box with machine screws. The shunt components were then soldered directly to the copper board with the shortest possible leads. The series components were supported by the shunt components (this arrangement can be seen in the photos). This construction provides a rigid mounting for the parts with minimal stray inductance and capacitance.

Shields were placed between the highpass and lowpass filters in each box to reduce mutual coupling. If you don't include the shields, isolation between the vhf transmitter and high-frequency receiver will be seriously impaired.

For the five-section networks, additional shields were required. The shields were made of double-sided copper board. They were soldered all along the seams together with the groundplane copper boards and the other shields, then fastened to solder lugs on the connectors where possible.

All coils were placed at right angles to each other in the same shielded area to avoid mutual coupling, which can cause filter performance to depart drastically from the theoretical predictions.

components

The fixed caps were micas with a 1000-volt rating. This rating is adequate for power levels up to the legal limit. Because of the high currents flowing in the shunt capacitors in the lowpass filters, the required capacitance was obtained by using two capacitors in parallel, which reduces any possible heating in the capacitors. Currents are highest when operating near the filter cutoff frequency and can easily reach 5 amperes with 1000 watts of input power.

Air variable capacitors could be used throughout, in place of the micas, provided the voltage rating is adequate. In the highpass sections, the micas were paralleled with air variables and glass piston capacitors to allow tuning. After the filters were tuned, it appeared that fixed units of the calculated values would have worked just as well, as judged from the positions of the variables.

All the inductors were wound of No. 12 (2.1-mm) tinned copper wire. Winding data were obtained from charts in the ARRL *Handbook*. Information on the dimensions of the coils appears in **table 1**.

tuning the filters

By far the best way to tune Chebychev filters is with a swept reflectometer. These filters were tuned this way, adjusting the coils by stretching and squeezing and by tuning the capacitors until the impedance match across the passband of each filter was within the desired limits. Not everyone has the facilities to adjust the networks in this manner. As an alternative, the filters should be adjusted one at a time into a dummy load with an SWR meter or a noise bridge set to 50 ohms. The frequencies to use are given in **table 2**. It's important not to vary the components too far from the calculated values; do-

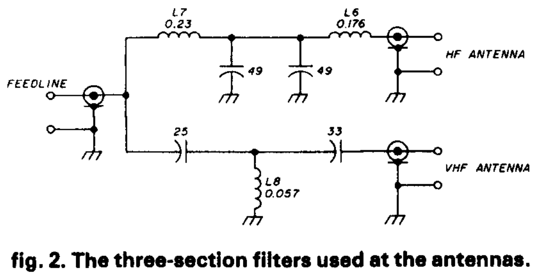


fig. 2. The three-section filters used at the antennas.

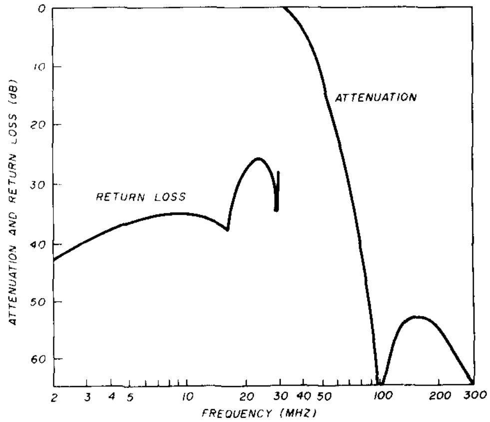


fig. 3. Measured frequency response and return loss for the five-section lowpass filter.

ing so may cause the isolation to be upset.

After tuning for best match at the frequencies indicated, check the match at other frequencies within the filter passbands. It may be necessary to retune somewhat if the impedance match is poor. Remember that the match should not necessarily be perfect at all frequencies, but the SWR should not be worse than 1.2 anywhere in the passband of either filter.

table 1. The inductors should be wound according to this data. The wire used is solid No. 12 (2.1 mm) with spacing between the turns equal to the wire diameter.

inductor	nominal inductance µh	inside diameter inches	(mm)	no. turns
L1	0.208	0.5	(12.7)	4.5
L2	0.413	0.75	(19.0)	5.0
L3	0.24	0.5	(12.7)	5.25
L4	0.044	0.5	(12.7)	1.25
L5	0.044	0.5	(12.7)	1.25
L6	0.176	0.5	(12.7)	4.0
L7	0.23	0.5	(12.7)	5.0
L8	0.057	0.375	(9.5)	2.0

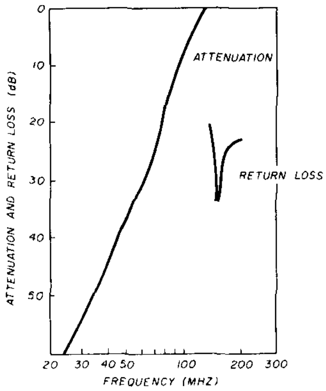


fig. 4. Measured frequency response and return loss for the five-section highpass filter.

table 2. Adjust the inductors (and variable capacitors, if used) for best match into a 50-ohm load at these frequencies.

filter	adjustment frequency (MHz)
5-section lowpass	28.3
5-section highpass	148.0
3-section lowpass	28.0
3-section highpass	147.0

diplexer performance

The two networks were measured with 50-ohm terminations on the unused ports. The results of the measurements are given in **figs. 3** through **8**. The impedance match is plotted as *return loss*. This quantity is 20 times the logarithm of the magnitude of the reflection coefficient. It was measured directly by the test equipment used. The reflection coefficient for which the filters were designed, 0.065, represents a

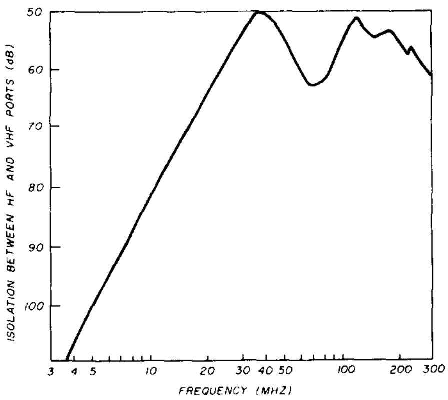


fig. 5. Measured isolation between the high-frequency and vhf ports of the five-section filters. The common port was terminated with 50 ohms.

table 3. These are actual measured losses in a 50-ohm circuit with the unused ports terminated. Resistive and mismatch losses are included.

filter	maximum loss (dB)	frequency (MHz)
5-section lowpass	0.1	21.0
5-section highpass	0.22	220.0
3-section lowpass	0.07	28.0
3-section highpass	0.05	225.0

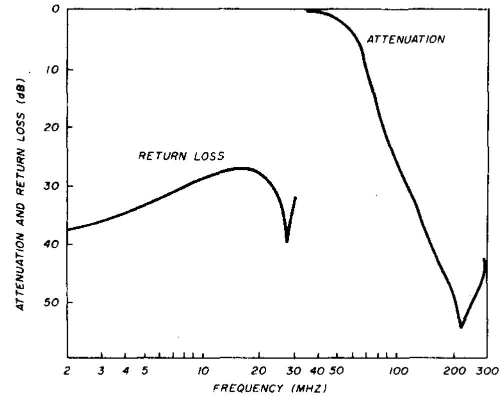


fig. 6. Measured frequency response and return loss of the three-section lowpass filter.

return loss of 23.7 dB and an SWR of 1.14. The diplexer insertion loss was surprisingly low. **Table 3** summarizes the measured losses through the filters.

Use of the filters shows that the isolation between the high-frequency and vhf equipment is more than adequate. The equipment was a Yaesu FT-301 with an FL-2100B and an Icom IC-22S. The only problem areas were at harmonics of the high-frequency transmitter that fell on frequencies in the 2-meter band. However, this was also a problem when operating with separate feedlines. Significant fifth-harmonic energy was picked up by the 2-meter transceiver even when it and the high-frequency transmitter were connected to dummy loads.

possible improvements

The layout of the filters would be much better if the boxes were long and narrow, with the common connection near the center of the assembly. Then the high-frequency and vhf ports would be separated by the greatest distance. Another layout improvement would be to shield separately each inductor in its own small compartment. This would greatly reduce mutual coupling between the coils.

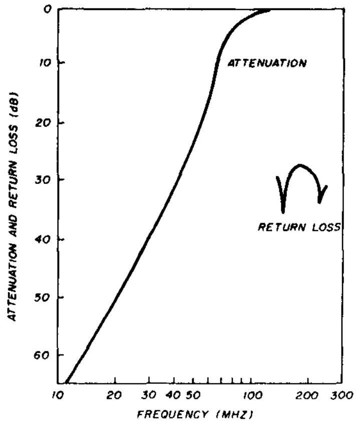


fig. 7. Measured frequency response and return loss of the three-section highpass filter.

The other possible improvement is to reduce the effective stray inductance of the shunt capacitors in the lowpass filters by paralleling more than two capacitors to obtain the required value. The self-resonant frequency of smaller capacitors would be moved higher in frequency, and the stopband attenuation and isolation would be greater.

using the diplexers

If antenna tuners or TVI filters are in use at your station, they must be placed between the transceiver and the diplexer, which can be a problem if the antenna tuner is used to compensate for fairly high standing-wave ratios. Possible voltage and current stresses on the components in the filters could easily damage them. It would be wise to restrict operation at maximum legal power to standing-wave ratios no higher than 2.5 on the main feedline.

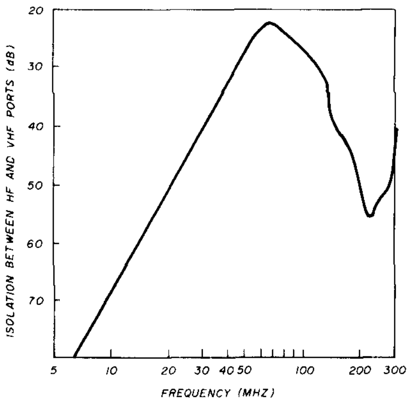


fig. 8. Measured isolation between the high-frequency and vhf ports of the three-section filters. The common port was terminated with 50 ohms.

For normal exciter power levels (under 300 watts input), there should be no problem with standing-wave ratios up to 5 under normal use, especially below the 20-meter band.

If your SWR meter is capable of operation on both hf and vhf, it may be placed in the common feedline and measurements can be made in either frequency range.

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Gainesville	Lafayette Radio
Hialeah	Heathkit Electronic Center
Jacksonville	Laraco of Florida Inc.
Lakeland	Lakeland Specialty Electronics
Miami	Herman Electronics
Miami	N&G Distributors
Oakland Park	Lafayette Radio
Orlando	Southeast Micro-Data Corp.
Pensacola	Grice Electronics Inc.
FLORIDA (Continued)	
Plantation	Heathkit Electronic Center
Tampa	AMF Microcomputer Center Inc.
Tampa	Heathkit Electronic Center
Tampa	Microcomputer Systems
GEORGIA	
Atlanta	Atlanta Computer Mart
Atlanta	CompuShop
Atlanta	Heathkit Electronic Center
Columbus	Radio Wholesale
Stone Mountain	Coleman's Electronics
HAWAII	
Honolulu	Industrial Electronic Inc.
Honolulu	Integrated Circuit Supply
Wailuku	Q.S.K. Electronics
IDAHO	
Boise	Custom Electronics
Caldwell	A-Gem Supply Inc.
ILLINOIS	
Carbondale	Picks Electronics
Chicago	Howard Electronic Sales
Chicago	Olson Electronics
Evanston	Tri-State Electronic Corp.
Geneva	Melvin Electronics
Groveland	Moyer Electronics
Mount Prospect	Tri-State Electronic Corp.
Niles	Computer Land
Peoria	Computer Land of Peoria
Peoria	Warren Radio Co.
Rockford	Computer Store of Rockford
Schaumburg	Data Domain of Schaumburg
Skokie	Lilipute Computer
Villa Park	Melvin Electronics
INDIANA	
Bloomington	Stansifer Radio Co. Inc.
Chesterton	Chesterton Electronics
East Chicago	Acro Electronics Corp.
Evansville	Hutch & Son Inc.
Gary	Calumet Electronic Supply
Indianapolis	Heathkit Electronic Center
Muncie	Pierce Electronics
South Bend	Radio Distributing Inc.
South Bend	Television Supply Co.
IOWA	
Ames	Electronic Supply Inc.
Davenport	Memory Bank
Des Moines	Gifford Brown Incorporated
Des Moines	Radio Trade Supply Co.
KANSAS	
Hutchinson	Hutchinson Electronics
Kansas City	Electronic Surplus Sales
Mission	Heathkit Electronic Center
Salina	Electronics Inc.
Wichita	Amateur Radio Equipment Co.
KENTUCKY	
Lexington	Radio Electronic Equipment Co.
Louisville	Peerless Electronic Equipment Co.
Richmond	Madison County Electronics
LOUISIANA	
Baton Rouge	Davis Electronics Supply Co.
Baton Rouge	Menard Electronics Inc.
Baton Rouge	Pelican Electronics
Kenner	Heathkit Electronic Center
Lake Charles	Gulf Power & Controls
Lake Charles	Wholesale Radio & Equipment
New Orleans	Wm. B. Allen Supply Co.
Shreveport	Industrial Electronic Supply
MARYLAND	
Annapolis	Computers Etc.
Annapolis	J&M Electronics Inc.
Baltimore	American Distributing Co.
Baltimore	Everything Electronic
Baltimore	Harco Electronics
Baltimore	Tritronics, Inc.
Churchville	Churchville Electronics
Damascus	Damascus CB
Frederick	Frederick Computers
Glen Burnie	Revacto
La Vale	J&M Electronics
Rockville	Heathkit Electronic Center
Silver Spring	Computers Etc.
Towson	Baynesville Electronic Inc.
Towson	Computers Etc.
Towson	Computers Unlimited
Towson	Heathkit Electronic Center
Wheaton	Computer Crafters
MASSACHUSETTS	
Littleton	Tel-Com Inc.
Peabody	Heathkit Electronic Center
Pittsfield	Pittsfield Radio Co. Inc.
Waltham	Computer Mart Inc.
Wellesley	Heathkit Electronic Center
West Springfield	Norbill's Electronics Inc.
Worcester	RM Electronics Inc.
MICHIGAN	
Adrian	E&B Electronics Inc.
Allen Park	Electronic Parts Co.
Ann Arbor	Wedemeyer Electronic Supply
Battle Creek	Warren Radio Co.
Bay City	Kinde Distributing Co.
Canton	The Electronic Connection
Clawson	Radio Supply & Engineering Co.
Coldwater	B&M Electronics
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Detroit	Radio Supply & Engineering Co.
Detroit	S&S Electronics
Oxford	Oxford Software Co.
East Detroit	Heathkit Electronic Center
Flint	Shand Electronics Inc.
Garden City	Computer Center
Grand Rapids	Micro Computer World
Grand Rapids	Radio Parts Inc.
Grand Rapids	T&W Electronics
Grand Rapids	Warren Radio Co.
MICHIGAN (Continued)	
Jackson	Fulton Radio Supply
Kalamazoo	Warren Radio
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Livonia	Norwest Electronics
Livonia	Radio Supply & Engineering Co. Inc.
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Muskegon	H.R. Electronics
Niles	Niles Radio Supply
Pontiac	Electronic Supply
Port Huron	Main TV Radio Electronics Inc.
Saginaw	Electronic Mart
St. Claire Shores	Bell Electronics
St. Joseph	Heath Customer Center
Taylor	Electronic World
Traverse City	Traverse City Electronic Supply
Westland	Olson Electronics
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Duluth	Northwest Radio of Duluth Inc.
Hopkins	Heathkit Electronic Center
Minneapolis	Pal Electronics
St. Paul	Heathkit Electronic Center
Winona	Hiawatha Electronics Inc.
MISSISSIPPI	
Jackson	Ellington Electronic Supply
MISSOURI	
Bridgeton	Forsythe Computers
El Dorado Springs	Beckman Electronics
Florissant	Computer Country
St. Louis	Computer Country (Downtown)
MONTANA	
Billings	Conley Radio Supply
Bozeman	Electronic Service & Dist. Inc.
Great Falls	Arts Electronics
NEBRASKA	
Lincoln	Computer Systems
Lincoln	Scott Electronic Supply Inc.
Norfolk	Caauwes Gun & Hobby Shop
North Platte	Scott Electronic Supply Corp.
Omaha	Midwest Computer
Omaha	Scott Electronic Supply Corp.
Omaha	Heathkit Electronic Center
NEVADA	
Las Vegas	Century 23
NEW JERSEY	
Cherry Hill	The Computer Emporium
Englewood	Radiomasters
Fair Lawn	Heathkit Electronic Center
Ocean	Heathkit Electronic Center
Pennsauken	Lafayette Radio
Pompton Lakes	Computer Corner
Ramsey	Typetronic Computer Store
Trenton	Lafayette Radio Electronics
Vineland	Lafayette Radio Electronics
NEW YORK	
Albany	Action Audio Inc.
Albany	Fort Orange Electronics
Albany	Greylock Electronics
Amherst	Heathkit Electronic Center
Buffalo	Radio Equipment Corp.
Cortland	Action Audio Inc.
Endwell	Computer Tree Inc.
Ithaca	Action Audio Inc.
Ithaca	Rad-Tronics
Jamestown	Warren Radio
Jericho	Heathkit Electronic Center
Kingston	Action Audio Inc.
Kingston	Greylock Electronics
Middleton	Action Audio Inc.
Newburgh	Action Audio Inc.
New York City	Aristo Craft
New York City	Digibyte Systems Corp.
New York City	Taft Electronics 45 Corp.
Orchard Park	Radio Equipment Corp.
Peekskill	Action Audio Inc.
Poughkeepsie	Greylock Electronics
North White Plains	Heathkit Electronic Center
Rochester	Heathkit Electronic Center
Troy	Trojan Electronics
Utica	Am-Com Electronics
Utica	Central Electronics
Wappingers Falls	Action Audio Inc.
White Plains	The Computer Corner
NORTH CAROLINA	
Raleigh	Byte Shop of Raleigh
NORTH DAKOTA	
Fargo	S&S Electronics
OHIO	
Cincinnati	Heathkit Electronic Center
Cincinnati	Micro-Solution
Cleveland	Heathkit Electronic Center
Columbus	Heathkit Electronic Center
Dover	T.V. Specialties
Fremont	Terra Tech College Bookstore
Reynoldsburg	Universal Amateur Radio
Toledo	Heathkit Electronic Center
Toledo	Warren Radio
Wickliffe	Amateur Electronic Supply
Youngstown	Ross Radio Company
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Oklahoma City	Bits, Bytes & Micros
Oklahoma City	Heathkit Electronic Center
Oklahoma City	Trice Wholesale Electronics
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Beaverton	Norvac Electronics
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OREGON (Continued)	
Salem	Comm-Shack
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Chambersburg	Sunrise Electronic Dist. Co.
Drexel Hill	Kass Electronics Distributors
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Frazer	Heathkit Electronic Center
Paoli	Personal Computer Corporation
Philadelphia	Heathkit Electronic Center
Philadelphia	Resco Electronics
Phoenixville	Stevens Electronics
Pittsburgh	Heathkit Electronic Center
Pittsburgh	South Hills Electronics
Wilkes Barre	Hamline Electronics
York	G.Y.C. Company
RHODE ISLAND	
Cranston	Jabbour Electronics City
Pawtucket	Jabbour Electronics City
Warwick	Heathkit Electronic Center
TENNESSEE	
Cookeville	Wagnon's Stereo Center
Knoxville	Shields Electronics Supply Inc.
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Memphis	Sere-Rose & Spencer Electronics
Memphis	Warren Radio Company
Nashville	Eddie Warner's Parts Company
Nashville	Electra Distributing Co.
Oakridge	National Electronics
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Beaumont	Electro-Hut
Brownsville	George's Electronics Mart
Dallas	Heathkit Electronic Center
Dallas	Ram Micro Systems
Fort Worth	The Ingenuity Concept
Garland	Tinker Tronics Inc.
Houston	Heathkit Electronic Center
Houston	Interactive Computers
Lubbock	Trice Wholesale Electronics
McAllen	Carlos Franco Electronics
San Antonio	Appliance & Equipment Co. Inc.
San Antonio	C&K Electronics Inc.
San Antonio	Heathkit Electronic Center
Waco	L&M Wholesale Electronics
UTAH	
Midvale	Heathkit Electronic Center
Provo	Alpine Electronic Supply Co.
Salt Lake City	Best Distributing
Salt Lake City	Computerland of Salt Lake
VERMONT	
Essex Junction	Lafayette Radio
VIRGINIA	
Alexandria	Computers Plus Inc.
Alexandria	Heathkit Electronic Center
Annadale	Arcade Electronics
Arlington	Arlington Electronic Wholesalers
Blacksburg	Scotty's Radio & TV Inc.
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Falls Church	Crossroad Electronics
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Hopewell	B&G Electronics
McLean	Tyson's Computer Emporium
Norfolk	Avec Electronics Corp.
Norfolk	Priest Electronics
Portsmouth	Electronics Unlimited
Richmond	Avec Electronics Corp.
Roanoke	Avec Electronics Corp.
Vienna	Electronic Equipment Bank
Virginia Beach	Heathkit Electronic Center
Warrenton	Radio Shack
Woodbridge	Virginia Micro Systems
WASHINGTON	
Bellevue	ABC Communications
Everett	ABC Communications
Kennewick	C&J Electronics Inc.
Longview	Progress Electronics
Moses Lake	Ron's Electronics
Pasco	Riverview Electronics
Richland	C&J Electronics
Seattle	ABC Communications
Seattle	Amateur Radio Supply
Seattle	Empire Electronics
Spokane	Don's Stereo Center
Spokane	Personal Computers
Tacoma	C&G Electronics
Tacoma	Northwest Radio Supply
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Morgantown	The Computer Corner
Morgantown	Electro Distributing Co. Inc.
Wheeling	Lafayette Radio Associate Store
WYOMING	
Cheyenne	Computer Concepts
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Calgary	CompuShop Alberta Ltd.
Ontario:	
Toronto	House of Computers
GUAM	Marianas Electronics
ENGLAND	
Berkshire	NewBear Computing Store
GUATEMALA	Electronica Pan-Americana
PANAMA	Tropelco, S.A.
SINGAPORE	Applied Digital System
SINGAPORE	Sys-Tech
SWEDEN	
Stockholm	LSI Electronics HB
WEST INDIES	
Port of Spain, Trinidad	The Hobby Centre

For Distributor Information, write or phone JIM-PAK® 1355 Shoreway Road, Belmont, CA 94002 (415) 595-5936

Jim-pak
electronic components

One-Stop Component Center (TM)

This is a partial listing of over 500 items available from authorized Jim-pak Distributors:

TTL

7400	2/.85	7490	.85
7402	2/.85	7493	.85
7404	2/.85	74100	2.25
7406	2/1.19	74109	2/1.19
7407	2/1.19	74121	.69
7408	2/.89	74123	.99
7410	2/.85	74150	1.95
7414	.99	74154	1.95
7417	2/1.10	74157	.99
7420	2/.85	74161	1.19
7447	1.19	74164	1.59
7474	.69	74174	1.59
7475	.79	74175	1.49
7476	.69	74192	1.19
7485	1.19	74193	1.19
7486	2/1.19	74367	.99
7489	2.99	74393	1.95


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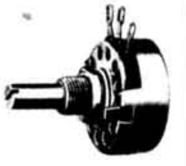
GB100	Cer. Caps. (100)	2.95	GB123	Heat Sinks (30)	3.95
GB101	Mylar Caps. (60)	4.95	GB127	Transistors (100)	3.95
GB102	Electrolytics (60)	4.95	GB137	Chokes (50)	3.95
GB103	Tantalums (40)	4.95	GB139	Term. Strips (40)	3.95
GB108	TTL IC's (50)	4.95	GB140	Spcr/Stndff (150)	2.95
GB110	Asst. LEDs (100)	5.95	GB141	Washers (200)	2.95
GB113	Trimmers (30)	4.95	GB145	Lugs (100)	2.95
GB116	¼w Resist. (200)	2.95	GB154	1&2w Resist. (100)	2.95
GB117	½W Resist. (200)	2.95	GB162	7-Seg. Dsplys. (50)	5.95
GB120	Slide Switch (25)	3.95	GB173	3/8" Pots. (100)	5.95

LS Schottky

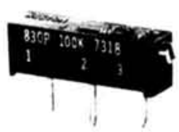
74LS00	.55	74LS109	.79
74LS02	.55	74LS123	1.95
74LS04	.69	74LS138	1.49
74LS08	.55	74LS139	1.49
74LS10	.55	74LS154	2.49
74LS14	1.09	74LS157	1.49
74LS30	.55	74LS161	1.79
74LS32	.69	74LS174	1.79
74LS38	.69	74LS175	1.79
74LS42	1.49	74LS192	1.89
74LS47	1.49	74LS193	1.89
74LS48	1.79	74LS221	1.95
74LS73	.79	74LS244	2.49
74LS74	.79	74LS245	3.49
74LS75	.99	74LS367	1.29
74LS85	1.95	74LS374	2.49
74LS90	1.09	81LS97	2.49

POTENTIOMETERS


2 Watt @ 70°C
7/8" Slotted Shaft
Linear Taper

1K 5K 10K
25K 50K 100K
1 Meg

CMU . . $2.95



3/4 Watt @ 70°C
15 Turn Pot.
Linear Taper

100Ω 500Ω 1K
5K 10K 50K
100K 500K 1Meg

830P . . $1.79

REGULATED POWER SUPPLY KIT


Uses LM309K. Heat Sink provided. PC board construction. Provides a solid 1 amp @ 5 volts. Can supply up to ±5V, ±9V and ±12V with JE205 Adapter. Includes components, hardware and instructions. 3½"x5"x2"H

JE200 . . . . $14.95

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Low Profile		Wire Wrap	
8 pin LP	2/.59	14 pin WW tin	.75
14 pin LP	2/.69	14 pin WW gold	1.09
16 pin LP	2/.79	16 pin WW tin	.79
18 pin LP	2/.89	16 pin WW gold	1.19
20 pin LP	2/.99	24 pin WW gold	1.69
22 pin LP	2/1.09	40 pin WW gold	2.75
24 pin LP	.79	14 p. plug/cover	1.29
28 pin LP	.82	16 p. plug/cover	1.39
36 pin LP	.99	24 p. plug/cover	1.95
40 pin LP	1.19	+ molex p./sockets	

DIODES & TRANSISTORS

IN751	2/.59	2N2219A	2/1.19
IN757	2/.59	2N2222A	2/.89
IN1188	2.69	2N2907A	2/.89
IN3600	5/.99	2N3055	.99
IN4001	4/.59	2N3772	2.25
IN4004	4/.69	2N3904	2/.69
IN4007	4/.79	2N3906	2/.69
IN4148	10/.99	2N4401	2/.79
IN4733	2/.69	2N4403	2/.79
IN4734	2/.69	2N5129	2/.69
IN4735	2/.69	2N5139	2/.69
IN4742	2/.69	2N5210	2/.79
IN4744	2/.69	2N5951	2/1.29

DESK TOP ENCLOSURES

DTE-8 (Pictured)	$31.95
DTE-11 (Pictured)	34.95
DTE-14	36.95
DTE-HK (Case for JE600)	47.95
DTE-AK (Case for JE610) (Pictured)	52.95

CMOS

4000	.69	4030	.79
4001	.69	4040	1.95
4002	.69	4044	1.39
4006	1.95	4046	2.49
4009	.89	4047	2.75
4010	.85	4049	.89
4011	.69	4050	.89
4013	.85	4051	1.95
4016	.85	4066	1.19
4017	1.49	4069	.79
4018	1.49	4070	.79
4020	2.19	4071	.79
4023	.49	4081	.69
4024	1.29	4093	1.19
4027	.89	4511	1.95

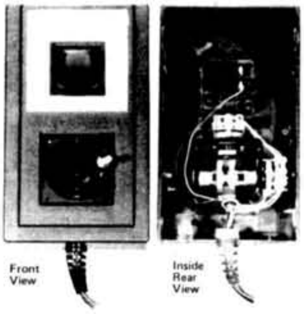
CAPACITORS

Dipped Tantalum		ELECTROLYTIC	
.1mfd@35V	2/.89	1mfd@50V	3/.69
.47mfd@35V	2/.89	4.7mfd@50V	2/.59
1mfd@35V	2/.89	10mfd@50V	2/.69
2.2mfd@25V	2/1.09	22mfd@50V	2/.79
3.3mfd@25V	2/1.19	47mfd@50V	2/.89
4.7mfd@25V	2/1.39	100mfd@50V	.59
10mfd@25V	1.19	220mfd@50V	.69
33mfd@25V	3.95	1000mfd@25V	1.19
100V MYLAR		2200mfd@16V	1.39
.001-.01mfd	4/.79	**50V CERAMIC**	
.022mfd	4/.89	10pf-.022mfd	4/.59
.047mfd	4/.99	.047mfd	4/.69
.1mfd	4/1.19	.1mfd	4/.79
.22mfd	4/1.29		

CONNECTORS

DB25P	D-Subminiature Plug	3.95
DB25S	D-Subminiature Socket	4.95
DB51226	Cover for DB25P/S	2.25
22/44SE	P.C. Edge	2.95
UG88/U	BNC Plug	2.19
UG89/U	BNC Jack	3.95
UG175/U	UHF Adapter	.59
SO239	UHF Panel Recp.	1.49
PL258	UHF Adapter	1.95
PL259	UHF Plug	1.95
UG260/U	BNC Plug	2.39
UG1094/U	BNC Bulkhead Recp.	1.49

VIDEO CONTROLLER



Ideal for all video games or remote control projects. Two mini. pots. 40K ohm each. SPST push button control. Five-wire connection cable 5' long. Rugged plastic case — 1½"H x 2-3/8"W x 4-5/16"L

JVC-40 . . . . $5.95 each

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AY-5-2376	88-Key Keyboard Encoder	13.95
2114-3	4K Static RAM (300ns)	9.49
MK4116	16K Dynamic RAM (250ns)	9.95
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2716	16K EPROM (+5V)	19.95

LINEAR

LM301N	.59	LM7805T	1.75
LM305H	1.39	LM7812T	1.75
LM307N	.75	LM7815T	1.75
LM308N	1.19	LM380N	1.49
LM309K	2.25	LM384N	2.49
LM310N	2.69	LM555N	.69
LM311N	1.49	LM556N	1.49
LM317T	2.29	LM565N	1.95
LM318N	2.95	LM566N	1.95
LM319N	2.95	LM567N	1.79
LM320K-5	2.25	LM723N	.79
LM7905T	1.75	LM741N	.65
LM7912T	1.75	LM1310N	2.95
LM7915T	1.75	LM1458N	.99
LM323K	5.95	LM1488N	1.59
LM324N	1.29	LM1489N	1.59
LM337T	2.29	LM1800N	4.49
LM339N	1.29	76477N	3.95

6

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the ham notebook

spring mounted beam saves rotor gears

What ham has not, at some time or another, had the gears torn out of his rotor drive motor when the beam has been whipped suddenly by a strong gust of wind or by a bad storm?

After experiencing this disaster several times in my sixty years in ham radio, I finally decided to do something about it. This time, when I put up my Mosley TA-33, I made sure the gears would stay in no matter what the wind velocity.

the cure

It's a simple measure and easy to accomplish (**fig. 1**). The mast from my rotor is a 2-inch piece of pipe. I slid a 6-foot (1.8 meter) piece of 1-1/2 inch pipe (this could be any other length of course) down inside the 2-inch pipe about 2 feet (0.6 meter) (this could vary). I slipped a heavy automobile shock absorber coil spring over both pipes so that the center of the spring came to the top of the 2-inch pipe. Then I welded the coil to the pipe: the top end of the coil to the 1-1/2-inch pipe; the bottom end to the 2-inch pipe. I made three weld spots around each pipe. The spring I used fit snugly around the 2-inch pipe, so welding directly to the pipe was easy.

At the top, I shimmed the spring with three pieces of 3/4-inch (2 cm) strap iron cut to about 1-inch (2.5 cm) long. This made the weld spots fit snugly to the 1-1/2-inch pipe. This precaution probably wouldn't be necessary, but it didn't take much more time and it made a neater looking weld.

Because of the lightweight construction of the TA-33 antenna, I didn't bother with an end thrust bearing at the bottom of the 1-1/2-inch pipe. The spring was heavy enough to take up the beam weight. However, with heavier and more complex beam antennas, it might be wise to do something along these lines. One simple method would be to slide a 2- or 3-inch (5 or 8 cm) cut of the 1-1/2-inch pipe inside the 2-inch pipe at the place you want the bottom end of the 1-1/2-inch pipe to rest, then drill through both pipe walls and secure the pipes with a bolt to hold the piece inside the 2-inch pipe. To avoid as much friction as possible, of course, the bottom of the 1-1/2-inch pipe and the top of the small inserted piece should be ground as flat as possible and packed with heavy machine grease.

springs

The heavier the spring the better. I came across a spring about 10 inches (25 cm) long made from 3/8 inch (9.5 mm) spring steel and 2-inches (51-cm) inside diameter. Many such springs are available in auto-part shops, usually from discarded shock absorbers. But I was lucky. I was driving past a shop one day and noticed a sign that said, HEAVY DUTY SPRINGS OF ALL KINDS. It turned out to be a spring manufacturer who

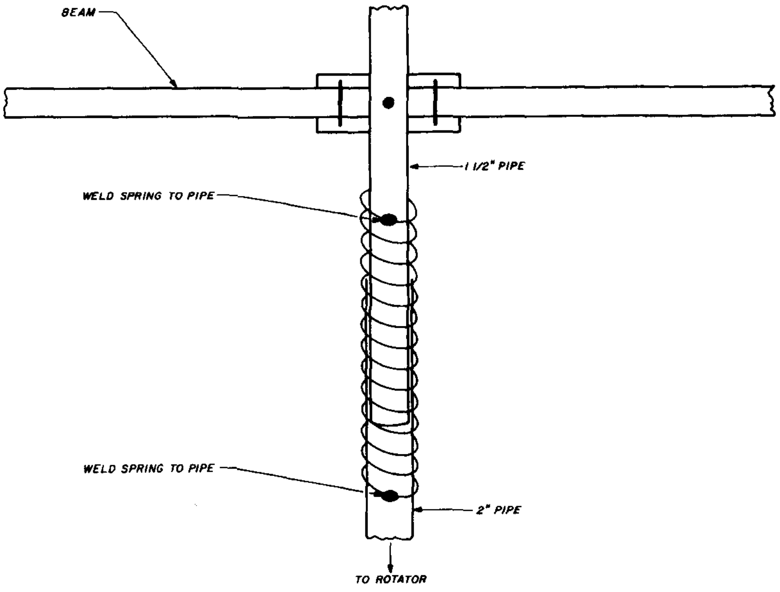

Fig. 1

made springs for the shock absorber people. I explained what I was looking for, and the shop foreman produced just what I wanted. When I asked, "How much?" he said, "Take it. It isn't worth the paperwork." Still some nice people around yet.

My beam has been up for six years. We have had all kinds of high winds, near-tornadoes, and gusts that shook the house. But the beam and the rotor gears are still intact. The beam bounces around a bit in high winds, but there is very little shock to the rotor gears. If I had it to do over, I'd try to find a heavier spring; but of course the nearer you get to a rigid connection, the less effective the arrangement becomes.

Russ Rennaker, W9CRC

calculator care

Many of the less-expensive small calculators aren't too well sealed against moisture and dirt. After living with the results of dirty contacts on the calculator keyboard of my unit, I decided to do something about it.

I opened the machine and squirted some aerosol switch-contact cleaner onto the bottom of the keyboard. I then cut and shaped a sandwich bag to fit around the calculator and taped the ends of the bag with Scotch™ tape. I poked a hole in the bag with a toothpick to accept the charger plug.

Now the calculator is protected from cigarette smoke, dirt, and grime. No more problems with contact bounce resulting in wrong entries when working long problems. The cost: about 0.5 cent.

Alf Wilson, W6NIF

varactor tuning tips

In tuning power varactor doublers, triplers, etc., there is often a sharp or sudden discontinuity in the tuning of one or more of the tuned circuits; a condition known as hysteresis.

While hysteresis is caused by some nonlinearities in the diode function, it seems that it may also be a result of the circuit Q aggravating diode nonlinearities. I figured that it might be possible to lessen the effect by a reduction in circuit Q. Accordingly, I reduced the bias resistor in my 144-to-432 MHz tripler from 92 to about 12. I was pleased to note that circuit performance was actually improved — tune up was easier, and there was no appreciable loss of power output.

Richard N. Coan, N3GN

power dissipation

Described here is a power-absorbing device commonly known as a dummy load. The circuit contains an active element so I have changed the name from dummy to active load.

an active load

The need for this circuit developed when I was trying to repair a 5-volt, 3-ampere power supply. No hot-dog-sized, 1.66-ohm resistors were available for load testing, so the circuit of **fig. 2A** was constructed and tested on the supply. Load current is controlled in both circuits (**figs. 2A** and **2B**) by R1. R2 limits the maximum base current to a safe value for the transistor used. One-hundred ohms is a nominal value. If the active load is to be used for more than a few seconds, adequate heatsinking must be provided for the transistor.

A provision for metering the current being consumed is included. I used the Simpson 260 volt ohmmeter on the 10-ampere scale.

other applications

This active load, when coupled to a properly designed heatsink, could be used in place of the Hot Mugger X1.[1] While these phenomena have not been fully investigated, an aluminum plate would probably exhibit an SWR of less than 3:1 over the operating range of the "coffee cup." Unfortunately, exact specifications for such a Hot Plate Matcher are beyond the scope of this article.

acknowledgments

I must acknowledge the contributions of David M. Newell, ex-K1KRG, who first introduced me to this circuit idea, and Donald S. Patterson, PS7ZAC, who developed the PNP version shown in **fig. 2A**.

reference

1. Burton, "The Hot Mugger X1," *73*, February, 1979, page 163.

Wm. Denison Y. Rich, PS7ZAD

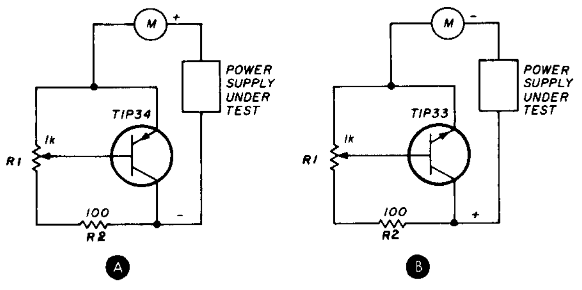

Fig. 2

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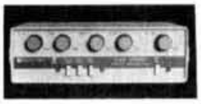

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Xitex introduces "Smart TU" for ASCII/Baudot/Morse

Xitex Corporation has just announced the addition of the UDT-170, Universal Data Transceiver, to its data-products line for RTTY and Morse operation. The UDT-170 connects directly between the user's ASCII or Baudot teletypewriter or video terminal, and the station transceiver. For the user who does not currently have an RTTY or video terminal, the Xitex SKT-100 video terminal is recommended.

The UDT-170 is the combination of a microprocessor-based data converter plus a high-performance RTTY Terminal Unit (TU). In the receive mode, the TU takes the RTTY or Morse signal from the receiver audio output and converts it to a dc signal, which is fed to the data converter portion of the UDT-170. Here, two single-chip microcomputers convert the ASCII, Baudot, or Morse input signal into an RS232 or 60-milliampere output signal, which has been regenerated to match the mode (ASCII or Baudot), Baud rate, and line length of the user's terminal.

In the transmit mode, the serial output from the keyboard on the user's terminal is fed into the data converter in the UDT-170 where it is continuously buffered and regenerated in the desired output mode

(ASCII, Baudot, or Morse) and data rate.

The UDT-170 will operate at any FSK shift from less than 100 Hz to over 1000 Hz; Baudot rates of 60, 67, 75, and 100 wpm; ASCII rates of 110 or 300 Baud; Morse rates from 1 to 150 wpm with "Auto Track"; and line lengths from 40 to 80 characters. Other features include a two-digit LED display for the copy rate (Morse only) and buffer states, and an optional CW "Ident" feature for RTTY operation.

The UDT-170 is packaged in an RFI-protected metal enclosure and operates on either 115 or 230 Vac, 50/60 Hz. For additional information contact Xitex Corporation, 9861 Chartwell Drive, Dallas, Texas 75243.

new energy-efficient voltage controls

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An all-new, rugged, aluminum housing provides a complete enclosure, and on the largest unit provides an integral carrying handle for ease of portability. All units feature fused, three-wire grounded circuitry for safety; and provide an on-off switch and pilot lamp in addition to a voltage-level adjustment knob. All controls are located on the front panel, which is recessed into the outer housing to minimize accidental readjustment. Models include the L-221 rated 1.75 A, the L-501 rated 4.5 A, and the L-1010 rated 10 A. All models are available from franchised Staco distributors throughout the country.

Applications include portable use, laboratory or bench applications, and incorporation into new or existing machines and equipment. The housing provides a means of custom mounting from either side, top, bottom, or rear of the unit, as the application requires.

Styles range from manual panel-mounted units through closed-loop voltage-regulator systems. Requests for engineering assistance may be addressed to the attention of Sales Manager, 301 Gaddis Boulevard, Dayton, Ohio 45403.

KLM multi-band vertical

KLM announces a new multiband vertical antenna. Designated 40-10V, the design uses a series of lossless linear loading and efficient High-Q air capacitor sections on 20, 15, and 10 meters, similar to those on the KT-34A and KT-34XA tribanders. Old style, power-robbing coils and capacitors have been eliminated.

In the KLM tradition, the 40-10V provides broadband coverage. All of 40 meters is accessible with no tuning adjustment at 1.5:1 VSWR or better. Optimized tuning is also possible using an adjustable element tip. Just two settings on each band provide complete coverage of 20, 15, and 10 meters at 1.5:1 VSWR or better.

The 40-10V is self-supporting; no guying is necessary. It is designed for mast, stake, or sidewall mounting. All aluminum tubing is strong, weather-resistant 6063-T832 alloy. All electrical hardware is stainless steel. Nominal feed impedance is 50 ohms. Windload is 2 square feet (0.6 square meters). Price is $109.95. For more information contact KLM Electronics, P.O. Box 816, Morgan Hill, California 95037.

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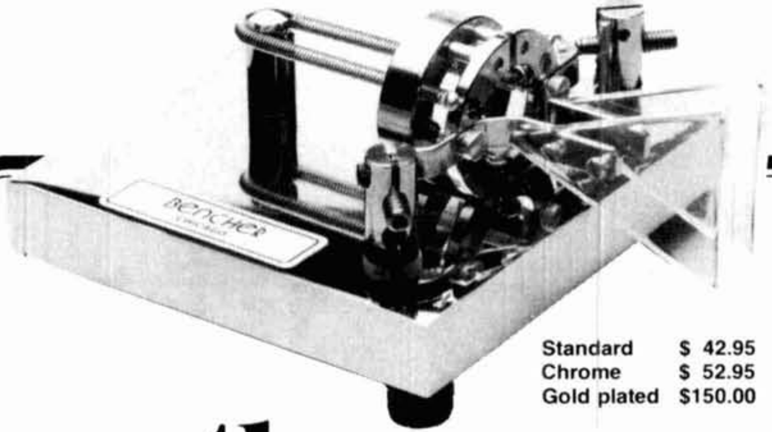

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NEW products

KLM SSV 80-40-15 antenna

The SSV 80-40-15 is the latest addition to KLM's unique new series of vertical, multi-band antennas, and, in the KLM tradition, features broadband response on 80, 40, and 15 meters. The SSV is free standing, with the lower half made up of three electrically active tripod legs. Excellent DX is possible, because the configuration of the legs contributes to a low angle of radiation on each band. Two of the legs are hinged at the base, allowing the SSV to be raised easily by two men. Only modest base preparations are needed. The upper half of the SSV is a single telescoping-whip section. It is quite flexible, and survives high winds by laying over to reduce its own wind load. Although the SSV stretches over 60 feet above ground, no guying is necessary. Overall weight is only 88 lb (39 kg). Feed impedance is 50 ohms.

A full 1/4-wave resonance is possible on 80 meters by the use of one tripod leg and the upper whip section. The adjustable tip allows the SSV to be tuned from below 3.5 MHz to 6.5 MHz, in 300-kHz steps, at 1.5:1 VSWR or better.

Resonance at 40 meters is quite broad thanks to the diameter of the base section (two of the tripod legs). Wide-range tuning is possible from 6.5 MHz and up. Performance on 40 meters appears better than a standard, ground-mounted, 1/4-wave vertical because shock excitation of the 80 meter section improves the radiation pattern.

Performance of the 3/4-wave, 15-meter section is also improved by shock excitation of the 80 meter section. The VSWR curve is very broad, with little change from band edge to band edge.

Performance approaching that of a full 1/4-wave vertical is also possible on 160 meters by simply adding inductance at the antenna base.

Experimental uses for the SSV abound. A wide-spectrum VSWR plot shows three more naturally occurring resonances that fall very close to the three new high frequency bands authorized at WARC-79 (10, 18, and 24 MHz) and are usable with slight retuning.

High-quality materials are used throughout the SSV. All aluminum tubing is drawn, seamless, 6063-T832 alloy. Tough fiberglass insulators insulate the SSV from ground and insulate the resonant sections. Base-mounting anchor-plates are supplied.

Price of the SSV 80-40-15 is $399.95. For more information, contact KLM Electronics, Inc., P.O. Box 816, Morgan Hill, California 95037.

B & W balun

Barker & Williamson, Inc., announce a new product for the Radio Amateur: the Model BC-1 Balun.

Specifications:

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Size	2¼ inch diameter; 7½ inches long (57 × 191 mm)
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For additional information contact Mr. Elmer Bush or Martin T. Zegel, Jr., at Barker & Williamson, Inc., 10 Canal Street, Bristol, Pennsylvania 19007.

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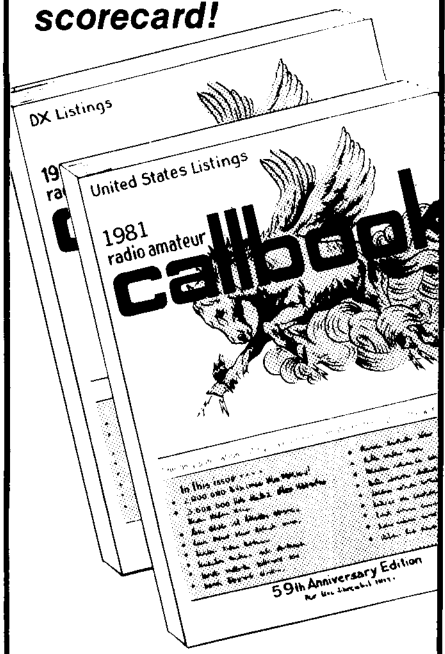

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2N2876	12.35	2N6545	12.38	MM1661	15.00
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2N2947	18.35	2N5849	21.29	MM2605	3.00
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2N3302	1.05	2N5946	14.69	MRF245	33.30
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2N3307	12.60	2N6081	10.05	MRF304	43.45
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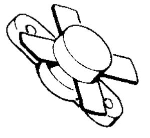

- Specified 12.5 Volt, 30 MHz Characteristics –
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  Minimum Gain = 12 dB
  Efficiency = 50%

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MRF472

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MRF475

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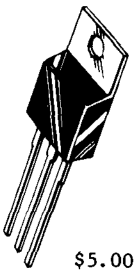

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  Minimum Efficiency = 50% (CW)
  Minimum Power Gain = 10 dB (PEP & CW)
- Common Collector Characterization

MRF458 $20.68

NPN SILICON RF POWER TRANSISTOR

. . . designed for power amplifier applications in industrial, commerical and amateur radio equipment to 30 MHz.

- Specified 12.5 Volt, 30 MHz Characteristics –
  Output Power = 80 Watts
  Minimum Gain = 12 dB
  Efficiency = 50%
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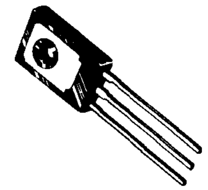

MHW710 - 2 $46.45
440 to 470MC

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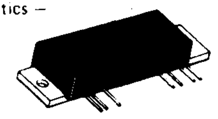

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General Microwave

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Hewlett Packard

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8436A	Bandpass Filter 8 to 12.4 GHz	75.00
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Microlab/FXR

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601-B18	X to N Adapter 8.2 - 12.4 GHz	35.00
Y610D	Coupler	75.00

Narda

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HEWLETT-PACKARD

805A Slotted Line 500MC to 4 GHz, 1.04 residual SWR. **$250.00**

809B Carriage with 806B Coaxial Slotted Section (.3 to 12 GHz), a X810B Slotted Section (8.2 to 12.4 GHz), a H810B Slotted Section (7.05 to 10 GHz), a X281A X to N adapter, a H281A H to N adapter, a HX292B H to X adapter, a 444A Probe (2.6 to 18 GHz), a PRD250 Probe (2.4 to 12.4 GHz) **$650.00**

340A Noise Figure Meter Automatically Measures and Displays IF and RF Amplifier Noise at 30 or 60 MHz. Bandwidth of 1 MHz. **$200.00**

340B Noise Figure Meter Automatically Measures and Displays IF and RF Amplifier Noise at 30 or 60 MHz. Bandwidth of 1 MHz. Input requirements – 60 to – 10 dBm. **$350.00**

AIL

74A Automatic Noise Figure Meter with a type 70 Diode Noise Generator 10 to 250 MHz, a type 71 Power Supply, a 07049 Noise Generator 3.95 to 5.85 GHz, a 07010 Noise Generator .20 to 2.6 GHz, a 0752 Noise Generator. **$650.00**

TEKTRONIX

661 90 Picosecond Rise Time Sampling Oscilloscope with a 4S1 350 Picosecond Dual Trace Sampling Plug-In DC to 1 GHz, 4S2 90 Picosecond Dual Trace Plug-In DC to 3.5 GHz, 4S3 350 Picosecond Dual Trace Plug-In DC to 1 GHz (all above Plug-Ins are 2mv/cm to 200mv/cm and with a 5T1 Plug-In Sampling System Timing. 1ns/cm to 100us/cm, (useful beyond 5 GHz). **$1000.00**

SPECTRUM ANALYZER PLUG-INS

1L5 50 Hz to 1 MHz, Center Frequency 50 Hz to 990 kHz, Dispersion – 10 Hz/cm to 100 kHz/cm, Deflection Factor 10uv/cm to 2v/cm. **$1000.00**

1L10 1 MHz to 36 MHz, Bandwidth resolution of 10 Hz to 1 kHz, Calibrated Dispersion from 10 Hz to 2 kHz, Sensitivity of – 100 dBm. **$900.00**

1L30 925 MHz to 10.5 GHz, Bandwidth resolution of 1 kHz to 100 kHz, Dispersion of 1 kHz to 10 MHz/cm, Sensitivity of – 75 dBm to – 105 dBm. **$1100.00**

1L40 1.5 GHz to 40 GHz about same specifications as above. **$1500.00**

3L10 1 MHz to 36 MHz same as 1L10 But For 560,561 Mainframe Oscilloscopes. **$1000.00**

HEWLETT-PACKARD

852A with a 8551B Spectrum Analyzer a Highly Versatile Instrument that Covers 10.1 MHz to 40 GHz. Sensitivity of up to – 100 dBm. Ten Calibrated Spectrum widths from 100 kHz to 2 GHz. Large 7 and 10cm Display.

The 852A is a Storage Display.	**$2000.00**
With The 851A Display (NOT STORAGE)	**$1500.00**
With the 851B Display (NOT STORAGE BUT NEWER)	**$1800.00**

WE ARE LOOKING FOR HEWLETT-PACKARD MODELS 8553B ECT. FOR THE 141S or T

NEW — TOLL-FREE NO. 800-528-0180 — please, orders only!

MHz electronics

R F CONNECTORS COAX

TYPE	DESCRIPTION	PRICE EACH
UG-273	Female BNC to PL-259	$3.00
UG-146/u	SO-239 to N Male	10.00
UG-83a/u	N Female to PL-259	10.00
UG-318/u	PL-259 to N Male	10.00
874	*N Female to General Radio*	15.00
UG-394b/u	BNC Male to N Female	10.00
UG-255/u	NBC Male to SO-239	5.00
UG-21e/u	N Cable Connector Male	4.00
UG-58a/u or UG-58b/u	N Female Panel	4.50
SO-239	UHF Female Panel	1.00
UG-1094a/u or UG-625b/u	BNC Female Bulkhead	1.35
UG-290a/u or UG-185/u	BNC Female	2.50
PL-259	UHF Cable Connector	1.00
UG-175 or UG-176	Adapter for RG58 or RG59 Cable for PL-259	.50
UG-88/u or UG-260/u	BNC Male 50 or 75 ohm	1.50
SO-239BM	SO-239 to PL-259 Quick Disconnect	3.00
UG-57b/u	N Male to Male	4.50
UG-27d/u	N 90° Male to Female	6.50
UG-274a/u	BNC T Male Female Male	5.00
UG-636a/u	*BNC Female to "C" Male*	10.00
UG-564/u	"C" Female to N Male	10.00
UG-635/u	BNC Male to "C" Female	10.00
UG-565a/u	N Female to 'C" Male	10.00
UG-201a/u	BNC Female to N Male	5.00
UG-306/u	BNC 90° Male to Female	3.00
M-358	UHF T Female Male Female	3.25
UG-491b/u	BNC Male to Male	5.00
UG-914/u	BNC Female to Female	3.00
PE9090	TNC Female to N Male	10.00
PE9089	TNC Male to N Female	10.00
PE9088	TNC Female to TNC Female	12.00
PE9087	TNC 90° Male to Female	20.00
PE9086	TNC Male to Male	12.00
PE9085	TNC Female to Female	20.00
PE9084, 9083, 9082	*TNC Panel and Bulkhead*	3.00
PE9081	BNC Male to F Female	5.00
PE9080	BNC Male to TNC Female	10.00
PE9079	N Female to SMA Female Panel	30.00
PE9078	BNC Female to SMA Female Panel	30.00
PE9077	"C" Female to SMC Female Bulkhead	30.00
PE9076	SMA Male for .141 semi-ridg	3.00
PE9075	SMA Male for .085 semi-ridg	3.00
PE9074	SMA Flange Female	5.00
PE9073	SMA Flange Male	5.00
PE9072	SMA Female Short	7.50
PE9071	SMA Male 50 ohm load	10.00
PE9070	SMA Female to Female	10.00
Tektronix 011-0049-01	50 ohm 2 watt term. BNC Female to Male	15.00
FXR AH-A92	0.5 dB SMA Male Female Att.	15.00
FXR AH-A93	*1.0 dB SMA Male Female Att.*	15.00
FXR AH-A94	1.5 dB SMA Male Female Att.	15.00

COAX CABLE SPECIAL SALE

Microdot RG-174
miniature 50 ohm coax cable for small jobs. This cable was made to meet military spec. (PRICE PER FOOT)
1 to 25 foot 15¢; 26 to 50 foot 12¢; 51 to 100 foot 11¢; 101 and up $10¢

Microdot RG-402U
.141 miniature 50 ohm hard line/semi-ridg coax for use with SMA/SMC etc. miniature coax connectors. This cable is very low loss and is used for High Frequency projects. (PRICE PER FOOT)
1 to 10 foot $5.00; 11 to 25 foot $4.00; *26 to 50 foot $3.00*

Microdot RG-402U with two Male SMA Connectors Assembled.
Approx. 10 to 15". $5.00

Microdot RG-402U with two Male N Connectors Assembled.
Approx. 10 to 20". $15.00

NEW — TOLL-FREE NO. 800-528-0180 — please, orders only!

MHz electronics

CRYSTALS — $4.99

KC/KHZ
15.75
24
26.25
32
49.71
70
81.9
96
100 (note)
114.1666
153.6
250
285.714
327.82
576
600
980
998.4

MC/MHZ
1.
1.024
1.05145
1.065158
1.077368
1.092105
1.125263
1.136316
1.165789
1.197368
1.3
1.3065
1.6896
1.6525
1.7
1.76375
1.77125
1.773125
1.78675
1.81875
1.845125
1.845625
1.84575
1.846
1.84825
1.84975
1.8575
1.908125
1.925
1.925125
1.927
1.932
1.982
1.985
1.9942
1.995975
1.96475
1.999659
2.
2.0285
2.05975
2.078
2.082
2.125
2.126175
2.12795
2.1315
2.133275
2.13505
2.1425
2.144625
2.14675

MC/MHZ
2.148875
2.151
2.153125
2.15375
2.15525
2.157375
2.1595
2.16375
2.165875
2.170125
2.17225
2.1765
2.17925
2.18475
2.18575
2.194125
2.198
2.207063
2.208313
2.209563
2.21812
2.210813
2.212063
2.214562
2.214563
2.215625
2.217938
2.21975
2.222125
2.22325
2.22675
2.23725
2.2395
2.24075
2.241
2.246
2.2475
2.264
2.2925
2.2975
2.3
2.32
2.326
2.32625
2.3525
2.35256
2.368
2.374
2.375
2.38725
2.394
2.395
2.396875
2.42
2.4375
2.44275
2.4495
2.45
2.482
2.486
2.5
2.51375
2.581
2.604
2.618
2.6245
2.62825
2.633125
2.63575
2.639
2.64325
2.647

MC/MHZ
2.65075
2.6545
2.65825
2.66
2.662
2.66575
2.6695
2.677
2.68075
2.681
2.6845
2.68825
2.69575
2.702
2.704
2.71075
2.715
2.716
2.723
2.73
2.7315
2.73225
2.732625
2.733
2.737
2.73975
2.742125
2.7425
2.744
2.7445
2.74475
2.746875
2.751
2.754
2.75525
2.762375
2.7735
2.776625
2.78
2.814
2.817
2.8225
2.835
2.85
2.854
2.854285
2.865
2.868
2.8725
2.876875
2.887
2.889
2.894
2.92545
2.931
2.94375
2.945
2.94675
2.952
2.966
2.97125
2.973
2.98
2.981
2.98325
2.987
3.
3.001
3.0235
3.049
3.053
3.062

MC/MHZ
3.067
3.074
3.1
3.1125
3.126
3.137
3.13975
3.1435
3.144
3.145
3.1545
3.158
3.1585
3.1615
3.1625
3.166
3.16975
3.177
3.181
3.1825
3.18475
3.1885
3.2035
3.20725
3.2165
3.2175
3.2315
3.23275
3.2365
3.23775
3.2385
3.238875
2.23925
3.24025
3.2405
3.241
3.2425
3.244
3.248875
3.24925
3.24975
3.2515
3.253625
3.255
3.256125
3.258625
3.261
3.261125
3.263625
3.266125
3.268625
3.271125
3.273625
2.33
3.4045
3.4115
3.4325
3.4535
3.4675
3.4815
3.541
3.579545
3.64
3.656
3.745
3.8
3.803
3.805
3.860
3.908
3.9168
4.

MC/MHZ
4.0457
4.096
4.1153
4.1299
4.26
4.335
4.6895
4.6965
4.7175
4.7245
4.7315
4.765
4.89
4.9037
4.93333
5.
5.13125
5.139583
5.147917
5.164583
5.1755
5.1768
5.25926
5.3037
5.33333
5.34815
5.3484
5.426636
5.436636
5.456
5.4675
5.499
5.5065
5.1111
5.5215
5.544
5.5515
5.559
5.5665
5.574
5.5815
5.58519
5.589
5.604
5.6115
5.619
5.6265
5.62963
5.6415
5.6715
5.68
5.7037
5.7105
5.733333
5.74815
5.80741
5.83704
5.85185
5.8968
5.92593
5.9525
6.
6.21
6.22222
6.25185
6.254167
6.28146
6.31111
6.321458
6.37037

MC/MHZ
6.380416
6.380833
6.381041
6.381666
6.382291
6.382916
6.383541
6.384166
6.384791
6.385416
6.42963
6.43104
6.45926
6.47
6.47111
6.48889
6.537
6.567
6.57778
6.582
6.612
6.627
6.6645
6.673
6.693
6.705
6.723
6.7305
6.738
6.75
6.75125
6.753
6.7562
6.7605
6.7712
6.77625
6.7833
6.81482
6.87407
6.9037
6.844444
6.88
6.91
6.92
6.933333
6.94
6.96296
7.01
7.125
7.225
7.25
7.255555
7.275
7.3435
7.35
7.36296
7.3728
7.39
7.42222
7.443
7.4585
7.4615
7.4685
7.4715
7.473
7.4785
7.4815
7.4985
7.62963
7.65926

NOTE 100KC is $9.99 each

NEW — TOLL-FREE NO. 800-528-0180 — please, orders only!

CRYSTALS — $4.99

MC/MHZ
7.67407
7.68889
7.71852
7.7985
7.8015
7.81
7.9
7.925
7.926667
7.95
7.975
8.
8.002
8.003333
8.0355
8.0835
8.04864
8.1
8.123
8.125
8.12625
8.14
8.15
8.15571
8.15714
8.175
8.2
8.284615
8.364
8.42308
8.5266
8.625
8.82
8.8285
8.837
8.8455
8.854
8.8625
8.871
8.8795
8.888
8.905
8.9135
8.9305
8.939
8.956
9.0265
9.327778
9.36
9.37491
9.425938
9.5075
9.545
9.555
9.565
9.585
9.643125
9.65
9.657292
9.7
9.75
9.8
9.85
9.9
9.934375
9.95
10.
10.01
10.02
10.021
10.20833
10.04
10.355
10.80375

MC/MHZ
10.8864
10.962
11.005
11.055
11.13
11.1805
11.228
11.2995
11.34
11.3565
11.50875
11.53375
11.55347
11.705
11.755
11.805
11.855
11.905
11.955
11.96125
12.925
12.93
13.102
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13.2455
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13.2945
13.3045
13.3145
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13.3345
13.3445
13.3545
13.824
14.315
15.02
15.016
15.036
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17.01018
17.015
17.065
17.115
17.165
17.215
17.28
17.9065
17.9165
17.9265
17.9365
17.9465
17.9665
17.975
17.9935
18.29
18.76563
19.006
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19.100308
19.103394
19.3483
19.3484
19.43125
19.45208
19.5385
19.6608
20.1
22.
22.22
23.25

MC/MHZ
23.575
26.375
26.62
26.64
26.66667
26.67
26.74
26.8965
26.958
26.965
27.005
27.045
27.095
27.126
27.185
27.205
27.225
27.5
27.7
27.77778
27.845
27.9
28.
28.615
28.7
28.728
28.775
28.8
28.805
28.835
28.855
28.88889
28.905
28.93888
29.896
29.9
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30.25
30.662
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31.11111
31.66667
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32.005156
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32.22222
32.6
32.936
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34.685
34.695
34.705
34.725
35.
35.02
35.03
35.04
35.07
35.08
35.11
35.12

MC/MHZ
35.14
35.18
35.19
35.2
35.3
35.36
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35.97625
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36.08
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36.3525
36.3875
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36.66667
37.
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37.77777
37.845
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38.33333
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40.51851

MC/MHZ
40.555556
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40.703704
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46.2
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49.21389
49.692
49.95
53.45
53.3
56.9
58.794
60.45
61.25
61.95
66.66667
67.52
67.82
67.94
68.1
68.12
68.18
68.375
68.48
68.60
71.015625
72.855
73.50
75.185
76.66667
82.75
83.
84.
90.833
93.1346
93.535
93.9353
94.3
102.2
106.85
115.83
121.5
126.4
128.
146.64
147.09

MHz electronics

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PHILIPS **HICKOK** **FLUKE**®

New Portable Digital Capacitance Meter

B+K PRECISION

MODEL **820**

- Measures capicitance from 0.1pF to 1 Farad
- Resolves to 0.1pF
- 10 ranges for accuracy and resolution
- 4 digit easy-to-read LED display
- 0.5% accuracy
- Special lead insertion jacks or banana jacks
- Fuse protected
- Uses either rechargeable or disposable batteries
- Overrange indication

Call For Our Price

DATA PRECISION

Model 938

0.1%, 3½-Digit, LCD DIGITAL CAPACITANCE METER

$179

- WIDE RANGING — from 199.9 pF full scale (0.1 pF resolution) up to 1999 μF full scale, in eight ranges...virtually every capacitance you'll ever need to measure.
- FAST AND EASY TO USE — Direct reading, pushbutton ranges. Just plug in and read.
- EXCEPTIONALLY ACCURATE — provides ±0.1% basic accuracy.
- TOUGH AND COMPACT — Built to take rough usage without loss of calibration accuracy. Fits and goes anywhere; takes very little bench space; always handy for quick capacitance checkout, matching, calibration, and tracking.
- PORTABLE — Palm-sized, light-weight, operates up to approximately 200 hours on a single 9V alkaline battery.
- EASY READING — big, clear, high-contrast *3½-digit LCD display, a full 0.5" high, readable* anywhere.
- VALUE PACKED — Outstanding measurement capability and dependability. Outperforms DC time-constant meters, and even bridges costing 2 to 5 times as much.
- RELIABLE — warranteed for 2 full years.

HICKOK

LX304 DIGITAL MULTIMETER

FAST, EASY, ONE HAND OPERATION

NEW

AVAILABLE NOW
$89.95

FLUKE DIGITAL MULTIMETERS

Model 8022A: The Troubleshooter

$139

- **Six functions**
  dc voltage
  ac voltage
  dc current
  ac current
  resistance
  diode test
- 3½-digit resolution
- 0.25% basic dc accuracy
- LCD display
- Overload protection

Model 8020A: The Analyst

$179

Model 8024A: The Investigator

$219

NEW

- **Seven functions**
  dc voltage
  ac voltage
  dc current
  ac current
  resistance
  diode test
  conductance (1/R)
- 3½-digit resolution
- 0.1% basic dc accuracy
- LCD display
- Overload protection
- **Free case**
- Two year parts and labor warranty

- **Nine functions**
  dc voltage
  ac voltage
  dc current
  ac current
  resistance
  diode test
  conductance (1/R)
  logic level and continuity detect
  temperature (K-type thermocouple)
- Peak hold on voltage and current functions
- Selectable audible indicator for continuity or level detection
- 3½-digit resolution
- 0.1% basic dc accuracy
- LCD display
- Overload protection

New Low Distortion Function Generator

B+K PRECISION

MODEL **3010**

Call For Our Price

- Generates sine, square and triangle waveforms
- Variable amplitude and fixed TTL square-wave outputs
- 0.1 Hz to 1MHz in six ranges
- Push button range and function selection
- Typical sine wave distortion under 0.5% from 0.1Hz to 100kHz
- Variable DC offset for engineering applications
- VCO external input for sweep-frequency tests

New Sweep/Function Generator

B+K PRECISION

MODEL **3020**

Call For Our Price

- Four instruments in one package—sweep generator, function generator, pulse generator, tone-burst generator.
- Covers 0.02Hz-2MHz
- 1000: 1 tuning range
- Low-distortion high-accuracy outputs
- Three-step attenuator plus vernier control
- Internal linear and log sweeps
- Tone-burst output is front-panel or externally programmable

More Details? CHECK—OFF Page 126

Simpson | B+K PRECISION | LEADER | DORIC

NLS Non-Linear Systems | WESTON | DATA PRECISION

PORTABLE OSCILLOSCOPES

BATTERY OPERATED

NLS Non-Linear Systems

MS-15

Single Trace 15MHz
Reg. price $349.
$299.95

MS-215

Dual Trace 15 MHz
Reg. price $465.
$399.95

MS-230

Dual Trace 30MHz
Regular price $598.
$499.95

KEITHLEY

Model 169
BENCH/PORTABLE DMM

- 3½ Digit liquid crystal display
- 0.25% basic accuracy
- 26 Ranges

$159.00

THESE 1980 B&K OSCILLOSCOPES ARE IN STOCK AND AVAILABLE FOR IMMEDIATE DELIVERY

B+K PRECISION

1479A	Dual-Trace	30 MHz
1477	Dual-Trace	15 MHz
1432	Dual-Trace	15 MHz Portable
1476	Dual-Trace	10 MHz
1466	Single-Trace	10 MHz
1405	Single-Trace	5 MHz

CALL FOR OUR EARLY BIRD SPECIAL LOW PRICE

WESTON

The Roadrunner
Model 6100

- 5 Range audible signaling function
- 0.5″ LCD display
- 6 Functions
- 29 Ranges

$139.00

80MHz Counter with Period Function

B+K PRECISION

Call For Our Price

MODEL **1820**

- 5Hz to 80MHz reading guaranteed—100MHz typical
- Period measurements from 5Hz to 1MHz.
- Period average, auto and manual positions
- One PPM resolution
- Totalizes to 999999 plus overflow
- Elapsed time measurements from .01 to 9999.99 seconds plus overflow
- One-megohm input resistance
- Bright .43″ high LED readouts

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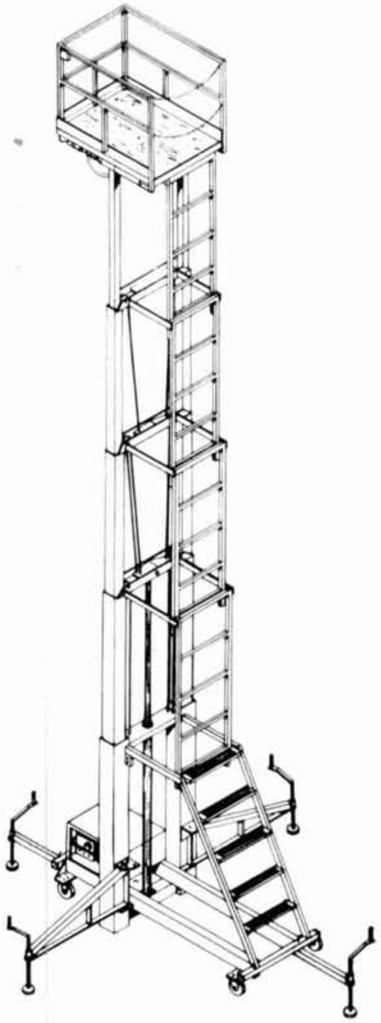

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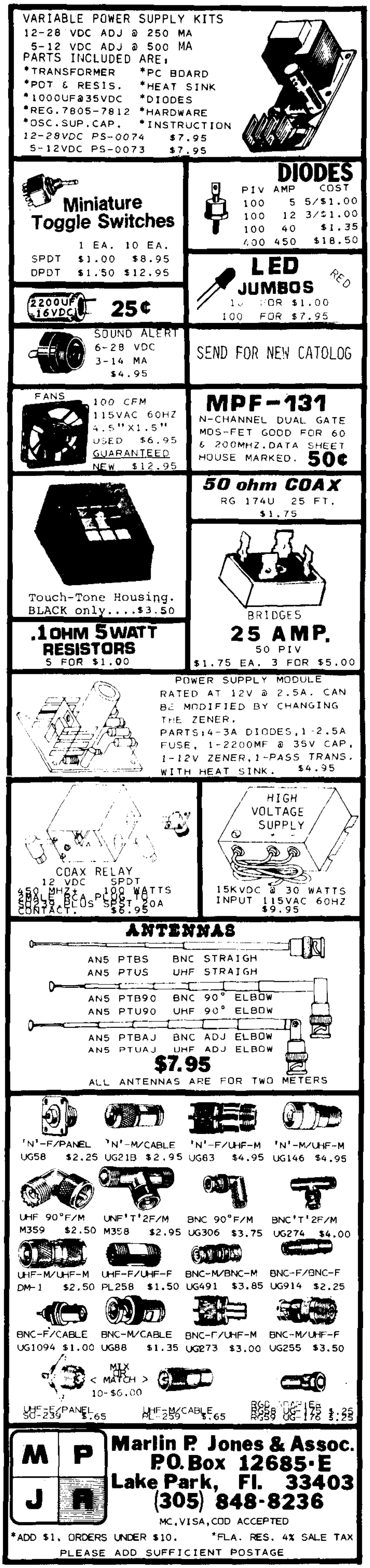

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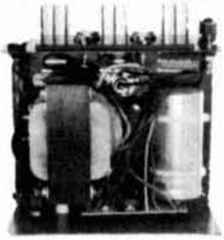

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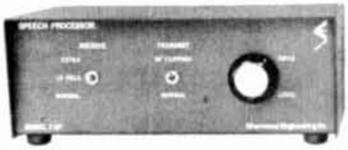

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ham radio cumulative index

1971-1980

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> To make the index easier to use only the years 1971-1980 are included, because most of the earlier material is now of limited interest. Refer to any December issue between 1970 and 1977 for a cumulative index covering 1968-1970. Copies of *ham radio* for December, 1977, may be purchased from Ham Radio's Bookstore for $2.50 postpaid.

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☐	MMB-10	mobile mounting bracket
☐	FTS-32E	CTCSS/burst encoder
☐	FTS-32ED	CTCSS encoder/decoder
☐	NC-1A	15-hr. desk charger
☐	NC-3	4-hr. quick charger
☐	NC-9B	wall charger
☐	PA-2	mobile battery eliminator/charger
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