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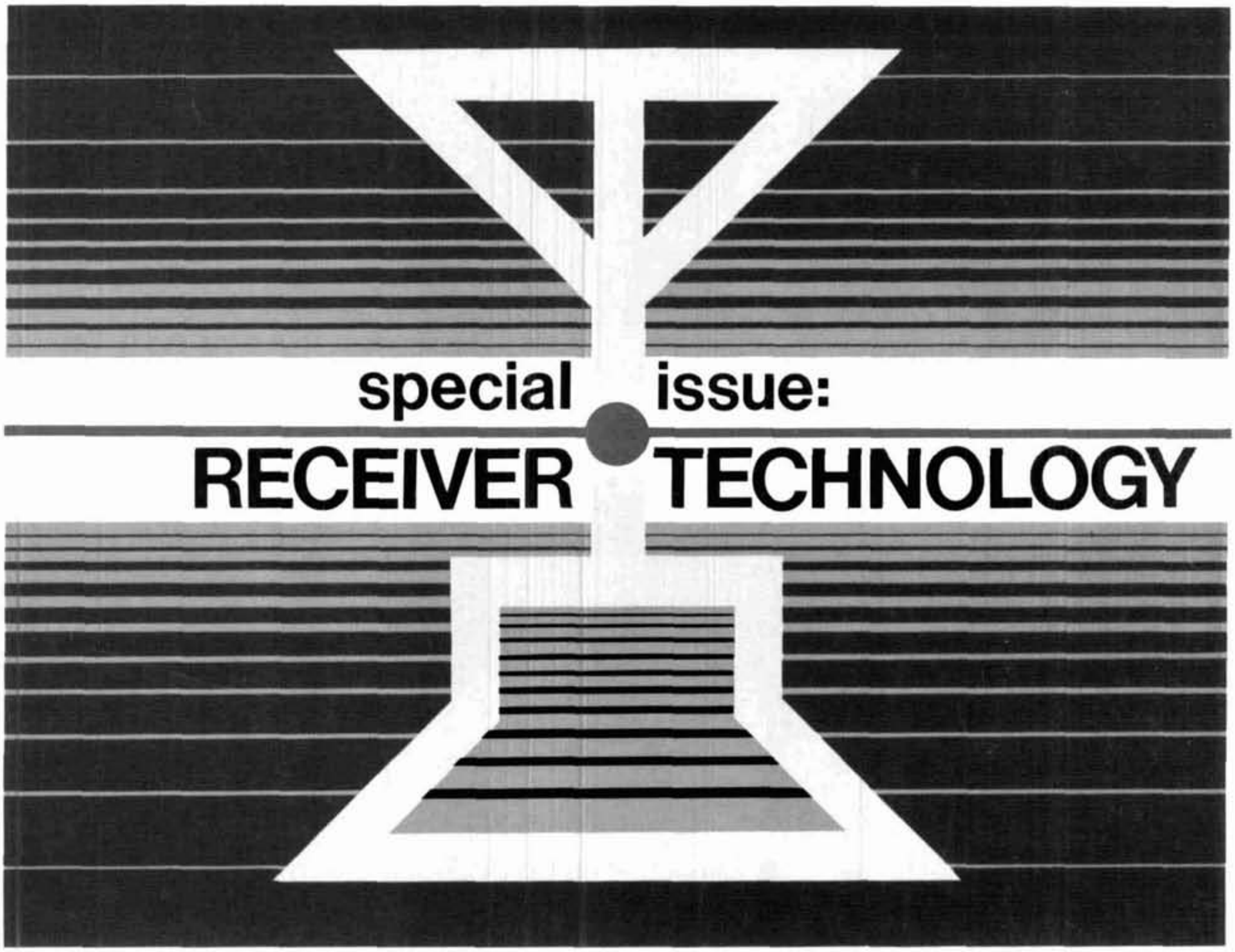
ham radio magazine



hr

focus on communications technology

understanding preamps • pulsewidth noise discriminator • predicting IMD • quieting the Russian Woodpecker • SW receiver modification • locating geostationary satellites • extended HF receiver • making PC boards • designing superhet coilsets • NEW: THE GUERRI REPORT

ICOM IC-751
The New Standard of Comparison

The IC-751 is the most advanced amateur HF transceiver available on the market today... the new standard of comparison.

Receiver. ICOM's 100KHz — 30MHz general coverage receiver with a specially designed DFM (Direct Feed Mixer) utilizes FETS in the receiver front end which gives extremely low intermodulation distortion, -19dBm intercept point, and a high dynamic range, 105dB. With cascaded filters, the IC-751 is virtually immune to overload. The FL-44 8-pole crystal in the third IF is standard and provides exceptional receiver selectivity.

Transmitter. An extremely low-noise PLL and conservative transmitter design give extremely low distortion products (-38dBm, third order) for a crystal clear transmit signal. A microphone tone control is provided to personalize the set to your particular voice. The 9 band solid-state transmitter is also a full 100% duty cycle (internal cooling fan standard) rated. For the CW operator, semi break-in or full QSK is possible.

32 Memories. An ultra versatile memory system allows storage of frequency and mode in each of the 32 memories. Data may be transferred from VFO to memory or from memory to VFO.

Standard Features. FM, FL-44A 455KHz high-grade SSB filter, SSB and FM squelch, built-in marker unit, convenient large controls, a new high-visibility fluorescent display and HM-12 Hand Mic.

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IC-PS30 System Power Supply.
The IC-PS30 25 Amp Switching Power Supply consolidates your power requirements by supplying up to four pieces of ICOM equipment, eliminating the need for independent AC power supplies for each. The IC-PS30 is designed to match all of ICOM's amateur equipment.

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The World System

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KENWOOD

...pacesetter in amateur radio

TS-930S "DX-traordinary"

TS-930S

We call it "DX-traordinary" because the TS-930S has now become the favorite rig of the serious contester! Its superior capability for full break-in split-frequency operation, the speed and convenience with which its eight memory channels can be accessed, its unsurpassed receiver dynamic range and its remarkable ability to select the desired signal during periods of heavy QRM, utilizing VBT, Slope tuning, IF Notch filtering, and tuneable audio filtering, have all combined to make this the rig that gives you the EXTRA EDGE!

The TS-930S is loaded with all the special features that you always wanted in an HF transceiver. Full coverage of the 160 through 10 meter bands, including the new WARC frequencies, (easily modified for HF MARS), plus a general coverage receiver that can tune any frequency from 150 kHz to 30 MHz. Operation in the SSB, CW, FSK, and AM modes, with selectable full or semi CW break-in. All solid-state, with 250 watts PEP input on SSB, CW, FSK, and 80 watts input on AM. SWR/power meter. Triple final protection circuits plus two cooling fans built-in. 10-Hz step synthesized frequency control. Available with optional automatic antenna tuner built-in, another industry first! Dual digital VFO's. Eight memory channels that store both frequency and band information, with internal battery back-up, (batteries not supplied). Dual mode adjustable noise blankers, especially effective in eliminating "woodpecker" type interference. SSB IF slope tuning, for maximum rejection of interference. CW variable bandwidth, with pitch and sidetone control. IF notch filter. Tuneable audio peaking filter. Unique six digit white fluorescent tube digital display is easy-on-the-eyes during those long contests. RF speech processor, for higher average "talk-power." SSB monitor circuit. 4-step RF attenuator. VOX. 100-kHz marker. AC power supply built-in, 120, 220, or 240 VAC.

TS-930S Optional Accessories:
AT-930 automatic antenna tuner, SP-930 external speaker, with selectable audio filters, YG-455C-1 (500 Hz), YG-455CN-1 (250 Hz), YK-88C-1 (500 Hz) CW filter, YK-88A-1 (6 kHz) AM filter, all plug-in type. SO-1 commercial stability TCXO, MC-60A deluxe desk microphone, MC-80 and MC-85 communications microphones, MC-42S mobile hand microphone, TL-922A linear amplifier (not for CW QSK), SM-220 station monitor, PC-1A phone patch, SW-2000 SWR/power meter, 160~6 meter, SW100A SWR/power/volt meter 160-2m HS-4, HS-5, HS-6, and HS-7 headphones.

Isn't it about time you stepped into the winner's circle?

More information on the TS-930S is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

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contents

REFLECTIONS

polish until it shines

It was a lazy afternoon. The air was warm, the sky was blue, and a soft sea breeze wafted gently across the deck. The place was Martinique, and a young radio operator from the SS *Brasil* — me — had the afternoon free. In those days my call sign was WMDT (all ships used four letters for identification), and I was at the halfway point in my fourth trip out to sea as a radio operator in the U.S. Merchant Marines.

Thinking back now, I recall the slow, undulating motion of the ship, the immense expanse of ocean, and the fresh smell of sea breeze created by the water splashing against the fantail. It was a wonderful experience for a lad of 19 to be able to visit many foreign ports, operate a high-power shipboard radio station (with four receivers), to receive room and board — and *be paid* — for the privilege!

On that lazy afternoon I decided to visit my counterparts (radio operators) aboard the SS *France* (FNRR). I suppose it was natural to want to see what equipment and antennas they had, what operating procedures they used, and in general, what their life was like aboard ship.

While the radio room on the *France* was larger than the *Brasil's*, they had about the same complement of transmitters and receivers as we had aboard our vessel, plus a high-resolution TV system used to pipe signals throughout the ship. Although our working conditions seemed similar, our feelings seemed to be quite different. The radio operators (there were about six, I believe) all appeared to be good, close friends, and they obviously enjoyed each other's company. I couldn't help but compare the atmosphere aboard my ship with that of the *France*. Though we were all friendly while on duty, we went our separate ways immediately after docking — I guess you could call our style "rugged individualism." I found myself preferring, however, the camaraderie shown by my new-found friends aboard this "foreign" liner.

What is a visit to France (or a French ship) without tasting the food? I was invited to lunch. In the cafeteria we enjoyed an excellent meal, several glasses of good wine, and amicable conversation. But suddenly my attention focused on one of the kitchen workers. I couldn't help noticing the considerable effort he was applying to the polishing of his equipment. Summoning up my best French, I went over to him and asked why he worked so hard. Were they that strict aboard the ship?

First he laughed. Then he became quite serious and said something that I'll probably never forget: "This is my job. I want to do the best I can at it. If I thought it were 'beneath me' to do this job, I'd get another."

I couldn't help thinking how many people I knew and had known who had what might be considered very good jobs, yet complained, for one reason or another, that they should have been doing something else. We have so much in this wonderful country of ours. We have resources and resourcefullness. Our children have the opportunity to acquire an excellent education, and we have the facilities to train them — and ourselves — for many different interesting jobs.

In Amateur Radio it's no different. We have the equipment, spectrum, technical resources, and obviously the time (just listen to some of our lengthy rag chews!) and yet I often come away from an evening on the air with the feeling that something's missing. We're all, it appears, "rugged individuals" diligently protective of our own frequencies and thoughts, content to do the same thing day after day. (For those who know my operating habits, perhaps I'm a fine one to talk . . . I *do* zero in on chasing quite a bit of DX.) I guess what I'm trying to say is that I'd be very happy to see what we have appreciated more and used more fully.

For my part I'm going to continue my experiments in antenna development and propagation studies, my two favorite technical subjects. But first I'm going to work on a more pressing problem — how to squeeze just two more hours into a 24-hour day. I don't think that's asking for too much.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

REALLOCATION OF THE TOP HALF OF 160 METERS TO RADIOLOCATION could take place in the very near future. In a mid-September Notice of Proposed Rule Making, the Commission has proposed moving non-government radiolocation operations from their present slot between the top end of AM broadcast and the bottom of 160 up to 1900-2000 kHz. The shift is based on the WARC '79 upward expansion of AM broadcast, displacing present radiolocation operation.

Ironically, The Importance Of Medium Frequency Radiolocation is being questioned in a Petition for Initiation of Inquiry Procedure filed by the ARRL just the day before the FCC's NPRM was released. In it the League asks that the actual spectrum requirements of the individual radiolocation users be specified along with the actual number of such stations that might be active in any geographical area. Though the ARRL petition addressed the needs of all non-government radiolocation, it specifically asked the Commission to consider whether radiolocation's real needs are sufficient to justify taking over the 1900-2000 kHz slot.

The League Has Now Petitioned The FCC To Withhold Consideration of the reallocation docket until after it considers the League's Inquiry Procedure petition.

A BILL STRENGTHENING FEDERAL LAW ON MALICIOUS INTERFERENCE has been introduced in the U.S. Senate by Barry Goldwater, K7UGA. In his bill, S-2975, Sen. Goldwater would make any operator of equipment used to maliciously interfere with any form of radio communications (or radar) subject to Section 501 of the Communications Act if he continues after receipt of written notice to stop. Section 501 provides for fines up to $10,000 and two years in prison; under present law the fine for malicious interference is only $500. In addition, the equipment used to generate the malicious interference could also be siezed.

RELIEF OF AMATEUR OPERATIONS FROM STATE AND LOCAL REGULATION is being sought by the ARRL. The League has asked the FCC to issue a "Declaratory Ruling of Limited Federal Preemption of State and Local Regulation of Amateur Radio Station Installation and Operation," to spell out just what limitations local and state authorities could place over federally-licensed Amateurs. A similar request regarding local regulation of TVRO satellite dishes was filed some time ago by United States Communications, Inc.

Comments From Concerned Amateurs, Particularly Those who've had problems with local regulators, are being sought by the Commission. An original and four copies should go to the Secretary, FCC, 1919 M St., NW, Washington, D.C. by November 9; refer to PRB-1. A copy of those Comments, along with any supporting documentation, would also be very helpful to the ARRL in its efforts. USCI's proposal on behalf of TVRO owners has generated strong opposition from a number of governmental organizations, and it's almost certain they'll resist the League's request with equal fervor.

THOUGH THERE'S BEEN NO REAL CHANGE IN THE 220 MHZ SITUATION since last month's Presstop, there have been some interesting developments. "220 Notes" Publisher K9XI has requested a Congressional investigation of the FCC's Office of Science and Technology, based on concerns that the OST may have been improperly involved in the STI petition that asked for reallocation of the 220-222 MHz slot to ACSB. "Westlink" reports Congress is getting plenty of mail on the subject, with Sen. Goldwater's office receiving about a thousand letters from concerned Amateurs and California Senator Pete Wilson almost 400.

WA2MCT's Petition To Permit Novices All-Mode 220 Privileges has been denied and dismissed by the FCC. In denying the petition Private Radio Bureau Chief Bob Foosaner noted that both the FCC and National Telecommunications Information Administration (NTIA) are conducting on-going studies of future 216-225 MHz uses, so it is "not appropriate to consider petitions which could have a major impact on the 220 MHz band..." at this time.

A SPREAD SPECTRUM FREQUENCY HOPPING 2-METER BEACON IS NOW ON THE AIR from Falls Church, Virginia. Start and stop frequencies are 144.5 and 147.7 MHz, on a 25-kHz spaced pseudo-random pattern. It's transmitting MCW on narrow band FM with a hop rate of 10 hops per second, sending a series of Vs followed by the station ID. Contact N4EZV for details.

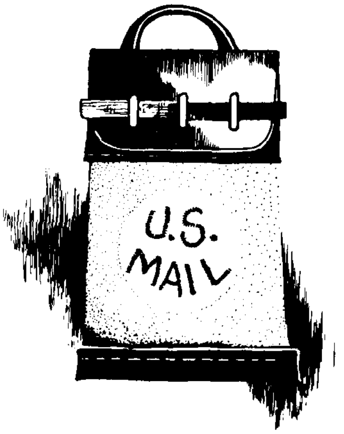
EXTENSIVE CHANGES IN THE VEC PROGRAM HAVE BEEN PROPOSED by W6NLG on behalf of the Sunnyvale VEC Amateur Radio Club, the newly appointed California VEC. They'd like the prior notification requirement relaxed, and more leniency with respect to the exams Advanced class VE's can administer. They'd also limit any VEC to a maximum of 3 call areas, to provide for local control. An RM number has not been assigned at the present time.

It Appears The FCC May Let The VEC Program Run As Is for the time being, until both it and the participants have enough experience to know what (if any) real bugs it has. However, it may act favorably on RM-4835, which would shorten the delay period for retaking a failed exam from the present 30 days to 7, despite ARRL opposition.

ARIZONA IS ADOPTING 20 KHZ SPACING ON 2 METER'S TOP END, effective immediately. No more "odd digit" coordination for either new repeaters or for changes in existing machines will be permitted, and a statewide program to move all odd digit systems will begin soon.

AN NPRM TO IMPLEMENT VARIOUS WARC BANDS IS DUE for FCC release very soon, possibly before this sees print. It's expected to include 24 and 902 MHz as well as 10 MHz (still operating under temporary authorization), and probably other WARC changes as well.

"NEW KID" THE ON THE BLOCK
Kantronics Challenger Terminal Unit
MARK SPACE POWER



comments

wait for the mailman

Dear HR:

Thanks for bringing VHF and UHF out of the dark ages and into the daylight. As I sat here carefully cutting out W1JR's article on propagation, (July, 1984) it occurred to me that it's the best primer I've ever read. The article now has a home on my research book shelf right next to authorities such as *Natural Electromagnetic Phenomena*, *Electronic Density Profiles in the Ionosphere and Exosphere*, and other noteworthy journals and papers.

WB3BGU's series on VHF and UHF Antenna Design (May-October, 1984) helps clear the smoke screen on design that has snowed hams for years. I have built and put up some large arrays over the years; Stan's notes are the best guide ever written for hams.

Since Rich Rosen took over as Editor-in-Chief, *ham radio* has moved to the number 1 position on my wait-for-the-mailman list. Keep up the great work.

Sid Liberman, WA2FXB
Woodbridge, New Jersey

Model 28 printer

Dear HR:

I have a TRS80 Color Computer,™ Kantronic Software, and an MFJ *TU-1224. I'd like to use my Model 28* as a printer. Can someone out there show me how?

John L. Gill
6000 Duda Road
House Springs, Missouri 63051

cheers

Dear HR:

Regarding your July, 1984 editorial, "The Number 1 Question," thanks! Not exactly for spelling out how to write a magazine article, but for announcing the birth of the "Superduper Louden-Boomer Metal Noodle." We in this area are using the "new and improved" version with extraordinary results and will shortly — yesterday, I believe — come out with an even more versatile one —the DASH 2 — on which I would be glad NOT to write a technical paper.

Seriously, though, I enjoy *ham radio* very much. Keep up the good work!

Frank Brumett, WB4CIZ
Lexington, Kentucky

wideband VCO design

Dear HR:

Your July, 1984, issue came just in time. I was showing my students how to use the Smith chart for finding the length of a transmission line to act as an inductor and I wanted a circuit to build. Alan Victor's article on wideband VCO design was just what I needed.

The circuit was easy to build, and because the resonator is shielded, it was immune to handling by the students. I was able to vary the frequency of the Colpitts oscillator throughout the FM radio band for the students to hear. This circuit helped my students in applying theory to a practical application.

Joe Avampato, W8DKR
Fort Mill, South Carolina

In the May, 1984, article, "Remote-controlled 40, 80, and 160-meter Vertical," reference was made to 4-inch O.D. irrigation pipe. Local inquiries produced the following information: 4-inch aluminum irrigation pipe with 0.050 inch wall is available in lengths up to 40 feet from Larchmont Engineering, P.O. Box 66, 11 Larchmont Lane, Lexington, Massachusetts 02173. The price is $2.38 per foot; other sizes are available. (Check your local phone book for additional sources.)

For additional sources of Ledex, also specified in W7LR's article, send an SASE to *ham radio*, Greenville, New Hampshire 03048.

Editor

KENWOOD

...pacesetter in amateur radio

TS-430S "Digital DX-terity!"

TS-430S

Digital DX-terity...that outstanding attribute built into every KENWOOD TS-430S that lets you QSY from band to band, frequency to frequency, and from mode to mode with the speed and ease that will give you a dominant position in DX operations.

KENWOOD'S TS-430S, a revolutionary, ultra-compact, HF transceiver has already won the hearts of radio Amateurs the world over. It covers 160-10 meters, including the new WARC bands (easily modified for HF MARS). Its high dynamic range receiver tunes from 150 kHz-30 MHz. It utilizes an innovative UP conversion PLL circuit for superior frequency stability and accuracy. Two digital VFO's allow fast split-frequency operations. A choice of USB, LSB, CW, or AM, with FM optional, are at the operators finger-tips. All Solid-state technology permits inputs of 250 watts PEP on SSB, 200 watts DC on CW, 120 watts on FM (optional), or 60 watts on AM. Final amplifier protection circuits and a cooling fan are built-in.

Eight memories store frequency, mode, and band data, with Lithium battery memory back-up. Memory scan and programmable automatic band scan help speed up operations. An IF shift circuit, a tuneable notch filter, and a Narrow-Wide switch for IF filter selection help eliminate QRM. It has a built-in speech processor. A fluorescent tube digital display makes tuning easy and fast. An all-mode squelch circuit, a noise blanker, and an RF attenuator control help clean up the signal. And there's a VOX circuit, plus semi-break-in, with side-tone. All-in-all, it just could be that the expression "Digital DX-terity" is a bit of an understatement.

TS-430S Optional Accessories:
In typical KENWOOD fashion, there are plenty of optional accessories for this great HF transceiver. There is a special power supply, the PS-430. An external speaker, the SP-430, is also available. And the MB-430 mounting bracket is available for mobile operation. The AT-250 automatic antenna tuner was designed primarily with the TS-430S in mind, and for those who prefer to "roll their own," the AT-130 antenna tuner is available. The FM-430 FM unit is available for FM operations. The YK-88C (500 Hz) or YK-88CN (270 Hz) CW filters, the YK-88SN SSB filter, and the YK-88A AM filter may be easily installed for serious DX-ing. An MC-60A deluxe desk microphone, MC-80 and MC-85 communications microphones, an MC-42S mobile hand mic., and an MC-55 8-pin mobile microphone, are available, depending on your requirements. TL-922A linear amplifier (not for CW QSK), SM-220 station monitor, PC-1A phone patch, SW-2000 SWR/power meter 160 ~ 6 meter, SW100A SWR/power/volt meter 160-2m, HS-4, HS-5, HS-6, HS-7 headphones, are also available.

More information on the TS-430S is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

Specifications and prices are subject to change without notice or obligation.

KENWOOD

... pacesetter in amateur radio

TS-711A

TS-711A Multi-function all-mode 2 m transceiver.

The TS-711A 2 m all-mode transceiver is the perfect base station unit. It features Kenwood's innovative D.C.S. circuitry that allows your TS-711A to respond only to signals that include a pre-selected digital code. The system recognizes 100,000 different 5-digit codes, making it possible for each station to have its own "private call," "group call," or "common call" code. Built-in dual digital VFO's provide commercial-grade frequency stability through the use of a TCXO (Temperature Compensated Crystal Oscillator). The new fluorescent multi-function display shows frequency, RIT shift, VFO A/B, SPLIT, ALERT, repeater offset, digital code, call sign code, and memory channel. 40 multi-function memories store frequency, mode, repeater offset and tone. It has programmable scan, memory scan, and mode scan. The Auto-mode function automatically selects the correct mode for the frequency being used. When a mode key is depressed, an audible "beeper" announces mode identification in International Morse Code.

The TS-711A has all-mode squelch, noise blanker, speech processor (SSB, FM), IF shift, RF power control, alert, and a unique channel Quick-Step tuning that varies tuning characteristics from conventional VFO feel, to stepping action when CH.Q switch is depressed.

Optional accessories:

• CD-10 Call Sign Display • TU-5 CTCSS Tone Unit • VS-1 Voice Synthesizer • MC-60A Deluxe Desk Mic • MC-80 Desk Mic • MC-85 Desk Mic • SP-430 External Speakers • MB-430 Mobile Mount • PG-2J DC Cable

TS-670

TS-670 All-mode "Quad Bander."

The TS-670 "Quad Bander" is a unique all-mode transceiver that covers the 6 meter VHF band and the 10, 15 and 40 meter HF bands. FM operation may be added with the optional FM-430. Key features include dual digital VFO's, 80 memory channels, memory scan, and programmable band scan. Direct keyboard frequency selection allows you to enter a frequency to either VFO or to a memory channel using the 10-button key-pad on the front panel. The 2-color fluorescent tube display indicates frequency to the nearest 100 Hz (10 Hz modifiable) and includes LED indicators that signal the specific functions in use. The optional GC-10 general coverage receiver unit allows continuous tuning from 500 kHz to 30 MHz. The VS-1 voice synthesizer unit is another popular option available. All this plus IF shift, all-mode squelch, CW semi-break-in with side tone, narrow-wide filter selection, noise blanker, and R.F. attenuator make the TS-670 "Quad Bander" the next transceiver you should own!

Optional accessories:

• GC-10 General Coverage Unit, 500 kHz to 30 MHz • VS-1 Voice Synthesizer • FM-430 FM Unit • YK-88C 500 Hz CW Filter • YK-88CN 270 Hz CW Filter • YK-88A 6 kHz AM Filter • PS-430 DC Power Supply • KPS-7A DC Power Supply • MC-60A Deluxe Desk Mic • MC-80 Desk Mic • MC-85 Multi-Function Desk Mic • VOX-4 VOX Unit

More information on the TS-711A and TS-670 is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, CA 90220.

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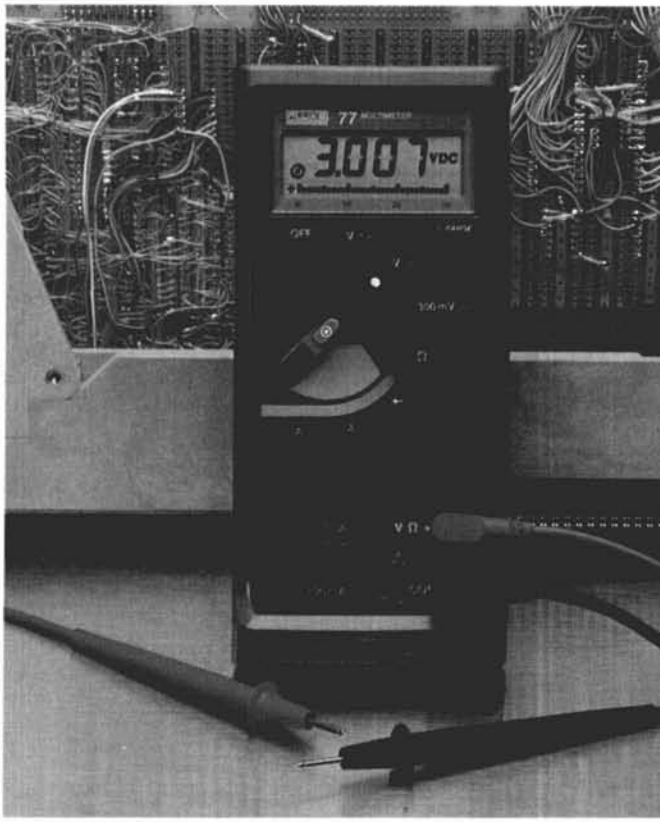

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quiet! preamp at work

Understanding preamplifiers means understanding all the important parameters of receiver performance

For years, the standard technique employed by Radio Amateurs to improve receiver sensitivity has been to precede their receivers with one or more stages of preamplification. Invariably a preamplifier that performs well on the bench will actually *degrade* the actual on-the-air system sensitivity. This article explores the relationship between gain, noise figure, bandwidth, distortion, and sensitivity in an attempt to answer the classic preamp question, "If a little is good, is a lot better?"

sensitivity

Sensitivity is a measure of the weakest input signal that will produce a specified output signal-to-noise ratio. We can quantify receiver performance in terms of *minimum discernible signal* sensitivity, which is the input level producing an output signal-to-noise ratio of unity; *tangential signal sensitivity* , which generally refers to the input level needed to produce an output signal-plus-noise to noise ratio of 6 dB or the RF level required to produce a detected signal which is 8 dB above the RMS noise level[1]; or *threshold,* which refers to the input amplitude required to produce a specified level of receiver quieting and is frequently employed in FM systems. All of these sensitivity measures are a function of the receiver circuitry's internally generated noise, bandwidth, and distortion.

Of these three parameters, the receive bandwidth can be considered fixed for a given application, and would ideally be wide enough to pass all the modulation sidebands of the desired signal, yet sufficiently narrow to exclude both background noise and any adjacent-channel signals. Because the response bandwidth of modern receivers is established primarily in the IF stages, it is relatively independent of the parameters of any preamplifier employed.

Both noise and distortion, on the other hand, are very much influenced by preamplifier performance. Most Radio Amateurs are now aware that preamplifier gain, by itself, does not necessarily assure an improvement in receiver sensitivity. Rather, to be beneficial in a system, the preamplifier must generate an internal noise level significantly lower than that generated by the receiver it precedes. The noise relationships in a cascade of stages are quantified by the now-familiar Friis Equation.[2] A well-known rule of thumb derived from the Friis Equation is that if a preamp's gain exceeds by at least 10 dB the noise figure of the receiver it precedes, the noise performance of the preamplifier will dominate the cascade.

Yet the above relationship serves merely to confuse the Amateur who measures a new preamp at a regional VHF Conference at, say, 3 dB noise figure for 15 dB gain, brings it home, installs it in front of a 10 dB noise-figure receiver, and finds its sensitivity actually degraded. What has been overlooked? Probably the effects of distortion.

distortion

A linear amplifier is one whose output signal is an exact replica of the input signal, measured in either

By H. Paul Shuch, N6TX, 14908 Sandy Lane, San Jose, California 95124

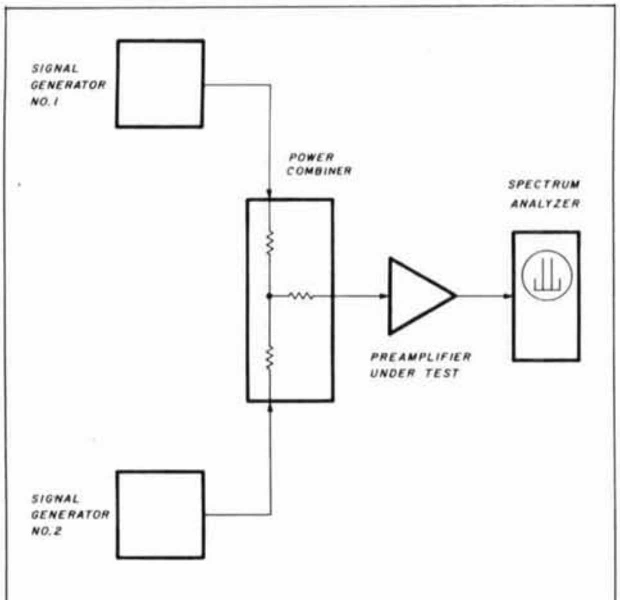

fig. 1. Two-tone intermodulation distortion test configuration.

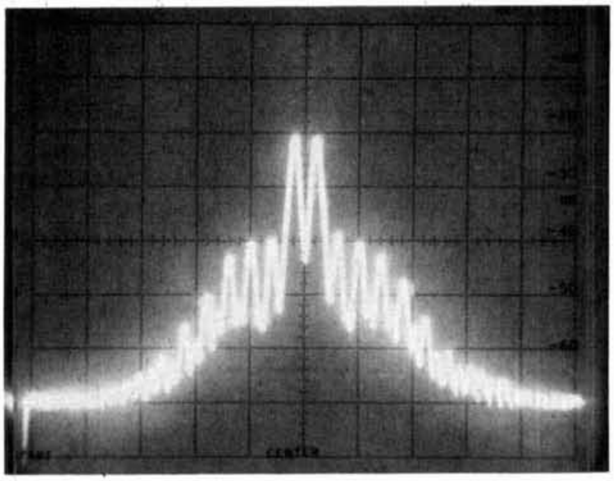

fig. 2. Typical intermodulation distortion spectrum display. Note the next pair of "signals" (IMD) are 20 dB down from the primary two-tone output.

the time or frequency domains, differing only in its increased amplitude. Try as we might, we cannot build truly linear amplifiers in the real world. Any nonlinearity introduced by an amplifier will manifest itself as a deviation from sinusoidal response when viewed in the time domain, or as the generation of new frequencies when measured in the frequency domain.

In a receive preamplifier, as in any non-linear device, the distortion products generated are integer multiples (harmonics) of the input frequency, plus their various sums and differences. Normally these distortion products would not degrade receiver sensitivity, as they would fall outside of the receiver's passband. Rare, however, is the receiver to which only a single input signal is applied. In our crowded spectra, we can anticipate countless signals of varying amplitudes within the passbands of our preamplifiers, only one of which (at a time) can be said to constitute "signal." All potentially interfering waveforms must, from a communications standpoint, be classified as noise.

It is these multiple input signals that give rise to both intermodulation (mixing of in-band signals) and cross-modulation (mixing of signals from in-band with out-of-band) distortion. When the harmonics of one signal mix with the harmonics of another, the resulting distortion products can fall within the receiver passband, degrading sensitivity.

dynamic range

Neglecting distortion effects, the weakest signal to which a receiver can respond is a function of its bandwidth and noise performance. If the multiple input signals applied to a receive system are all relatively low in amplitude, their distortion products may fall below this sensitivity limit, and be negligible. But if the input signals are of sufficient amplitude, their distortion products may appear strong enough to degrade reception of the desired signal. Thus, noise figure of a receiver generally determines the weakest signal to which it can respond. Maximum spurious free input signal, a function of a receiver's linearity, establishes an upper limit for the range of signal amplitudes to which the receiver can respond without generating perceptible distortion. The difference between sensitivity and maximum spur-free input levels is called spurious-free dynamic range, and represents a primary limitation in receiver performance.

Dynamic range is generally degraded by the addition of a preamplifier in front of a receiver. Although the low inherent circuit noise of a preamplifier may significantly improve minimum discernible signal sensitivity, degradation occurs because any additional gain in a system increases the amplitude of the desired signal, but increases the amplitude of the distortion products at an even greater rate, diminishing the maximum spurious-free input signal level. Thus, at least with respect to preamplifier gain, the old axiom, "If a little is good, a lot is better" can get us into trouble. Preamplifiers should be used only when actually necessary to improve weak-signal performance, and then only with as much gain as is actually necessary to establish the required system noise performance.

Even so, preamplifiers can result in a net degradation in system sensitivity. Some preamps are worse than others in this respect; as far as dynamic range is concerned, not all preamps are created equal. We need to measure and quantify their dynamic range, as well as their noise figure, in order to accurately predict their impact on system performance.

```
10 '---------> IMD.BAS <------------------------
20 '            Rev. A, 24 May '84
30 '            by N6TX
40 '  COPYRIGHT (C) 1984 MICROCOMM
50 '
60 '  Determines Spurious-Free Dynamic Range from
70 '  Spectrum Analyzer Two-Tone IMD Measurements
80 '
90 '------------------------------------------
100 CLR$ = CHR$(26) ' Defines Clear-Screen String
110 PRINT CLR$
120 PRINT "DO YOU WISH OUTPUT ROUTED TO:"
130 PRINT
140 PRINT "                    PRINTER (P)"
150 INPUT "              or    SCREEN  (S)";PR$
160 IF PR$="P" OR PR$="p" OR PR$="S" OR PR$="s" GOTO 200
170 PRINT CLR$
180 PRINT "YOU MUST RESPOND WITH 'P' OR 'S'" : PRINT
190 GOTO 120
200 '------------------------------------------
210 PRINT CLR$
220 PRINT "TONES -->             |   |"
230 PRINT "                      |   |"
240 PRINT "                      |   |"
250 PRINT "                      |   |"
260 PRINT " IMD  -->         |   |   |   |"
270 PRINT "                  |   |   |   |"
280 PRINT "              ----------------" : PRINT
290 PRINT "FOR TWO-TONE OUTPUT SPECTRUM AS INDICATED ABOVE,"
300 PRINT
310 INPUT "ENTER Two-Tone Output Amplitude, in dBm ",TONES
320 INPUT "ENTER Third-Order IMD Amplitude, in dBm ",IMD
330 INPUT "ENTER System Gain,               in dB  ",GAIN
340 INPUT "ENTER System Noise Figure,       in dB  ",NF
350 INPUT "ENTER System Bandwidth,          in kHz ",BW
360 I3=TONES + ((TONES - IMD) / 2)
370 MDS = NF - 144 + 10*(LOG(BW)/LOG(10))
380 INMAX = ((2/3)*(I3-GAIN))+(MDS/3)
390 RANGE = INMAX - MDS
400 '------------------------------------------
410 PRINT CLR$
420 PRINT "INTERMODULATION ANALYSIS BY MICROCOMM"
430 PRINT : PRINT
440 PRINT USING "####.# dBm -->          |   |    ";TONES
450 PRINT       "                        |   |    "
460 PRINT       "                        |   |    "
470 PRINT       "                        |   |    "
480 PRINT USING "####.# dBm -->      |   |   |   |";IMD
490 PRINT       "                    |   |   |   |"
500 PRINT       "                ----------------"
510 PRINT : PRINT
520 PRINT USING "SYSTEM GAIN          = ####.# dB";GAIN
530 PRINT USING "SYSTEM NOISE FIGURE  = ####.# dB";NF
540 PRINT USING "SYSTEM BANDWIDTH     = ####.# kHz";BW
550 PRINT
560 PRINT USING "OUTPUT THIRD ORDER INTERCEPT POINT = ####.# dBm";I3
570 PRINT USING "MINIMUM DISCERNIBLE INPUT SIGNAL   = ####.# dBm";MDS
580 PRINT USING "MAXIMUM SPURIOUS-FREE INPUT SIGNAL = ####.# dBm";INMAX
590 PRINT USING "SPURIOUS-FREE DYNAMIC RANGE        = ####.# dB";RANGE
600 '------------------------------------------
610 IF PR$="S" OR PR$="s" THEN GOTO 800
620 LPRINT "INTERMODULATION ANALYSIS BY MICROCOMM"
630 LPRINT : LPRINT
640 LPRINT USING "####.# dBm -->          |   |    "; TONES
650 LPRINT       "                        |   |    "
660 LPRINT       "                        |   |    "
670 LPRINT       "                        |   |    "
680 LPRINT USING "####.# dBm -->      |   |   |   |"; IMD
690 LPRINT       "                    |   |   |   |"
700 LPRINT       "                ----------------"
710 LPRINT : LPRINT
720 LPRINT USING "SYSTEM GAIN          = ####.# dB";GAIN
730 LPRINT USING "SYSTEM NOISE FIGURE  = ####.# dB";NF
740 LPRINT USING "SYSTEM BANDWIDTH     = ####.# kHz";BW
750 LPRINT
760 LPRINT USING "     OUTPUT THIRD ORDER INTERCEPT POINT = ####.# dBm"; I3
770 LPRINT USING "     MINIMUM DISCERNIBLE INPUT SIGNAL   = ####.# dBm";MDS
780 LPRINT USING "     MAXIMUM SPURIOUS-FREE INPUT SIGNAL = ####.# dBm";INMAX
790 LPRINT USING "     SPURIOUS-FREE DYNAMIC RANGE        = ####.# dB";RANGE
792 LPRINT
794 LPRINT
796 LPRINT
798 LPRINT
800 '------------------------------------------
810 PRINT : PRINT
820 INPUT "TYPE <return> TO CONTINUE, 'Q' TO QUIT ",D$
830 IF D$ = "Q" OR D$ = "q" THEN GOTO 850
840 GOTO 100
850 '
860 '-------------------------------------------------------------------
870 '
880 '         EQUATIONS EXECUTED
890 '
900 'MINIMUM DISCERNIBLE SIGNAL  =  -174 dBm/Hz  + NF (dB)  + 10 * LOG BW (Hz)
910 '
920 'OUTPUT INTERCEPT POINT      =  P (tones)  +  [ P (tones) - P (imd) ] /  2
930 '
940 'MAXIMUM INPUT SIGNAL LEVEL  =  (2/3) * (INTERCEPT - GAIN) + (M. D. S. /3)
950 '
960 'SPURIOUS FREE DYNAMIC RANGE =  MAXIMUM INPUT - MINIMUM DISCERNIBLE SIGNAL
970 '
980 '
990 '-------------------------------------------------------------------
1000 END
```

fig. 3. CP/M BASIC language program listing to determine spurious-free dynamic range from spectrum analyzer two-tone IMD measurements.

gain compression

Inferences about an amplifier's dynamic range can be drawn by applying to its input a single signal of varying amplitude and observing the amplitude present at the output. In its linear region, the amplifier will produce a 1-dB change in output signal amplitude for every 1-dB change in the applied signal. That is, the gain of the amplifier is independent of applied signal level. But as the upper limit of dynamic range is approached, output signal changes will be unable to keep pace with the input. That is, the gain of the amplifier compresses at the upper end of its dynamic range. The output level at which the amplifier is exhibiting 1 dB less gain that it was under weak-signal conditions is referred to as its *output 1-dB compression point,* and is an indicator of the amplifier's immunity to intermodulation and cross-modulation distortion.

For a given noise figure, the preamplifier with the highest compression point will offer the greatest spurious-free dynamic range. But correlating the two parameters directly is difficult because the relationship between compression and distortion varies between active devices, and between circuit configurations.

Another indicator of dynamic range relates to the fact that if you continue to increase the drive level to an amplifier beyond the compression point, the gain further decreases. Eventually, the amplification of the desired signal is degraded to a point at which its amplitude at the output of the amplifier, and those of the intermodulation distortion products, would be the same. The output level at which this should occur is called the *output intercept point.** Intercept point is more readily correlated to dynamic range than is compression point, but is difficult to measure directly. To best quantify dynamic range limitations, it is necessary to test the preamplifier in its actual operating environment — that is, under multiple-signal conditions.

```
INTERMODULATION ANALYSIS BY MICROCOMM

  -6.0 dBm -->          |   |
                        |   |
                        |   |
                        |   |
 -26.0 dBm -->      |   |   |   |
                    |   |   |   |
                ----------------

SYSTEM GAIN          =   -6.0 dB
SYSTEM NOISE FIGURE  =    7.0 dB
SYSTEM BANDWIDTH     =    2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    4.0 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   = -133.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =  -37.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =   95.5 dB
```

fig. 4. IMD analysis of a double-balanced mixer with a +7 dBm injected LO level.

two-tone testing

In the method of dynamic range testing prevalent in industry, two sinusoidal signals of equal amplitude are applied to the input of the device under test, and the resulting output spectrum monitored in the frequency domain. The two input signals, or tones, may be generated by summing the outputs of the two signal generators in a power combiner, or by applying a single RF source to the LO input of a balanced mix-

*For many amplifier circuits, this is a *theoretical,* rather than an *attainable,* level, because the active device may burn out before this output level is reached.

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Calibrated Output DCV/ DMM Range Setting DCV	190 mV/ 200 mV	1.90V/ 2V	19.9V/ 20V	199V/ 200V	990V/ 1000V
Unit #1					
Actual Reading	190.6	1.904	19.94	199.2	993
% Accuracy	0.316	0.211	0.201	0.101	0.303
Unit #2					
Actual Reading	190.5	1.905	19.97	199.1	992
% Accuracy	0.263	0.263	0.352	0.050	0.202
Unit #3					
Actual Reading	190.3	1.901	19.92	199.3	992
% Accuracy	0.158	0.053	0.101	0.151	0.202
Unit #4					
Actual Reading	190.4	1.902	19.92	199.4	993
% Accuracy	0.211	0.105	0.101	0.201	0.303
Unit #5					
Actual Reading	190.4	1.901	19.89	198.7	989
% Accuracy	0.211	0.053	0.050	0.151	0.101

And we guarantee accuracy for 1 full year.

AWARDED FOR KEEPING UP WITH OUR ADVANCING TECHNOLOGY

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```
INTERMODULATION ANALYSIS BY MICROCOMM

 -7.0 dBm -->          | |
                       | |
                       | |
                       | |
 -65.0 dBm -->       | | | |
                     | | | |
                   ----------------

SYSTEM GAIN           =    -7.0 dB
SYSTEM NOISE FIGURE   =     8.0 dB
SYSTEM BANDWIDTH      =     2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    22.0 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   =  -132.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -24.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =   107.5 dB
```

fig. 5. IMD analysis of a double-balanced mixer with a +17 dBm injected LO level.

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -14.0 dBm -->         | |
                       | |
                       | |
                       | |
 -25.0 dBm -->       | | | |
                     | | | |
                   ----------------

SYSTEM GAIN           =    12.0 dB
SYSTEM NOISE FIGURE   =     2.0 dB
SYSTEM BANDWIDTH      =     2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    -8.5 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   =  -138.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =   -59.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =    78.5 dB
```

fig. 6. IMD analysis of a bipolar junction transistor preamplifier.

er, a suitable audio signal generator to the mixer's IF input, and applying to the device under test the double-sideband (two-tone) signal appearing at the mixer's RF port. In either case, the two tones must be separated in frequency sufficiently to be individually resolved on the spectrum analyzer's display, yet sufficiently close in frequency to both fall within the response bandwidth of the device under test.

A typical interconnection of instruments for two-tone dynamic range analysis is shown in **fig. 1**, and a typical resulting spectrum is displayed in **fig. 2.** Note that the distortion products of greatest amplitude (in this case, the pair of signals immediately adjacent to the two applied tones) are roughly two divisions, or 20 dB, below the amplitude of the desired output tones. The intermodulation distortion level of this particular amplifier, measured at this particular signal level, is thus -20 dB.

If the vertical axis of the spectrum analyzer is calibrated in absolute amplitude (typically in dBm), the output power per tone, the PEP output power (6 dB above the level of each individual tone), and power of the individual distortion products can be readily determined. And from these values, with minimal number crunching, we can determine the dynamic range of the preamplifier.

data analysis

The mathematical relationships applied next are, as is said in college texts, "beyond the scope of this course." However, I have included in **fig. 3** a listing of a Micro-soft™ BASIC program that performs the complete analysis. Although written to run under the CP/M™ operating system, the program can likely be modified to run on any of the popular home computers using their version of BASIC. **Figures 4** through **8** are sample executions of the IMD program for various receiver configurations. Comparing these printouts will enable us to draw some significant conclusions with regard to the utility of preamplifiers in VHF and UHF communications systems.

mixer design considerations

As a rule, balanced mixers offer excellent dynamic range and intermodulation distortion performance, although their weak-signal sensitivity leaves something to be desired. Mixers are designed to operate at different levels of local oscillator injection, and generally, the higher the LO level employed, the higher will be the mixer's compression level. However, raising the LO injection above perhaps 5 milliwatts tends to degrade mixer conversion efficiency and noise figure. Nonetheless, as **figs. 3** and **4** indicate, so-called high level mixers offer sufficiently improved dynamic range to override the considerations of slightly degraded sensitivity, in most applications.

Not shown in the computer runs, but worthy of consideration, are the so-called "starved LO" mixers. These devices use an extremely low LO injection level with external DC bias of their mixer diodes, and excel in low-noise performance. Their dynamic range, however, is severely degraded, typically 12 to 15 dB below that of even the "low-level" balanced mixer shown in **fig. 3**. Thus, except in those applications in which it is impractical to generate 5 milliwatts or more of LO injection, starved LO operation should be avoided.

The same is true for harmonic mixers. These devices are extremely popular in microwave TV receive converters, and employ LO injection at half the normal frequency, with the mixer diodes serving double duty as frequency multipliers. Obviously, the more frequencies we generate within a mixer, the more spurs will be available to bite us later. I recommend multiplying in a stage separate from that doing the heterodyne conversion.

preamp design considerations

Most receive preamplifiers operate with their active devices drawing relatively low quiescent current. This is done because high device current generates high thermal activity, which degrades noise performance significantly. Unfortunately, biasing any active device

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -9.0 dBm -->          | |
                       | |
                       | |
                       | |
 -22.0 dBm -->       | | | |
                     | | | |
                 ----------------

SYSTEM GAIN           =   17.0 dB
SYSTEM NOISE FIGURE   =    3.0 dB
SYSTEM BANDWIDTH      =    2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =    -2.5 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   = -137.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =  -58.7 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =   78.5 dB
```

fig. 7. IMD analysis of a MOSFET preamplifier.

```
INTERMODULATION ANALYSIS BY MICROCOMM

 -2.0 dBm -->          | |
                       | |
                       | |
                       | |
 -28.0 dBm -->       | | | |
                     | | | |
                 ----------------

SYSTEM GAIN           =   24.0 dB
SYSTEM NOISE FIGURE   =    1.0 dB
SYSTEM BANDWIDTH      =    2.4 kHz

     OUTPUT THIRD ORDER INTERCEPT POINT =   11.0 dBm
     MINIMUM DISCERNIBLE INPUT SIGNAL   = -139.2 dBm
     MAXIMUM SPURIOUS-FREE INPUT SIGNAL =  -55.1 dBm
     SPURIOUS-FREE DYNAMIC RANGE        =   84.1 dB
```

fig. 8. IMD analysis of a GaAs FET preamplifier.

N6TX TABLE 1
RECEIVER DYNAMIC RANGE COMPARISON

	STANDARD LEVEL DBM	HIGH LEVEL DBM	BIPOLAR JUNCTION TRANS	MOS FIELD EFFECT	GaAs FIELD EFFECT
Conversion Gain (dB)	-6	-7	12	17	24
Noise Figure (dB)	7	8	2	3	1
IMD Intercept (dBm)	+4	+22	-8	-2	+11
SSB Sensitivity (dBm)	-133	-132	-138	-137	-139
Maximum Input (dBm)	-38	-25	-60	-59	-55
Dynamic Range (dB)	95	107	78	78	84

NOTE: Measurements were performed at 144 MHz, with representative devices. Results at other frequencies will vary, but comparisons will be similar.

near cutoff tends to limit its dynamic range, such that the "optimum" bias point from a noise figure standpoint often coincides with the "worst" bias point as far as dynamic range and actual system sensitivity are concerned. Remember, although we talk about desiring high "signal to noise ratio," what we really need for maximum sensitivity is a signal level that is high relative to the sum of noise and distortion. If we can considerably reduce IMD interference by giving up some slight amount of noise performance, the overall system sensitivity has to improve!

Joe Reisert, W1JR — probably the most prominent UHF DXer of our time — has long advocated designing bias circuits for preamplifiers so that device quiescent current can be readily and remotely varied.[3] This way the user can optimize noise figure when operating conditions call for it, and readily improve dynamic range, at a sacrifice in noise performance, should interference conditions dictate. Since all RF design is a series of compromises, Joe's approach seems to offer the best of all possible worlds.

There has long been controversy in Amateur circles over the relative merits of bipolar junction transistors and MOS field effect devices as VHF preamplifiers. Bipolar advocates boast the excellent low-noise performance of these devices, while those preferring the MOS devices cite their higher gain and stable operation, which eliminates the need for neutralization. **Figures 6** and **7** seem to indicate that neither device holds a clear advantage as far as overall system performance is concerned. The two representative amplifiers I tested in preparing this manuscript exhibited identical dynamic range.

Gallium-Arsenide Field Effect Transistors, on the other hand, are the undisputed winner in all areas of VHF and UHF performance. As indicated in **fig. 8**, the GaAs FET offers exceptional high gain, low noise, and wide dynamic range performance. If only they weren't so expensive!

summary

In evaluating receiver performance, it is necessary to consider dynamic range limitations, as well as noise figure, to select the combination of devices and circuits that will yield the best overall sensitivity. **Table 1** summarizes the results of testing various competing mixer and preamplifier technologies. Although the tests were performed at 2 meters, we can generalize the results to other VHF and UHF bands as well.

It appears that best receiver performance will be achieved by cascading a GaAs FET preamplifier with a high-level doubly-balanced mixer. Two-tone analysis confirms that such a combination has considerable immunity to intermodulation and cross-modulation interference, while maintaining an impressively low-noise figure.

references

1. H.W. Bode, *Network Analysis and Feedback Amplifier Design*, D. Van Nostrand Co., Inc., New York, New York, 1945.
2. H.T. Friis, *"Noise Figures of Radio Receivers," Proceedings of the IRE*, July, 1944, page 419.
3. Joe Reisert, W1JR, "Ultra Low-Noise 432 MHz Preamplifier," *ham radio*, March, 1975, page 8.

This article was adapted from a paper originally presented at the 18th Conference of the Central States VHF Society, held in Cedar Rapids, Iowa, on 28 July, 1984, and appeared in the Proceedings of that conference.

ham radio

a pulsewidth noise discriminator

Impulse noise control works on time duration rather than amplitude

I think most hams would agree that the best impulse-noise squelch would be one that stopped each noise burst at its source. Unfortunately there are too many noise bursts arriving from too many directions to make such a thing possible. It's usually difficult to locate a local source; even if you do, the person responsible for the noise is often unwilling or unable to cooperate. Once launched, these disturbances seem determined to enter our receiving systems; when they do, they're repeatedly amplified, modified, and stretched as they race from antenna terminals to speaker. In short, once admitted, these unfriendly signals are actually made worse — often *much* worse — by your receiver's own circuits.

What happens at the output after a burst of noise arrives at your receiver's input is quite predictable. Some of the fast-changing wave front is absorbed by the first tuned circuit, then released in the form of ringing at this filter's natural resonant frequency. The resulting damped oscillation is then translated by a local oscillator, resulting in a rapid rise at the input of the IF filter, which absorbs some of the energy and then releases it in the form of ringing at *its* resonant frequency . . . and so on.

basic noise control methods

Over the years a great many circuits have been tried in an effort to control impulse noise. Successful methods have been of two basic types: the noise blanker and the noise limiter. The well known Lamb filter (often called a hole-puncher, noise silencer, or noise blanker) takes a sample of each noise pulse from a receiver stage as near to the antenna as possible and, using fast circuits, forms a blanking pulse that momentarily blocks the receiver's IF stage just before the ringing pulse of noise energy arrives. The blanking pulse is designed to embrace the ringing time caused by the filter characteristic. Some rise and fall time is usually added so the blanking function will not itself generate audible clicks at the receiver's output. This system has been around for a long time; properly designed, it works very well. But one problem with this technique is that it must be designed to go into action only on noise pulses that are significantly larger than desired signals in order to avoid the creation of excessive distortion.

Another form of impulse noise control is called the "peak limiter" system. Again, this method is restricted to noise signals whose peak amplitudes are above that of desired signals. When a noise burst is received, the desired signal is momentarily suppressed and the interference is limited in peak amplitude. Perhaps the best features of this noise control system are its simplicity and its ability to reduce possible damage to our ears caused by otherwise nearly unlimited sharp audio sound transients.

Neither method is effective in removing noise bursts of low to moderate amplitude, or of durations of greater than a few microseconds, which usually includes those "woodpecker" style noise disturbances. The majority of disturbances fall in the latter category, with high amplitude disturbances in the minority.

pulsewidth noise discriminator

This article is about a third method that effectively handles a wide range of impulse noise amplitude levels and can be used either by itself or in conjunction with the more familiar methods described above. Furthermore it can be added at the *audio* output of any receiver. I have called this method the PND — Pulsewidth Noise Discriminator. Rather than working on peak amplitudes, this system makes use of time duration differences between the character of almost all desired signals and impulse noise. Impulse noise bursts at their origin exist for only a few nanoseconds to

By Don E. Hildreth, W6NRW, P.O. Box 60003, Sunnyvale, California 94088

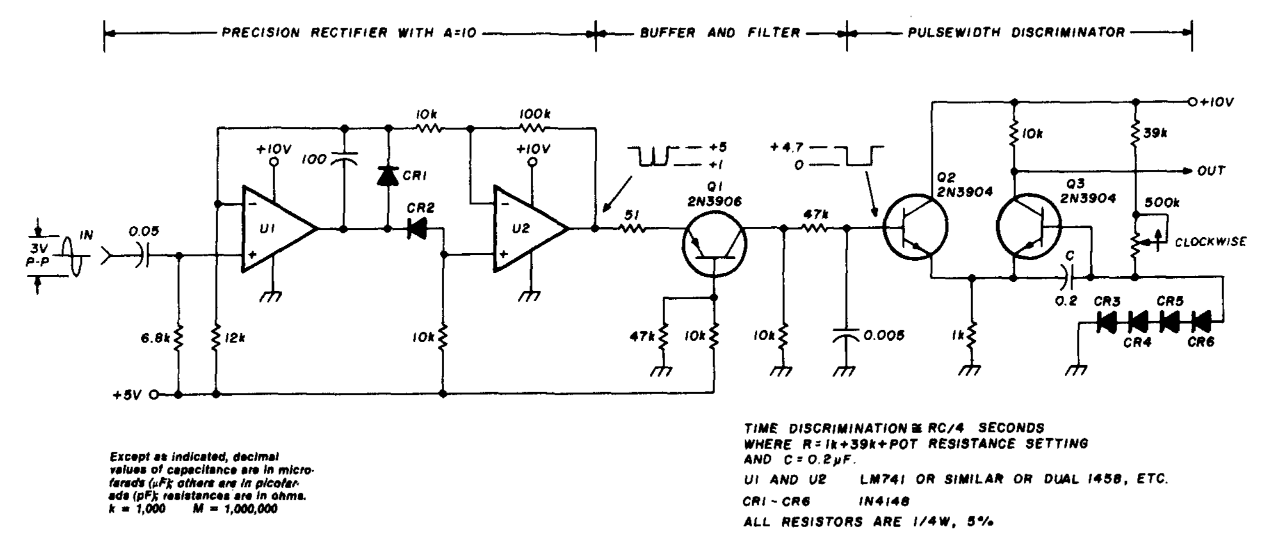


fig. 1. Basic PND system.

microseconds, but they are then transformed by receiver circuits into ringing bursts lasting from a little less than one millisecond up to as much as ten milliseconds, depending on the shape and bandwidth of the narrowest filter being used.[1] Concurrently, the vowel parts of desired SSB voice signals are fractions of a second long; even the shortest parts of CW are typically of 50 to 100 milliseconds duration for the "dit" and "space" lengths.

Using this data, a basic pulse width discriminator is designed to ignore any signal until it has existed for a selectable period of time — 10 milliseconds, for example — and will consequently block a noise pulse stretched by a typical 100 Hz bandwidth filter.

The circuit of **fig. 1** is the basis of this sytem. When no signal is present, Q1 and Q2 are ON and Q3 is OFF. When a signal arrives, the precision rectifier formed by U1 and U2 develops a negative gate that turns off Q1 and Q2. However, Q3 remains off until the no-signal reverse charge of a nominal -2 volts on C, Q3's base to emitter capacitor, is bled off and reversed to the required level of approximately +0.7 volts by charging through R, which is made up of 1K + 39K + the 500K pot setting. When Q3 goes on, its collector voltage drops. It is this signal that is used to accept a desired signal or to reject those that are too short for completion of the timing cycle as determined by the effective value of R. A second and very desirable feature of the basic pulsewidth discriminator part of the circuit is that it resets very quickly, thereby avoiding the integration of noise pulses provided they are not too close together (equal to or less than the selected time discrimination period). This feature makes the full-wave rectification provided by U1 and U2 plus a small amount of filtering necessary to avoid a functional dropout between the cycles of a desired signal. Using the specified circuit values, a minimum of 1 volt RMS is required for normal operation.

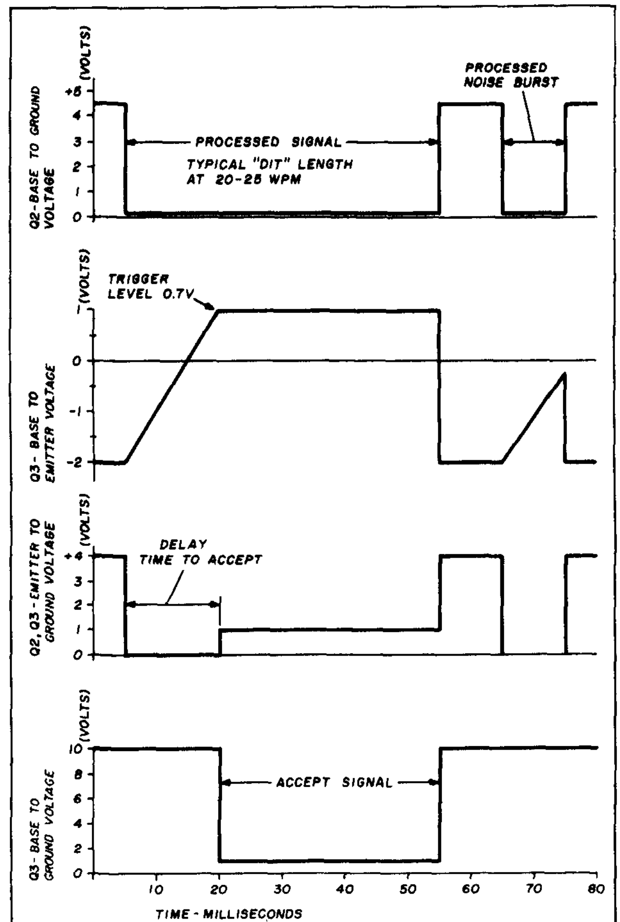


fig. 2. Idealized PND waveform.

Fig. 2 shows what happens as a function of time. Since the pulsewidth discriminator circuit would ignore a ringing or sinusoidal signal of any period, the incoming signal is transformed into a negative gate. An ideal gate pulse is shown for clarity. This use of an oscilloscope is convenient at the Q2 and Q3 emitter junction, to observe and measure the time discrimination performance.

a PND application

Fig. 3 shows an application in which an audio filter feeds the pulse noise discriminator and the output of the discriminator is used to key an audio white-noise generator for listening to CW. (Of course you can have it key a tone oscillator if your prefer.) The mixer control is arranged to enable listening to CW directly through the filter, or to the noise generator, or to any mixture of the two. By connecting the control in this experimental way, you can test and experience how well this idea works: simply compare a non-noise discriminated 750 Hz CW output to the processed audio-noise CW output that is driven by the PND.

In operation, the + 10 volt level at Q3's collector is used to squelch noise output from the dual op-amp noise generator shown in **fig. 3** or in the SSB output of **fig. 4**. When the PND system recognizes a signal, the Q3 collector level drops, opening the transistor switch Q2 of **fig. 3** or Q1 of **fig. 4**. One undesirable feature of this noise control system relates to operating convenience: your receiver's output is normally OFF until a qualified signal appears. If you like to be aware of the noise floor, as I do, this can be a disadvantage, so I usually run the mixer control midway when looking for DX. Then, depending on conditions, I decide which way to twist this control. If I want to avoid the tinkling roar present when listening to a low-level signal through a narrow CW filter, I turn the control to admit only the keyed-noise signal with its silence between characters. But if I want to make use of the ear-brain filter capabilities (when there is more than one signal in the filter passband) then I turn the control to allow only the signal through the 750 Hz filter[2] to reach the power amplifier.

With the mixer control set toward keyed-noise operation, and with the delay control set at minimum, slowly increase the receiver gain at a no-signal spot

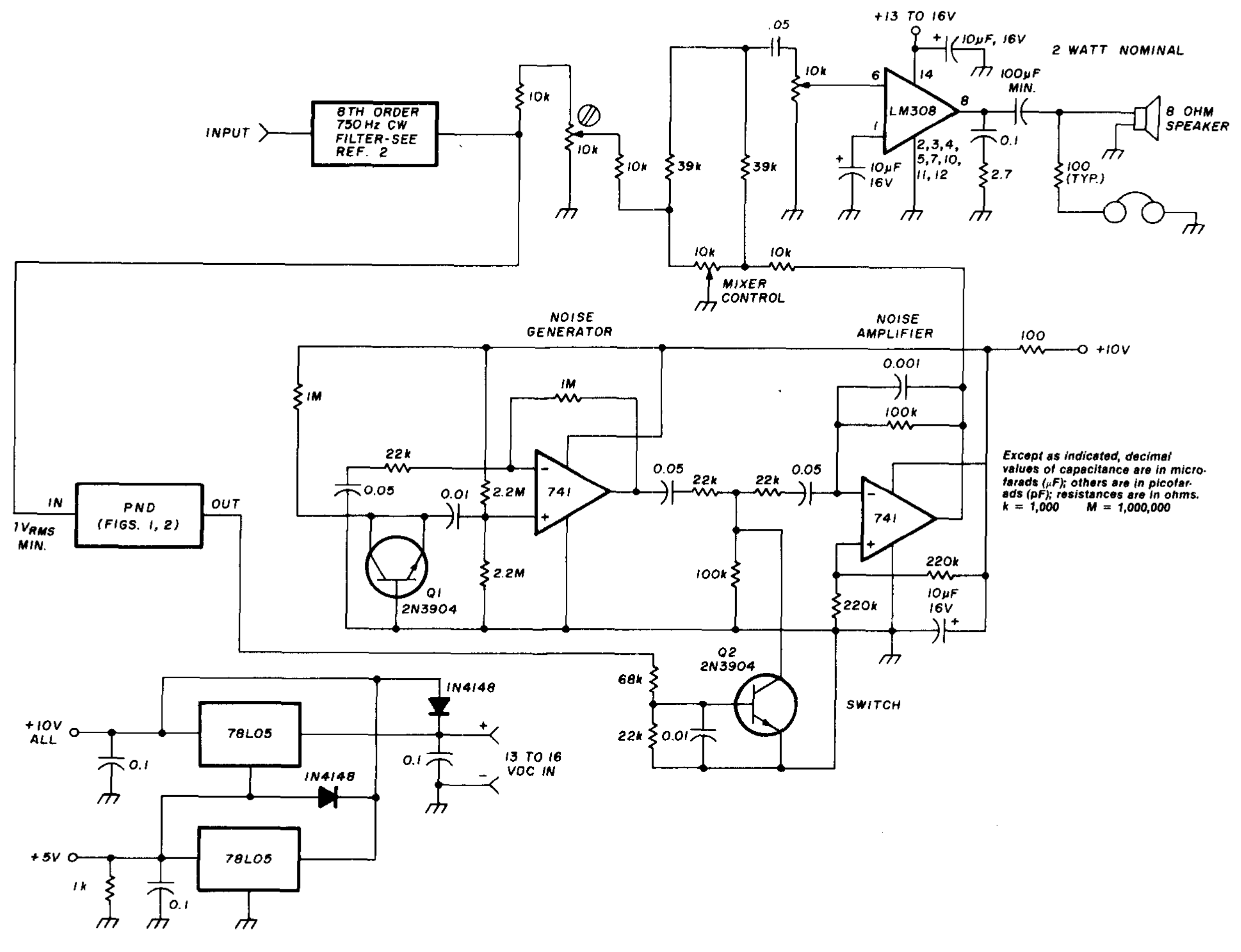
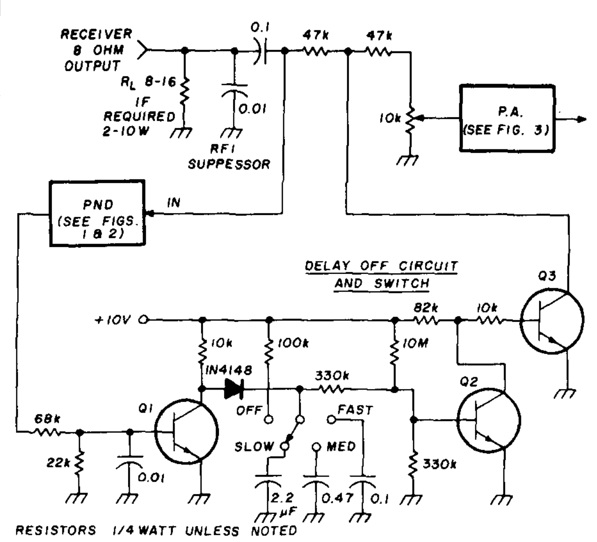


fig. 3. A PND application.



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. 4. Combined PND and squelch for SSB.

on the dial until the noise function starts to be heard sporadically. Increase the delay until the noise stops, then increase the delay just a bit more. This will match the delay to your filter bandwidth. If you are in a location that is too radio-quiet at the time, set the 500K pot at about mid-range for a nominal 12 millisecond delay. This will usually enable the rejection of noise pulses as they are stretched by a typical 100 Hz bandwidth filter. For best operation, the amplitude of a signal being received should be adjusted by your receiver's output level control to just a little above that required for reliable keying of the noise generator. Once this is done, the control at the output of the filter is set so that either the perceived level of keyed noise or the tone level are about the same when the mixer control is in any position. Adjust the volume control for the overall listening level desired. Once all of these settings are made, you will usually work with only your receiver's output control and the mixer control.

single sideband

Although the electronic switch as driven by the PND system can follow code quite well, its use on SSB would result in choppy voice reception. To use PND on sideband it is best to insert a delay circuit between the PND circuitry and the electronic switch to keep the controlled audio stage ON for a short time after the PND has shut off. This technique was used with my laryngeal squelch[3] and that part of the circuit is included in **fig. 4**. If you use the voice-filter system,[2] you can either use PND to open both channels, or you can feed a sample of the vowel filter to PND and control the consonant filter with the delay/switch combination. If you use the vowel filter, the delay setting is about the same as that used for CW. However, if the normal 3 kHz voice bandwidth is used the delay may usually be reduced to its minimum. The PND system, in its basic form shown here, is not as effective for SSB as it is when used to key the noise generator for CW, for although noise is rejected between a voice signal's ON times, it can appear in addition to the desired signal during the ON periods. Circuit development to improve on this problem is being studied, but requirements are much more complex.

general considerations

Although PND can handle most impulse noise problems unaided by the more prevalent noise silencers, it is still best to have a Lamb-type noise blanker in addition. Since the basic noise blanker blocks out noise bursts early in the receiver, it reduces the probability that strong pulses will drive one or several amplifier stages into heavy saturation, which can block a receiver for periods much longer than the offending noise-pulse length, and this is something that PND cannot help. Also, the front-end blanker system can suppress auto ignition pulses produced by an engine at moderately high RPMs, while PND is limited to rejecting auto ignition at idling RPMs when a 100 Hz bandwidth is used and at higher RPMs only when the bandwidth is increased. At the same time, PND can wipe out those woodpecker noise sources while their ON periods are too long for the Lamb blanker to handle. PND can be most effective when it is used in conjunction with either a blanker or a limiter, but use with a blanker is my first choice.

Because we are accustomed to using noise floor as a guide — and with PND you lose this reference — it's easy to run up your receiver gain much too high. Unfortunately, when the gain is too high, the noise-floor itself, even without an antenna, creates what amounts to a constant ringing level as it is stuffed through filters. Moreover, since the noise-floor in linear receivers is not limited, amplitude variations on this ringing can make a weird form of noise-floor-generated CW by the PND. When you hear this, just back off on the gain control a little to achieve silence.

references

1. Don E. Hildreth, W6NRW, "Graphic Filter Design," *ham radio*, April, 1984, page 38.
2. Don E. Hildreth, W6NRW, "Communications Audio Processor for Reception," *ham radio*, January, 1980, page 71.
3. Don E. Hildreth, W6NRW, "Smart Squelch," *ham radio*, June, 1983, page 37.

ham radio

A printed-circuit board (**fig. 3**) is available for $9.95 postpaid; the PC board *plus parts* (no controls) is available for $32.95 postpaid. Contact the author, Don Hildreth, W6NRW, P.O. Box 60003, Sunnyvale, California 94088.

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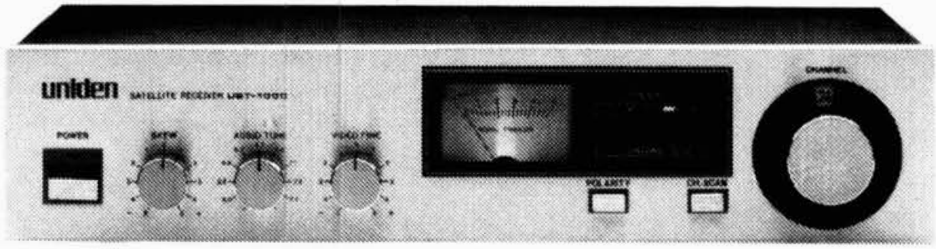

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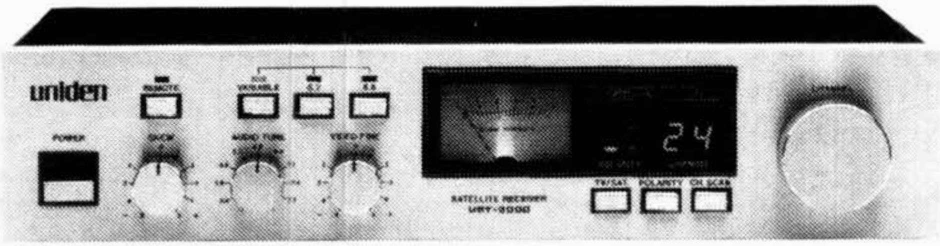

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IMD and intercept points
of cascaded stages

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Intercept point is a useful concept in predicting the spurious intermodulation products generated by components, systems, or subsystems. Only the second-order and third-order products are significant. The intercept point is the power level at which the spurious response equals the fundamental response. The value may be referred to the input or output. Usually the output is referenced for amplifiers and mixers, and the input for receivers. The intercept points for the second-order and third-order products may be the same or different, depending on the circuit of the device. Typically the responses are plotted using log-log scales with the values in dBm as shown in **figs. 1** and **2**.

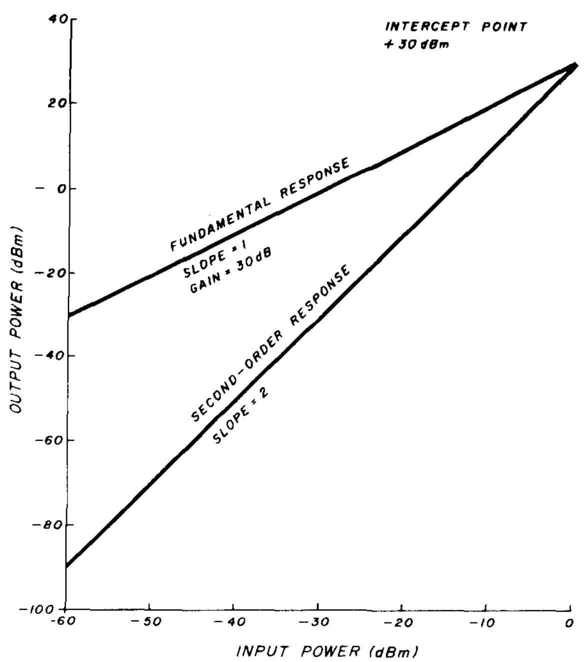

fig. 1. Typical second-order response.

Assuming two signals at frequencies f_A and f_B, and $f_A > f_B$, the second-order products are: $f_A + f_B$, $f_A - f_B$, $2f_A$, and $2f_B$. The second harmonics are not strictly intermodulation products, but may be predicted in the same manner except that their amplitudes are 6 dB less than the sum and difference products. If the two fundamental frequencies are almost equal, the $f_A - f_B$ term is near zero frequency and the remaining product is at about twice the fundamental frequencies. Half-octave filters can be used to attenuate the second-order products. Refer to **fig. 3** for the worst case with f_A and f_B at the band edges of a half-octave filter.

Third-order intermodulation products present the most serious problem for devices having bandwidths less than one-half octave. For two signals at frequencies f_A and f_B, the third-order products are: $2f_A + f_B$, $2f_A - f_B$, $2f_B + f_A$, and $2f_B - f_A$. For a narrow-band device centered at 20 MHz, two signals, $f_A = 20.50$ MHz and $f_B = 20.25$ MHz, will generate the third-order product $2f_B - f_A$ at exactly 20 MHz. For three signals at frequencies f_A, f_B, and f_C, the third-order products are: $\pm f_A \pm f_B \pm f_C$. Third-order products of three signals are seldom considered except for multi-frequency systems such as cable TV.

measurement techniques

For single or cascaded components, intercept point is measured by driving the device with two equal amplitude signals and measuring the fundamental outputs and intermodulation products on a spectrum analyzer.

The concept of intercept point for a receiver is usually limited to the RF front end. It is meaningless for the IF passband because of the nonlinearities of detection

By William Richardson, W3IMG, 1003 Wagner Road, Baltimore, Maryland 21204

and gain control, as well as the high overall gain. Two equal level signals, outside the IF passband but within or as close as possible to the RF passband, are selected so that an intermodulation product is at the receiver center frequency. Their levels are simultaneously increased until an output (intermod) signal of about 10 dB signal-to-noise is observed. Record the level of the signals. Then the two signals are removed, and a single signal at the receiver center frequency is adjusted in level to produce the same output. Its level is also recorded. For third-order products the two signals are usually placed within the RF passband. However, for second-order products the signals fall out of the RF passband if the RF bandwidth is less than an octave. Refer to **fig. 3**. The most important second-order product in a receiver comes from a signal at one-half center frequency that doubles into the center frequency. This latter measurement is made by increasing the amplitude of a signal at one-half the receiver center frequency until an output signal of about 10 dB signal-to-noise is observed. Record the input level. This signal source must be well filtered so that its second harmonic is well below the second-order response. Next the input signal is tuned to the receiver center frequency and its level is adjusted to produce the same output, and this level is recorded.

When connecting two signals to the input, the insertion loss of the combiner must be subtracted from each generator output. A second precaution is to make all measurements at least 10 dB below the 1 dB compression point. Otherwise the device will be operating in its large signal area.

second-order products

Refer to **fig. 1**. The slope of the second-order response is 2. As the fundamental output decreases by 1 dB, the second-order intermodulation products decrease by 2 dB. For two equal signals, the function may be expressed as:

$$IP = P + IMR \qquad (1)$$

IP is the second-order intercept point in dBm, P is the fundamental response in dBm, and IMR is the ratio between the fundamental and second-order responses in dB. In the case of a receiver, P is the level of the two signals or half-frequency input, and IMR is the ratio of P to the level of the signal at center frequency.

For example, if the fundamental outputs of an amplifier are -10 dBm and the second-order intermodulation products are -45 dBm, the second-order intercept point is:

$$IP = -10 + 35 = +25 \text{ dBm}$$

Knowing the intercept point, this equation will predict the second-order intermodulation products for known signal levels. For a single input signal, the same

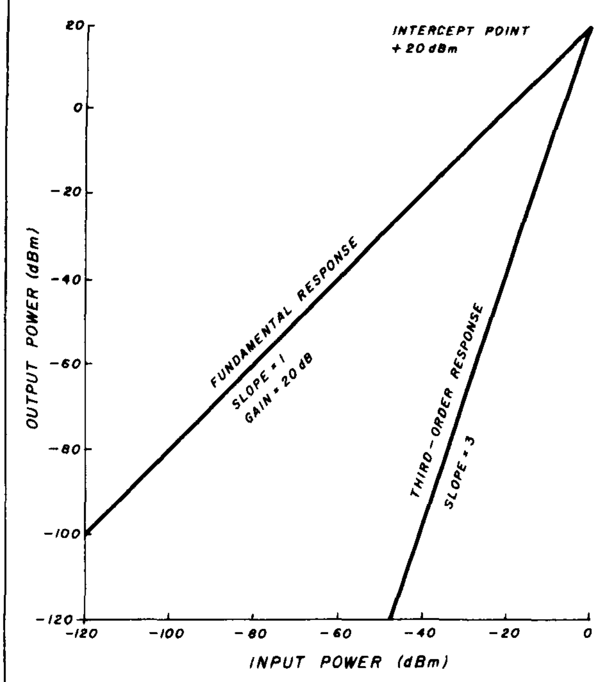

fig. 2. Typical third-order response.

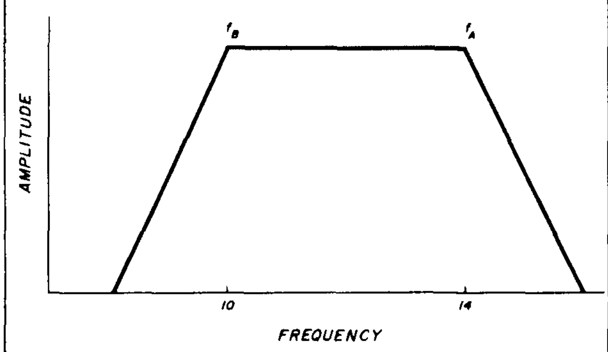

fig. 3. Half-octave filter second-order products.

methods can be used to calculate the intercept point if the second harmonic is measured, or to predict the second harmonic if the intercept point is known. However, the second-harmonic response is 6 dB less than a second-order intermodulation product.

The levels of the second-order intermodulation products are proportional to the product of the levels of the fundamental signals. For signals of unequal levels, the level of equal signals that produce the same intermodulation products can be calculated. If the levels are in dBm, add the two and divide by one-half. For two signals, one at -20 dBm and the other at -26 dBm, the equivalent equal amplitude signals are

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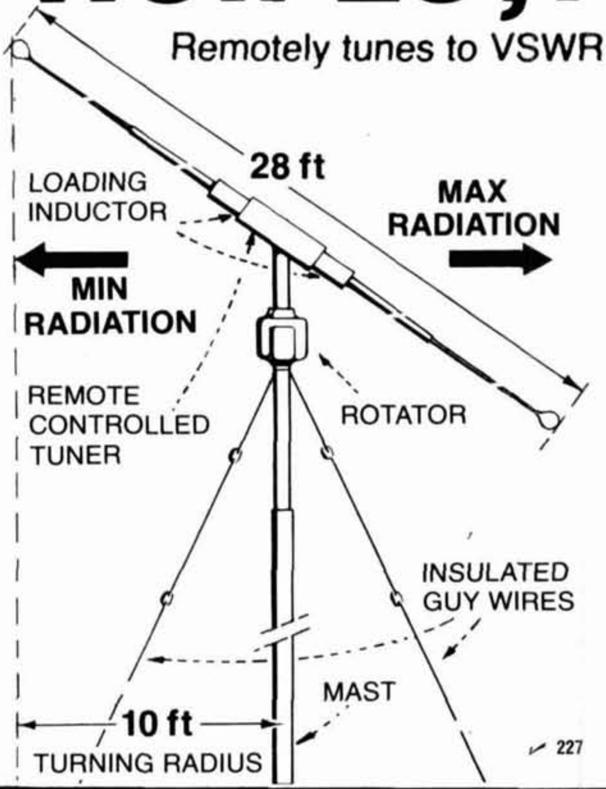

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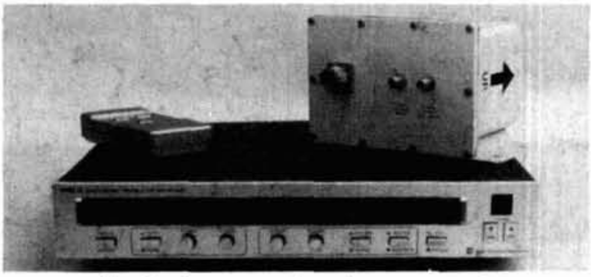

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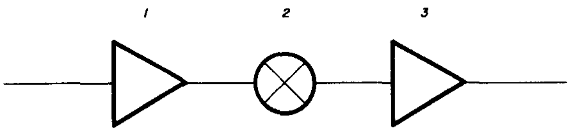


stage	IP dB	IP ratio	gain dB	gain ratio
1	+20	100	10	10
2	+13	20	−6	0.25
3	+30	1000	20	100

using cascaded stage equation

$$\frac{1}{\sqrt{IP}} = \frac{1}{\sqrt{IP_1 \cdot G_2 \cdot G_3}} + \frac{1}{\sqrt{IP_2 \cdot G_3}} + \frac{1}{\sqrt{IP_3}}$$

$$= \frac{1}{\sqrt{100 \cdot (0.25) \cdot 100}} + \frac{1}{\sqrt{20 \cdot 100}} + \frac{1}{\sqrt{1000}}$$

$$= 0.0200 + 0.0224 + 0.0316 = 0.074$$

$$IP = 182 = 22.6 \text{ dBm}$$

Alternate method

$$\Delta Ip = 20 \log \left[1 + \sqrt{\frac{IP_2}{IP_1 \cdot G_2}} \right]$$

For stages 1 and 2:

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{20}{100 \cdot 0.25}} \right] = 20 \log 1.89 = 5.5 \text{ dB}$$

Combining stages 1 and 2

$$IP = IP_2 - \Delta IP = 13 - 5.5 = 7.5 \text{ dBm} = 5.6 \text{ milliwatts}$$

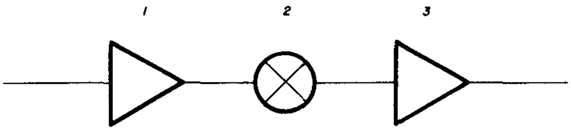
Now cascade this with stage 3

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{1000}{5.6 \cdot 100}} \right] = 20 \log 2.34 = = 7.4 \text{ dB}$$

$$IP = IP_3 - \Delta IP = 30 - 7.4 = 22.6 \text{ dBm}$$

Input intercept point = 22.6 − 34.0 = −11.4 dBm

fig. 4. Example shows two methods of calculating cascaded stages second-order intercept point.



stage	IP dB	IP ratio	gain dB	gain ratio
1	+20	100	10	10
2	+13	20	−6	0.25
3	+30	1000	20	100

using cascaded stage equation:

$$\frac{1}{IP} = \frac{1}{IP_1 \cdot G_2 \cdot G_3} + \frac{1}{IP_2 \cdot G_3} + \frac{1}{IP_3}$$

$$\frac{1}{IP} = \frac{1}{100 \cdot (0.25) \cdot 100} + \frac{1}{20 \cdot 100} + \frac{1}{1000}$$

$$= 0.00040 + 0.00050 + 0.0010 = 0.0019$$

$$IP = 526 = 27.2 \text{ dBm}$$

Alternate method

$$\Delta IP = 10 \log \left[1 + \frac{IP_2}{IP_1 \cdot G_2} \right]$$

For stages 1 and 2

$$\Delta IP = 10 \log \left[1 + \frac{20}{100 \cdot 0.25} \right] = 10 \log 1.80 = 2.6 \text{ dB}$$

Combining stages 1 and 2

$$IP = IP_2 - \Delta IP = 13 - 2.6 = 10.4 \text{ dBm} = 11.0 \text{ milliwatts}$$

Now cascade this with stage 3

$$\Delta IP = 10 \log \left[1 + \frac{1000}{11 \cdot 100} \right] = 10 \log 1.91 = 2.8 \text{ dB}$$

$$IP = IP_3 - \Delta IP = 30 - 2.8 = 27.2 \text{ dBm}$$

Input intercept point = 27.2 − 24 = 3.2 dBm

fig. 5. Example shows two methods of calculating cascaded stages third-order intercept point.

each at −23 dBm. If the levels are in milliwatts, the equivalent equal signal levels are each:

$$\sqrt{A_A \cdot A_B}$$

where A_A and A_B are individual signal levels.

If the second-order intercept point and gain of each stage are known, the overall intercept point of the cascaded stages may be found from the following formula:

$$\frac{1}{\sqrt{IP}} = \frac{1}{\sqrt{IP_1 \cdot G_2 \cdot G_3 \cdot G_4}} + \frac{1}{\sqrt{IP_2 \cdot G_3 \cdot G_4}} + \frac{1}{\sqrt{IP_3 \cdot G_4}} + \frac{1}{\sqrt{IP_4}} \quad (2)$$

Each term of the formula has the intercept point of the stage multiplied by the gain of all of the following stages. The terms are numerical ratios, not dB or dBm. A look at each term will indicate the contribution of each stage to the overall system intercept point.

Another method of calculating the overall intercept point of cascaded stages is to use the formula:

$$\Delta IP = 20 \log \left[1 + \sqrt{\frac{IP_2}{IP_1 \cdot G_2}} \right] \quad (3)$$

The ΔIP is in dB and is subtracted from the second stage intercept point to give the overall intercept point of the two cascaded stages. For more than two stages, this formula is used for the first two stages, and that result is then used with the third stage and so forth.

Both formulas assume the worst case in which the intermodulation products within each stage *add in phase*. If a linear stage is part of the system, it must be included with its actual gain (or loss) and an infinite intercept point.

Refer to **fig. 4** for sample calculations of the intercept point of three cascaded stages. The output intercept point is calculated. The input intercept point

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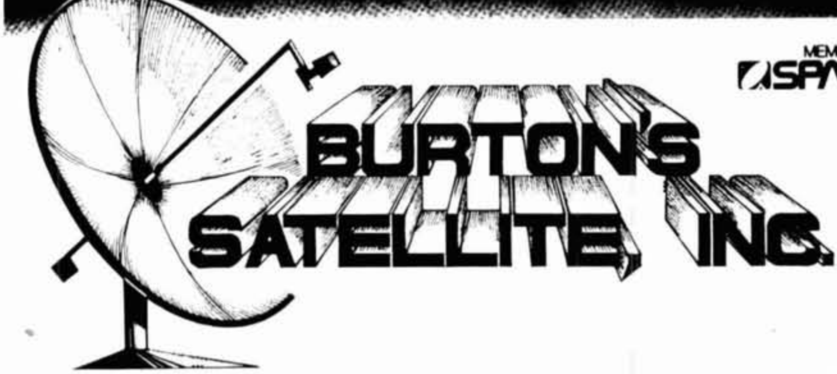

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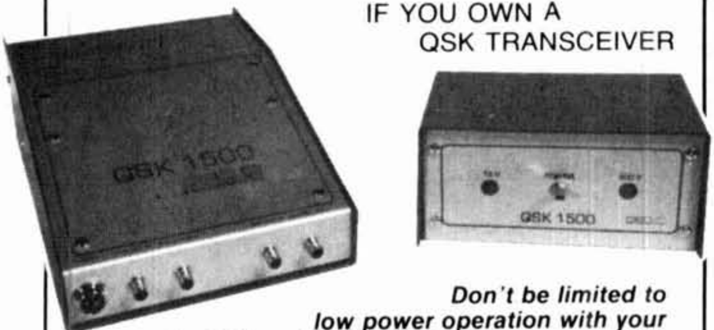

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equals the output intercept point reduced by the total gain.

third-order products

Refer to **fig. 2.** The slope of the third-order response is 3. As the fundamental output decreases by 1 dB, the third-order intermodulation products decrease by 3 dB. For equal signals, the curve may be expressed as:

$$IP = P + 1/2(IMR) \qquad \textbf{(4)}$$

IP is the third-order intercept point in dBm, and IMR is the ratio between the fundamental and third-order responses in dB. For the case of a receiver, P is the level of the two input signals and IMR is the ratio of P to the level of the signal at center frequency.

For example, if the fundamental outputs of an amplifier are −10 dBm and the third-order intermodulation products are −50 dBm, the third order intercept point is:

$$IP = -10 + 1/2(40) = +10 \ dBm$$

Knowing the intercept point, the equation will predict the third-order intermodulation products for known signal levels.

The levels of the third-order intermodulation products are proportional to (1) the cube root of the product of three signals or (2) in the case of two signals, the cube root of the square of the higher level signal times the other. For signals of unequal levels, the equivalent equal level signals that produce the same intermodulation products can be calculated. If the levels are in dBm, for the two signals add 2/3 of the larger to 1/3 the smaller. If one signal is at −20 dBm and the other at −32 dBm, the equivalent equal level signals are at −24 dBm. For three signals, add 1/3 of each level in dBm. If the levels are in milliwatts, the equivalent levels are

$$\sqrt[3]{A_A \cdot A_B \cdot A_C} \text{ or } \sqrt[3]{A^2_A \cdot A_B}$$

where A_A is the highest level.

If the third-order intercept point and gain of each stage are known, the overall intercept point of cascaded stages may be found from the following formula:

$$\frac{1}{IP} = \frac{1}{IP_1 \cdot G_2 \cdot G_3 \cdot G_4} + \frac{1}{IP_2 \cdot G_3 \cdot G_4} + \frac{1}{IP_3 \cdot G_4} + \frac{1}{IP_4} \qquad \textbf{(4)}$$

Each term of the formula has the intercept point of the stage multiplied by the gain of all the following stages. The terms are numerical ratios, not dB or dBm. A look at each term indicates the contribution of each stage to the overall system intercept point.

Another method of calculating the overall intercept point of cascaded stages is to use the formula:*

$$\Delta IP = 10 log \left[1 + \frac{IP_2}{IP_1 \cdot G_2} \right] \qquad \textbf{(5)}$$

The ΔIP is in dB and is subtracted from the second-

```
10 PRINT"INTERCEPT POINT OF CASCADED STAGES"
20 PRINT:PRINT
30 PRINT"2  SECOND-ORDER INTERMODULATION"
40 PRINT"3  THIRD-ORDER INTERMODULATION"
50 INPUT"SELECT 2 OR 3";A$
60 INPUT"ENTER NUMBER OF STAGES";C
70 CLS
80 FOR N=1 TO C
90 PRINT"ENTER INTERCEPT POINT FOR STAGE";N;"IN DBM"
100 INPUT I(N)
110 PRINT"ENTER GAIN OF STAGE";N;"IN DB"
120 INPUT G(N)
130 NEXT N
140 CLS
150 PRINT"IP(DBM)","GAIN(DB)
160 PRINT
170 FOR N=1 TO C
180 PRINT I(N),G(N)
190 NEXT N
200 PRINT:INPUT"IS DATA OK Y/N";B$
210 IF B$="N" THEN CLS
220 IF B$="N" THEN GOTO 80
230 FOR N=1 TO C
240 I(N)=10[(I(N)/10)
250 G(N)=10[(G(N)/10)
260 NEXT N
270 E(C)=1:D(C)=1/I(C):IF A$="2" THEN D(C)=SQR(D(C))
280 FOR N=(C-1) TO 1 STEP -1
290 E(N)=G(N+1)*E(N+1)
300 D(N)=I(N)*E(N)
310 D(N)=1/D(N)
320 IF A$="2" THEN D(N)=SQR(D(N))
330 NEXT N
340 FOR N=1 TO C
350 IP(N)=D(N)+IP(N-1)
360 NEXT N
370 IP(C)=1/IP(C)
380 IP(C)=10*LOG(IP(C))/LOG(10):PRINT
390 IF A$="2" THEN IP(C)=2*IP(C)
400 IF A$="2" THEN PRINT "SECOND-ORDER INTERCEPT POINT IS ";IP(C);" DBM"
410 IF A$="3" THEN PRINT "THIRD-ORDER INTERCEPT POINT IS ";IP(C);" DBM"
```

fig. 6. TRS-80 Model III program listing determines the intercept point of cascaded stages.

```
INTERCEPT POINT OF CASCADED STAGES


2  SECOND-ORDER INTERMODULATION
3  THIRD-ORDER INTERMODULATION
SELECT 2 OR 3? 3
ENTER NUMBER OF STAGES? 3
ENTER INTERCEPT POINT FOR STAGE 1 IN DBM
? 20
ENTER GAIN OF STAGE 1 IN DB
? 10
ENTER INTERCEPT POINT FOR STAGE 2 IN DBM
? 13
ENTER GAIN OF STAGE 2 IN DB
? -6
ENTER INTERCEPT POINT FOR STAGE 3 IN DBM
? 30
ENTER GAIN OF STAGE 3 IN DB
? 20
IP(DBM)           GAIN(DB)

 20                10
 13               -6
 30                20

IS DATA OK Y/N? Y

THIRD-ORDER INTERCEPT POINT IS  27.2141  DBM
```

fig. 7. Three-stage device IMD intercept point calculation is simple with user-friendly program.

*Note differences between **eqs. 2** and **4** and **3** and **5**. — **Editor.**

stage intercept point to give the overall intercept point of the two cascaded stages. For more than two stages, this formula is used for the first two stages, and the result is then used with the third stage and so forth.

Both formulas assume the worst case in which the intermodulation products within each stage add in phase. If a linear stage is part of the system, it must be included with its actual gain (or loss) and an infinite intercept point.

Refer to **fig. 5** for a sample calculation of third-order intercept point for three cascaded stages. The output intercept point is calculated. The input intercept point equals the output intercept point reduced by the total gain.

computer program aids calculation

Figure 6 lists the steps of a typical BASIC language computer program for calculating the second-order and third-order intercept points of cascaded stages if the values for the individual stages are known.

Figure 7 is a TRS-80 Model III™ printout showing a typical calculation of third-order IMD intercept point of a three-stage device.

ham radio

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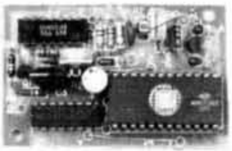

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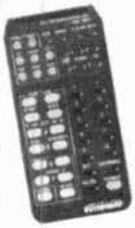

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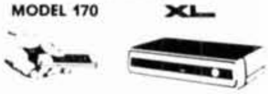

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the Russian Woodpecker: a continuing nuisance

What it is and what can be done

It never fails . . . you're working some choice DX on 20 meters for your 300th country or you're in one of the Area traffic nets, trying to pass a message to another ham a thousand miles away. Suddenly, without warning, the band is shattered by something that sounds like a cross between a machine gun and a jackhammer. No, it's not the neighbor's power saw or the family microwave oven . . . it's the Russian Woodpecker in full operation. With the interference level running 10 to 20 dB over S9, your much-needed contact is buried under this avalanche of QRM and heard no more. The only solution is to turn off the rig and cool down with a tall, cold 807.

What is the Woodpecker? Why is it in operation? And why does the Soviet Union persist in creating this level of interference worldwide? What can we do about it? And what have other radio services and users of the spectrum done? This article will explore the aspects of this problem and suggest some possible solutions.

Basically, the Russian Woodpecker is an extremely powerful over-the-horizon radar system. It operates over most of the HF band, with effective radiated power levels of some 10 to 50 megawatts. To understand the nature of this problem, we need to have a basic understanding of how OTHB (Over-The-Horizon-Backscatter) radar operates, some sense of the history of experimentation and operation in this field, and an educated awareness of Soviet diplomatic response to complaints about the interference their system is generating.

basic radar operation

It has long been recognized that radar can be operated on any frequency. The earliest radar systems — built by the British and responsible for much of that nation's success during the Battle of Britain — were operated at a frequency around 30 MHz. This was due, in part, to 30 MHz being the highest frequency at which significant levels of power could be generated. Later radar systems were operated at much higher frequencies as technology developed tubes capable of generating multi-kilowatt levels of power at shorter and shorter wavelengths. Moving through the spectrum from VHF to UHF and finally into the microwave regions, radar engineers have traditionally sought the highest possible frequency of operation for several reasons. Shorter wavelengths bring increased target resolution and give the system, as a whole, increased immunity to interference, either natural or man-made. In addition, highly directional antennas become physically smaller, making possible the design of mobile radar units with greater target discrimination.

Unfortunately, all these radar systems suffer from a single common defect: they can operate only on line-of-sight. This means that at greater distances, the target must be at higher altitudes in order to be detected. Aircraft or cruise missiles flying at very low altitudes can escape radar detection

By **Bradley Wells, KR7L**, 5053 37th Avenue, S.W., Seattle, Washington 98126

until they are almost on top of the radar transmitter. Thus, several aircraft flying at tree-top level could approach and neutralize radar installations undetected, leaving a blind spot through which an enemy could pour aircraft or missiles. This scenario, dealing with the problem of low level detection, has left many a defense planner, both American and Soviet, in a cold sweat.

lower frequencies provided new opportunities

It had been recognized by many that some form of high-frequency radar, utilizing backscatter techniques, could detect these low-level targets. Since the radar signal would be reflected off the ionosphere and illuminate the target from above, there would be no escape from this type of detection system. It was also recognized that there were several inherent problems in this approach. First, the ionosphere was thought to be in a state of continuous flux, unable to provide stable refraction characteristics for any length of time; second, there would be continuous interference both to and from other users of the HF spectrum; and, finally, the reception of backscattered signals would require extremely complex detection systems.

By the early 1970's, scientific inquiry and experiments brought new light to this gloomy picture. The widespread use of ionospheric sounders, both ground-based and satellite, had shown the ionosphere to be more stable than previously thought. It was discovered that the refractive characteristics of the ionosphere changed very little in the short term — that is, for periods of approximately 30 minutes, the ionosphere is remarkably homogeneous. During the course of a day, these characteristics change in response to shifts of solar flux and geomagnetic activity. This meant a radar system would have to be capable of operating over much of the HF band to provide coverage of selected areas. Simply put, the radar would have to be frequency-agile to follow these changes in the Maximum Usable Frequency (MUF).

The explosion of computer technology made possible the correlation and analysis of weak backscattered signals on a real-time basis. Using cross-correlation reception techniques coupled with the development of magnetic drums for data storage, high-speed computers were used to sort out interference in the system. These computers could not only discern a weak target signal from ground clutter but also selectively filter out other users of the HF spectrum.

early OTHBs

In the 1950's, the United States Naval Research Laboratory and other groups began small-scale experiments with OTHB radar. These early experiments led to the solution of some of the major problems in designing a functional HF radar. Among these problems were the following:

- The return from prospective targets would be 40 to 80 dB weaker than the ground return (i.e., ground clutter).
- It was not known whether sufficient angular resolution could be developed at HF wavelengths to permit accurate target identification.
- Extremely precise doppler techniques would have to be used to permit identification.

The magnitude of this doppler problem may be seen in the following equation:

$$fd = \frac{2\,V_r f_o}{C}$$

fd represents the doppler shift, f_o represents the radar carrier frequency, V_r is the target relative velocity, and C is the speed of light. For aircraft type targets, the doppler shift varies from tenths of a Hertz upward to 50 Hz. This is dependent upon the operating frequency. The development of technologies to deal with these and other problems have resulted in the operation of both American and Soviet OTHB radar systems.

Both the American CONUS OTH-B (Continental United States Over-The-Horizon Backscatter — see sidebar, page 43) and the Soviet Woodpecker share characteristics common to all HF radars. The interaction of these characteristics may be seen from an examination of the radar range equation:

$$R_{MAX} = \frac{P_{AV}\, G_T\, G_R\, \lambda^2\, \sigma\, F_P\, T_C}{(4\pi)^3\, N_o\, (S/N)\, L_S}$$

where R_{MAX} = maximum range
P_{AV} = average power
G_T = gain of transmitting antenna
G_R = gain of receiving antenna
λ = wavelength
σ = target cross section
F_P = propagation effects factor
T_C = coherent processing time
N_o = noise power/unit bandwidth
S/N = signal to noise ratio required for detection
L_S = system losses

The major differences between HF and microwave radar systems are related to the following:

- Propagation effects — energy loss over ionospheric paths, polarization mismatch between transmitted and received signals, and gains or losses related to the dynamic nature of the transmission path.
- The amount of noise injected into the system by natural sources (i.e., distant thunderstorms) and, more importantly, by other users of the HF spectrum

(e.g., international broadcasting, Amateurs, Maritime mobile, etc.).

• Processing time (the number of hits integrated divided by the pulse-repetition frequency) — important since doppler radar requires a dwell time of T_C seconds to realize a frequency resolution of $1/T_C$ Hertz.

The transmitted waveform for HF radar systems is similar to that of microwave systems. It can be CW, pulse, FM-CW, or some other coded mode of transmission. OTHB radar have different problems with detection at minimum ranges than do microwave radars. This is because of the existence of a skip zone — that region, familiar to all hams, from which no signal is received. This skip zone accounts for HF radar's inability to detect targets closer than 1000 km to the transmitter.

A long pulse is used in HF radar to increase the sensitivity of the system and may reduce to interference levels associated with pulse modulation. In addition, the pulse repetition frequency is normally low to avoid range ambiguities. A PRF of 50 Hz will yield an unambigious range of some 3000 km. Individual pulse widths may range from tens of microseconds to several milliseconds depending upon the sensitivity desired and the desire to reduce interference to other services.

antenna requirements are severe

OTHB radar places more demands upon the antenna system than do other types of radar. The antenna must be physically large because of the low frequencies involved, be capable of handling very large amounts of power, exhibit gain and directivity over a wide range of frequencies, and be steerable in both elevation and azimuth. Typically the antenna consists of phased arrays of vertical bowtie driven elements in front of screen reflectors. The antenna lobes are steered in azimuth and elevation by shifting the phase relationships between individual active elements. Normally, separate antennas are used for transmitting and receiving. While this increases the problems of synchronizing the transmitter with the receiver, it is more than offset by the simplification of antenna construction. To place the first lobe as near horiziontal as possible, an extended ground screen is placed in front of the array. This ground screen may extend up to 3000 meters in front of and be as wide as the antenna array.

Changes in the ionosphere bring about changes in the MUF. HF radar adapts to these changing conditions by shifting its operating frequency. The ionosphere is probed with a sounder and the profiles are updated constantly. This gives real-time information as to what the best operating frequency for coverage of a particular area of interest will be. The relationship between vertical profiles and transmission paths can be seen from **figs. 1** and **2**. As the transmission frequency approaches the MUF, the paths lengthen, providing the maximum distance in a single hop transmission. Operating at or near the MUF greatly reduces path losses. Since these radar systems are not limited to a few discrete bands of frequencies, as are other services (including hams), they can follow the MUF quite closely.

The reliability of HF radar is related to antenna size, radiated power, and the range of frequencies used. These factors can overcome shortcomings in the reliability of the ionospheric paths. The ionosphere places limits on operation in both summer and winter, but for different reasons. In summer, ionization extends well into the lower regions, which normally contain neutral particles. Thus strengthened, the D-layer causes increased path loss. During the winter, decreased solar radiation creates lower electron densities in the F-layer and results in lower frequencies being required for reliable transmission. Several other problems exist because of changes within the ionosphere. These problems include the following:

• Propagation velocity is frequency dependent which places lower limits on pulse length and range resolution.

• The refractive characteristics of the ionosphere allow specific areas to be covered only by a narrow range of frequencies at any given moment.

the solar jammer

At frequencies in the high HF and low VHF range, natural extra-terrestrial sources of interference can play havoc with radar systems. During the height of the Battle of Britain, for example, British radars operating around 30 MHz were suddenly jammed by a strange, unknown signal. The interference became so severe that the British High Command felt sure it was some new and very effective form of German jamming. A group of engineers and astronomers, led by Stanley Hey, was detailed to locate the source and develop countermeasures. Together they determined that the interference appeared to originate in the area of the Sun. After photographs revealed a large sunspot group on the surface of the Sun, Hey concluded that the intensity of interference was related to the size and position of the sunspots on the solar surface. This discovery, confirmed by other investigators, led to the post-war development of solar radio astronomy.

table 1. Current capabilities of United States and USSR OTHB radars.

range coverage:	1000-4000 km, with longer ranges possible through multiple-hop transmissions.
angle coverage:	360 degrees in azimuth possible, but less than 120 degrees typical.
range resolution:	As low as 2 km, with 20-40 km typical.
absolute range accuracy:	10-20 km, assuming accurate and timely assessments of the ionosphere and optimum operating frequency.
angle resolution:	Determined by beamwidth. It can be as low as 1 degree, which corresponds to 50 km at a range of 3000 km.
Doppler resolution:	Generally resolution of 1/10 Hz is possible. At an operating frequency of 20 MHz it corresponds to a target velocity of just under 2 MPH.
level of interference:	Dependent upon such factors as frequency of operation, antenna design, power level, type of modulation, area of illumination, and duration of operation at any particular frequency.

• The propagation medium is filled with unwanted clutter from meteor and auroral ionization in addition to other areas of scattering that compete with target returns.

present OTHBs

The current capabilities of OTHB radars, both American and Soviet, are shown in **table 1**.

it all started with Ivan The Terrible

The initial evidence of Soviet OTHB radar capability surfaced in mid-1976. The first of these radar units, nicknamed "The Kiev Buzzsaw" or "Ivan The Terrible" was a 2-million watt transmitter operating near the city of Kiev, augmented by a smaller installation near the Black Sea town of Nikolayev. From these initial efforts, the Soviets have expanded their system into a fully functional early-warning high-frequency radar. Most of the early information concerning the Russian Woodpecker, as it is now known, came from the worldwide efforts of Amateur Radio operators. Even today, little hard information is available concerning the physical make-up of these radar installations. Western intelligence reports remain classified and, or course, the Russians appear reluctant to volunteer anything.

The Woodpecker is part of some 7000 surveillance radar systems deployed at over 1200 sites across the length and breadth of the USSR. While it was initially thought that the Woodpecker was designed for aircraft or ship detection, recent information indicates that it is actually a ballistic missile early warning system. There are currently three of these OTHB radars in operation. Two of them pointed at the United States and the other was directed at central China. These radar systems operate in conjunction with satellite detection systems to provide upwards of 30 minutes warning of an ICBM strike launched from sites within the United States or China. This HF radar launch detection system is not as accurate or reliable as a satellite system, but the two working together give 24 hour-a-day coverage of missile silos.

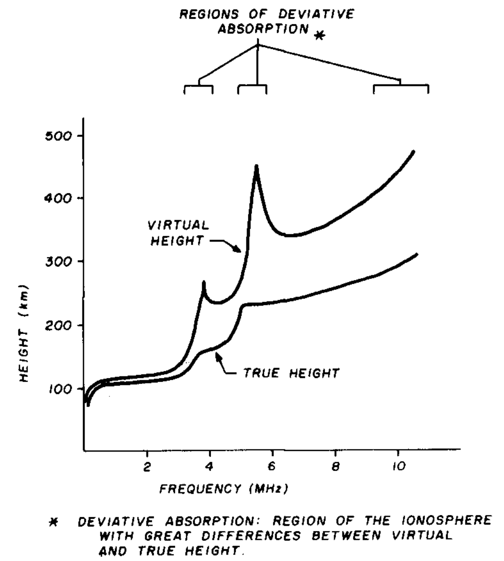

fig. 1. Daytime vertical profile of the ionosphere.*

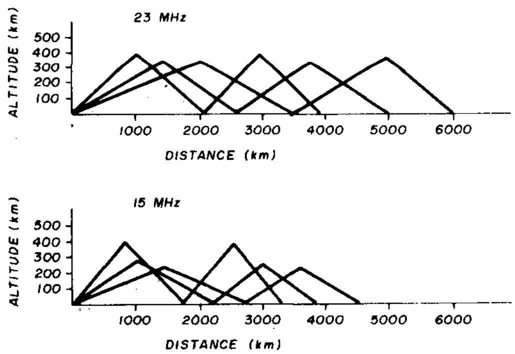

fig. 2. Virtual path traces at two frequencies based on data in *fig. 1*. Radiation angles of 4-12-20 degrees above horizontal. (Adapted from "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, June, 1974.)

*Adapted, with permission, from James M. Headrick's "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, Volume 62, Nov. 6, June, 1974. ©1974, IEEE.

Since its beginning in 1976, the Soviet OTHB systems have increased their power and currently operate at the 20 to 50 megawatt level. Their system utilizes pulse modulation, in contrast to the American CONUS OTHB, which transmits FM-CW. The PRF (Pulse Repetition Frequency) is normally 10 per second, although additional analysis has suggested each pulse actually consists of a pulse train of up to twenty different square wave pulses with some less than two milliseconds long, giving an effective PRF of 800 pps. The modulation scheme employed by the Russians has undergone some evolutionary changes since the inception of this radar system. Currently, the modulation, though still a pulse system, causes the radar signal to be spread in frequency. This permits frequency compression on the receiving end and results in "processing gain" for the system as a whole. In addition, this spread-spectrum technique allows the detection system to more easily discriminate against other stations on frequency. Unfortunately, these wide-band signals have further increased the interference levels to other, legitimate users of the HF band.

Currently, the radar signals no longer sit on one frequency for extended periods of time, as they once did. This is due, in part, to the protests of other users of the HF spectrum, but also to the Soviets' efforts to utilize the optimum transmission frequency. At the present time, the signals move up and down the band in 100 kHz steps at intervals of 30 seconds to 10 minutes.

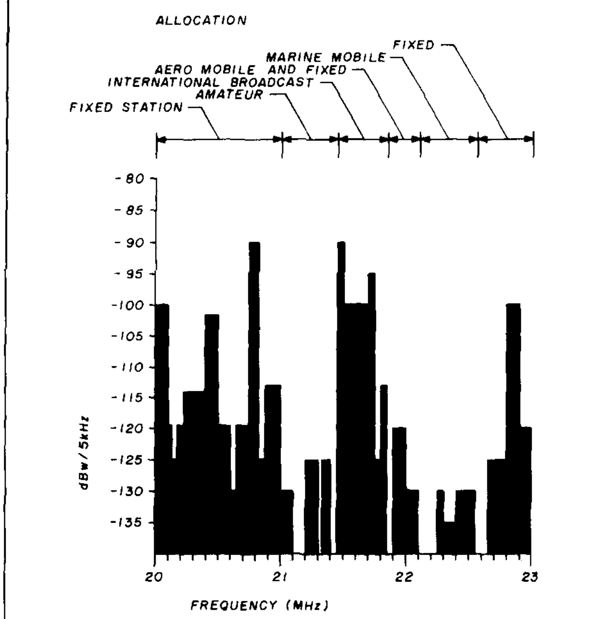

fig. 3. Typical power spectrum at HF. Note correlation with user allocation. Noise floor — 140 dBw (140 dB below 1 watt). (Adapted from "Over-The-Horizon Radar on the HF Band," *Proceedings of the IEEE*, June, 1974.)

why hams are most affected

It is also noteworthy that the Woodpecker chooses parts of the HF spectrum with low rates of RF occupancy. Certain portions of the band have few users per unit time and those users operate with low levels of radiated power. These areas of the spectrum are a natural for radar operation, placing less stringent requirements on the detection system. As can be seen in **fig. 3**, the Amateur bands fit nicely into this category. This helps to explain why hams have suffered the most. In addition, Amateurs tend to have limited political "pull" with their governments and, thus, are less able to bring pressure to bear to curb this interference than are other users of the spectrum. Other services, such as international broadcasting, can overcome the Woodpecker by raising their effective radiated power into the megawatt range and thereby swamping out the Russian radar.

Worldwide response to the Woodpecker arose almost immediately after its first transmission. In July, 1976, the Federal Communications Commission sent a telegram — prompted by complaints from ham operators about interference levels on the 20-meter band — to the Soviet Ministry of Post and Telecommunications. With no response from the USSR, the FCC sent three more cables. Still there was no response. In October, the FCC filed a formal complaint with the International Frequency Registration Board.

Additional complaints poured in from Amateur, Maritime, and aeronautical operators in other countries. In addition to the United States, and European nations, countries in the region of the Baltic Sea as well as Australia and New Zealand voiced strong protests. Early in 1977, the Soviet Union admitted that their experiments might cause some interference to other radio facilities for short periods of time. As worldwide pressure mounted, the USSR agreed to cut back on these radar transmissions. In reality, the Woodpecker remained on the air for the same amount of time, but its signals moved back and forth through the HF band rather than staying in one spot for extended periods of time.

In 1979, this issue surfaced, but was never pressed, at the World Administrative Radio Conference. In retrospect, this was probably for the best. This conference resulted in substantial gains for the Amateur community that might never have come about if the Conference had been disrupted over the Woodpecker issue.

Soviets ignore treaty

The USSR is signatory to international telecommunications treaties that spell out, in detail, the allocations for broadcasting. However, the Soviets have made full use of an escape clause included in all of these treaties. Simply put, a nation may ignore the treaty if such action is deemed to be in the best interests of its national defense. In addition, telecommunications treaties are only as good as a nation's willingness to abide by them. There is no practical way to force compliance by other countries. Most nations observe these treaties rather closely, however, realizing that compliance is in the best interest of the world community.

The current position of the United States was recently stated by Dr. William Schneider, Under Secretary of State for Security Assistance, Science and Technology. In an interview, Schneider commented, "We are making every effort to encourage the USSR to comply with their treaty obligations. In this regard, I hope we will be more effective in the future than we have been in the past."[1] In reality, this means that the Soviets will continue to use the Woodpecker as long as it suits their needs or until they develop a completely accurate and reliable satellite surveillance system for ICBM launch detection.

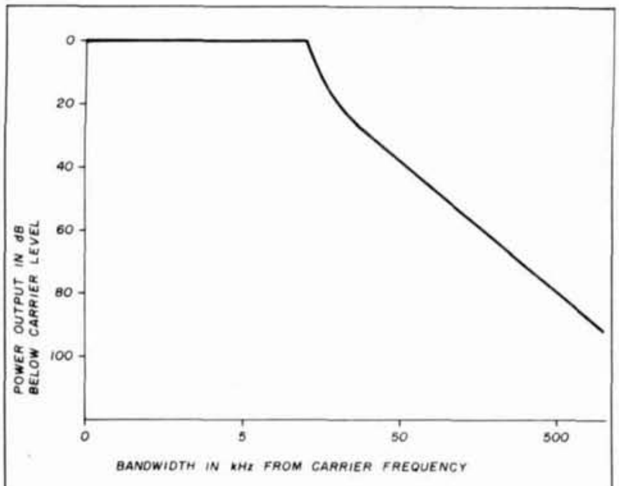

fig. 4. Spectral distribution of signal from the Russian Woodpecker (assuming 50 Mw carrier ERP, note that the Woodpecker still has 5000 watts ERP 50 kHz either side of center frequency).

what can we do?

So what can you do the next time the Woodpecker blows the 20-meter band apart? There's no point in complaining to the FCC or Department of State — they're not interested. They have literally *thousands* of complaints on file and don't need any more. They are fully aware of the problem and realize how little they can do to change it. Cranking your keyer up to 99 WPM and shooting a string of pulses in the direction of the Soviet Union is equally futile. Because the radar is designed to ignore this type of interference, all this accomplishes is additional QRM for other hams.

Perhaps the best solution to Woodpecker interference lies in the field of electronics. The technology is available to eliminate this pest at the receiver. The newer transceivers, such as the Kenwood TS-930, the ICOM IC-751, and the Yaesu FT-1, among others, have dual noise blankers, one of which is designed to eliminate long pulse noise such as that from the Woodpecker. This trend is likely to continue until most new rigs have this capability.

All this doesn't help those of us who aren't quite ready to buy a new state-of-the-art transceiver. What can we do? We have two choices. The first is to build a custom noise blanker for our existing rigs. Circuits to eliminate the Woodpecker have been published in Amateur magazines and in the ARRL's *Radio Amateur's Handbook*. The second choice is the purchase of a "Moscow Muffler," a Woodpecker noise blanker manufactured by AEA (Advanced Electronics Applications) of Lynnwood, Washington. Installed between the transceiver and antenna, this unit effectively blanks out the Woodpecker by means of an RF sensing unit that automatically takes it out of the circuit when the transmitter is keyed. The blanking width and synchronization are both adjustable. The basic sync rate may be switched between 10 and 16 Hz to allow for blanking when both OTHB radars are in operation.

It does not appear that the Woodpecker will dis-

interference not inevitable

The USA's CONUS OTH-B radar has received widespread publicity in technical, professional, and Amateur publications. At the onset of operation, the project's organizers actually *solicited* interference reports from all users of the HF spectrum. A committee was set up to handle the expected deluge of complaints; after two years of operation, only eight reports had been received. Of these, seven were disallowed because the radar had not been in operation at the time the alleged interference occurred or because the radar was operating on a frequency far removed from the one specified in the complaint. The eighth report was not a complaint at all, but rather a report from an SWL looking for confirmation. This absence of interference to other services is due to the nature of the American radar and the care exercised in the selection of clear frequencies.

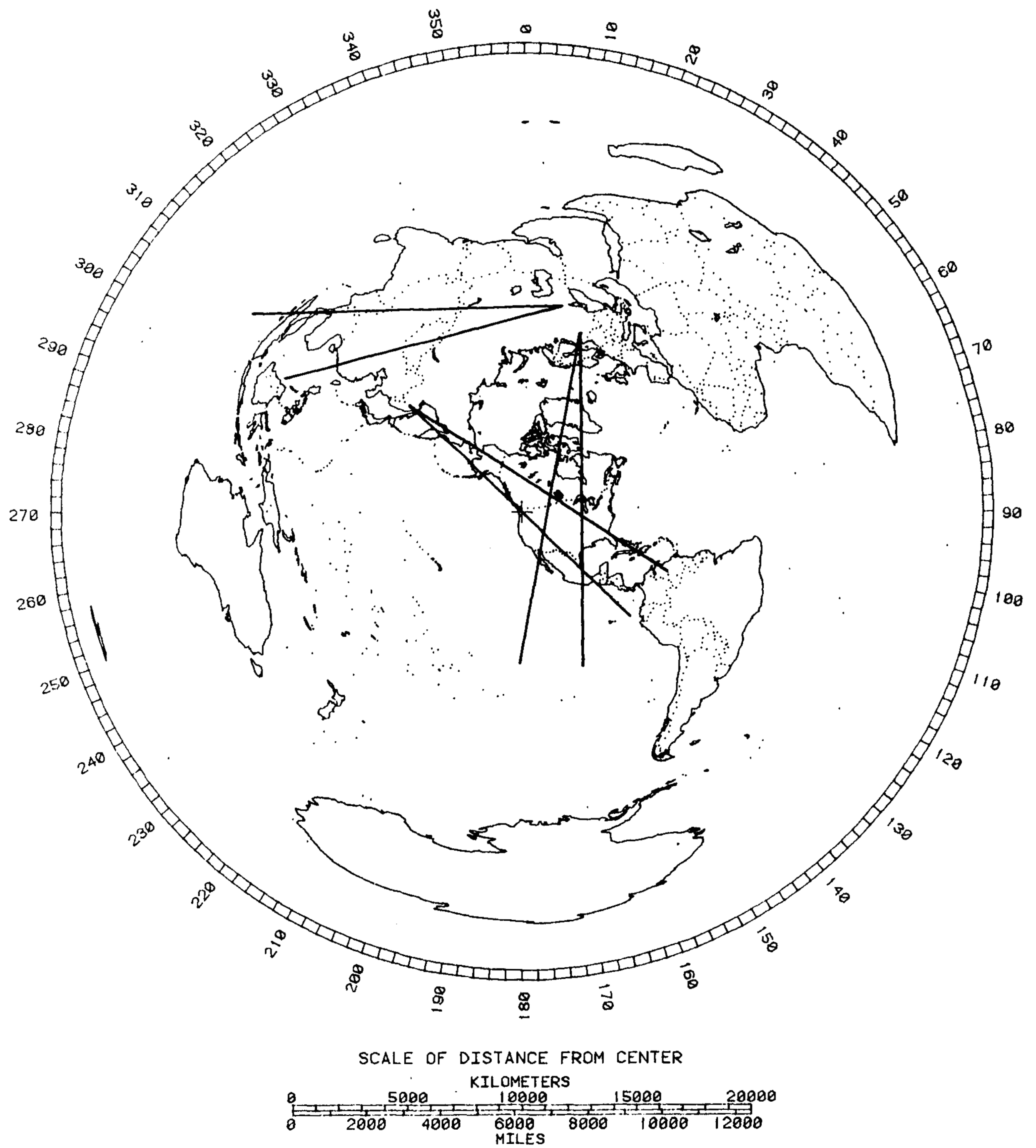

fig. 5. Coverage of Soviet "Woodpecker" ICBM launch detection systems. (Note: this base map appeared in KR7L's "Fundamentals of Grayline Propagation," ***ham radio*****, August, 1984, page 77. Azimuthal-equidistant map prepared by Bill Johnston, N5KR, 1808 Pomona Drive, Las Cruces, New Mexico 88001.)**

appear within the near future. The Soviet Union will continue operation despite world opinion, as long as it deems the practice necessary. The ultimate practical solution will be the inclusion into Amateur equipment of noise blankers capable of removing this interference. Advancing electronic technology will provide the solution to a worldwide problem that apparently cannot be resolved by diplomatic methods.

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1. Theodore J. Cohen, "CQ Interviews . . . Dr. William Schneider, K2TT, Under Secretary of State for Security Assistance, Science and Technology Department of State, Washington, D. C.," *CQ*, February, 1983, pages 11-13.

fig. 6. The MOSCOW MUFFLER® by Advanced Electronics Applications, Inc. This transceiver accessory is capable of removing interference from the Woodpecker.

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(See "Publisher's Log," April, 1984, page 6, for details.)

603-878-1441

a double conversion portable SW receiver

Modify an earlier design for additional coverage, built-in frequency counter

In the April, 1984 issue of *ham radio*, Jack described the construction of a compact, portable, high performance shortwave receiver for the 46- through 100-meter bands.[1] This article provides a design for an extended coverage receiver based on that design, but offering front-end RF tuning and a built-in frequency counter and power supply. Helpful circuit hints applicable to other receiver designs are also described. The April article should be reviewed for schematics, component values and contruction details. **Figures 1, 2, and 3** show several views of the new receiver from different angles, including component layout and shielding requirements — **Editor.**

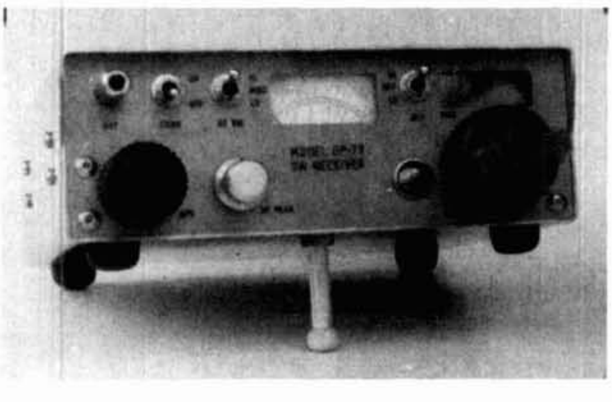

fig. 1. Top row of controls includes antenna input, on/off switch, RF bandswitching. Bottom row includes AF gain knob, RF tuning, main tuning.

Good converter design calls for an examination of *all mixing* by-products for each choice of local oscillator and desired input signal range and minimizes in-band spurious responses.[2] The frequency conversion scheme finally decided upon extends front-end coverage to include signals in the 9.3 to 10 MHz range. The incoming signal is downconverted to 3.3 to 4.0 MHz using a crystal oscillator and active mixer. The digital display is made to "track" by converting the "3" MHz readout to "9" MHz simply by switching the "F" LED segment, thereby eliminating the need for elaborate frequency readout conversion schemes. To accomplish this a 4-pole, 2-position C&K miniature switch performs the following functions:

- supplies +12 VDC to the converter board
- bandswitches the converter input
- bandswitches the converter output
- switches the "F" LED between "3" and "9"

The converter has been designed for a broadband response and the RF and MIX trimmers should be stagger-tuned for flatest front-end response. The converter schematics and the wiring of the CONV bandswitch are detailed in **fig. 4.**

construction details

In addition to the schematics and photos shown, the following information should be useful.

Power transformer. This should supply 14 volts at 120 mA. A 15-volt unit would probably be better to use because it would deliver (under load) a DC voltage closer to that of a car battery. Presently, power drain

By Jack Perolo, PY2PE1C, P.O. Box 2390, Sao Paolo, Brazil

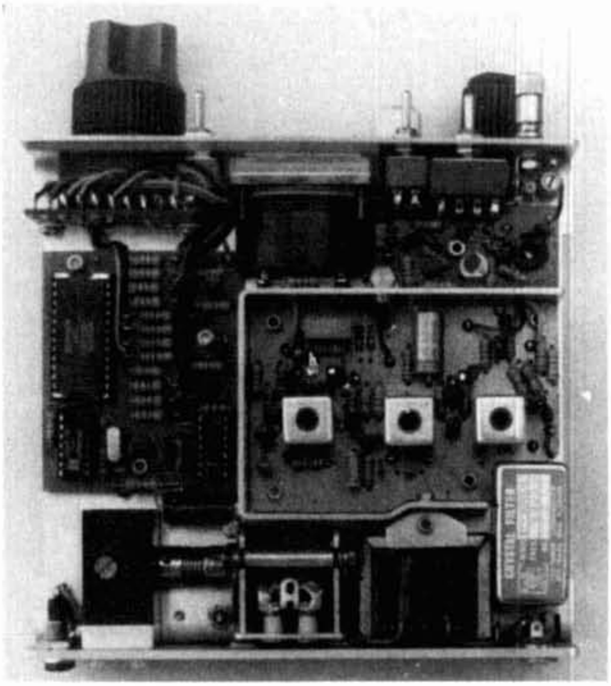

fig. 2. Top view. At top left is the 4-digit display, wired to the frequency counter below it. At top center is the S-meter, with C&K switches to its right. The input 9 MHz trap is at top right. The IF strip is below the S-meter, shielded with 1/16 inch aluminum sheet. At bottom left is the 60:1 ratio Muffett gear reducer connected with flexible coupling to the 104 pF variable capacitor. To the right of the variable capacitor is the power transformer, followed by the 9 MHz crystal filter. The back panel has provisions for two AF output jacks, DC power (12 volt) input jacks, and a 110 VAC connector (see fig. 3).

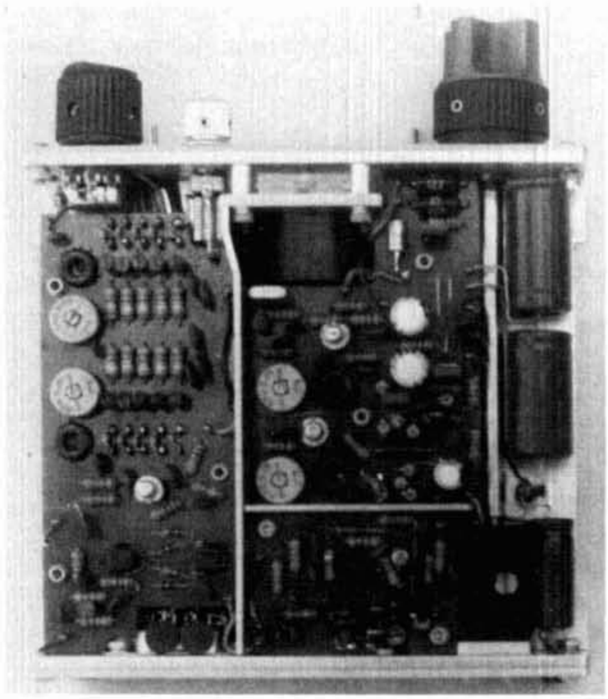

fig. 3. Bottom view. At top left are the audio frequency potentiometer and 5-24 pF Jackson RF trimmer. S-meter is at top center; below the S-meter is the PC board for power supply, AF strip and the front-end converter. The power supply electrolytic capacitors are at its right. At left is the RF/Mixer PC; notice shield between it and the front-end converter. The 110 VAC connector is at bottom. At bottom right is the VFO PC board, with electronic bandswitching circuitry, shielded from both the RF PC and front end converter PC; the VFO PC board ends near the gear reducer, seen at bottom right. All RF transistors are mounted in sockets to ease replacement in case of failure.

is 120-125 mA with the 9 MHz converter on, dropping to about 100-105 mA with the converter off. (Saw off the original brackets; use a bolted pillar and pressure holder to keep it in place.)

Space-saving techniques. In order to make room for the converter and power transformer, a new layout was developed. The audio strip PC board was redesigned and reduced in size, with all components vertically mounted. This reduced it from 1.57 x 2 inches (40 x 50 mm) to 1.18 x 1.57 inches (30 x 40 mm).

The PC board that houses the audio strip also includes the power supply, the zener diodes, and the front-end converter. The 1/16 inch thick epoxy board measures 2-3/8 x 3-1/2 inches (60 x 90 mm). Separate diodes were used in the supply to avoid confusion in case 110 VAC and external power were left on at the same time.

Gear reducer. Zero backlash, Muffett size 1 with gear ratio 60:1. Available in the U.K. for $75.00.

Cabinet. 2 x 6 x 6 inches (50 x 150 x 150 mm) HWD.

ham band options

The 80-meter band, covered in the earlier project, is included in the current version. The 40-meter band can be covered by the direct method or by the converter method. Using the direct method extends the coverage of the basic receiver to 7 MHz. (The MHz digit of the display must be read.) The converter option would employ a 4 MHz crystal; the digit problem appears to be easier to solve, but some spurious signals are likely to appear within the band. Coverage on 30 and 20 meters can be implemented by modifying the front-end conversion using a single set of coils and electronically bandswitching the parallel capacitors and the oscillator crystal. One can cover 7, 10, and 14 MHz with the same basic converter by increasing PC board size slightly.

The frequency counter can easily handle the fifth digit (tens of MHz) because the 7207 IC has provisions for it, but the power supply must be sized accordingly for the extra load. Care should also be taken in the layout and design of the front panel, which is presently

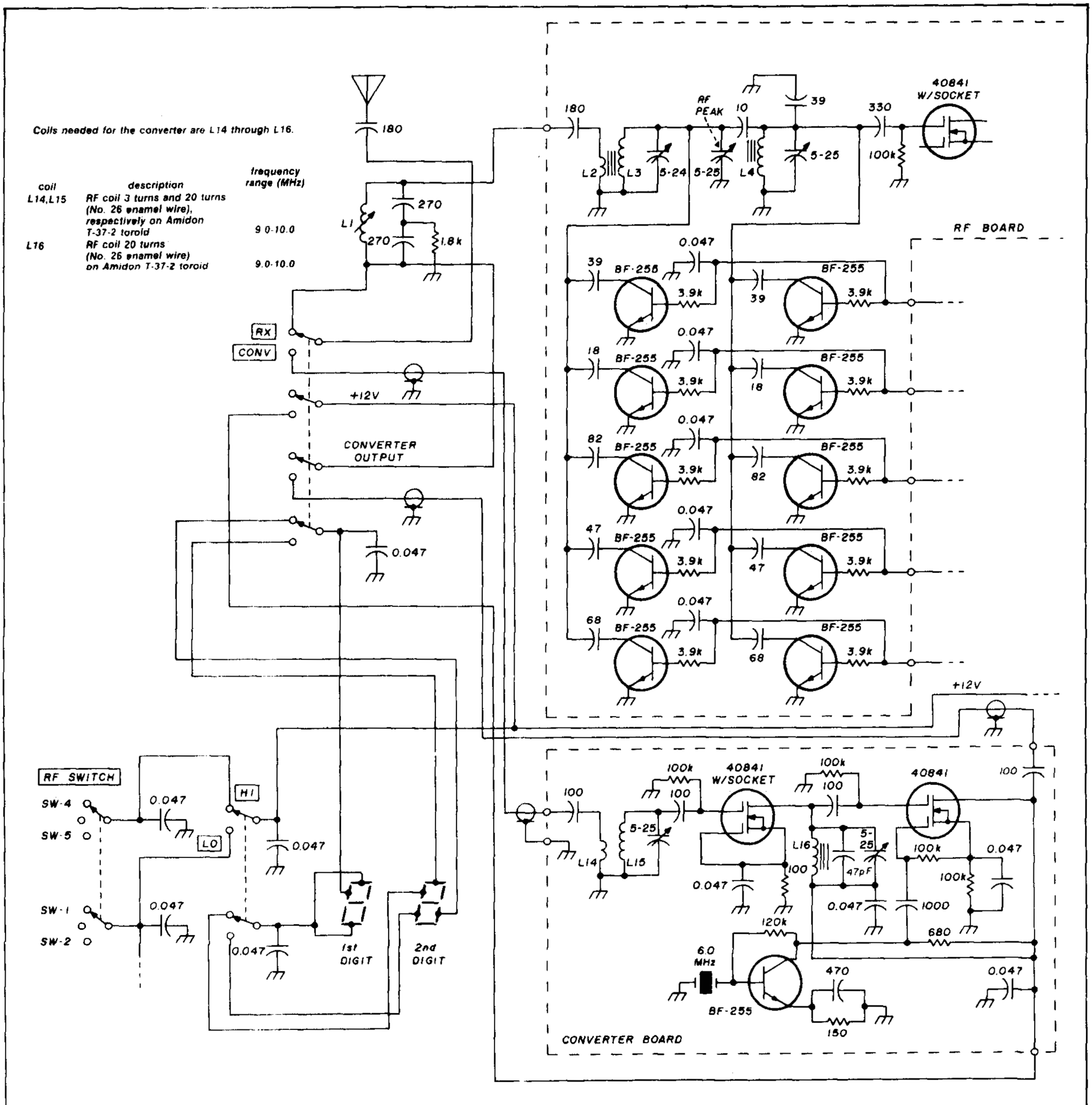

fig. 4. Schematics of the GP-78 converter and bandswitch wiring. (RF coils shown, L1 through L4, are detailed in reference 1.) Shielded bandswitch connections use miniature coaxial cable, type RG-174/U or equivalent. The 6.0 MHz crystal is a type HC-25/U by ICM. Only the first two digits (DG) of the frequency display are shown, as the third and fourth digits are wired in parallel to digit No. 2. The new control added "RF PEAK" is a 5-24 pF Jackson miniature trimmer, controllable from the front panel (see fig. 1). The remainder of this circuit (not shown) is identical to the schematics shown in reference 1. The converter shown could also be adapted to the earlier (or other) projects as a separate accessory, fitting into a 3 × 3 × 0.75 inch box, borrowing the +12 volts from the receiver.

very crowded; the addition of an extra digit would require widening the panel.

acknowledgement

Thanks go to Fernando, PY2DQU, for his support and encouragement on this project.

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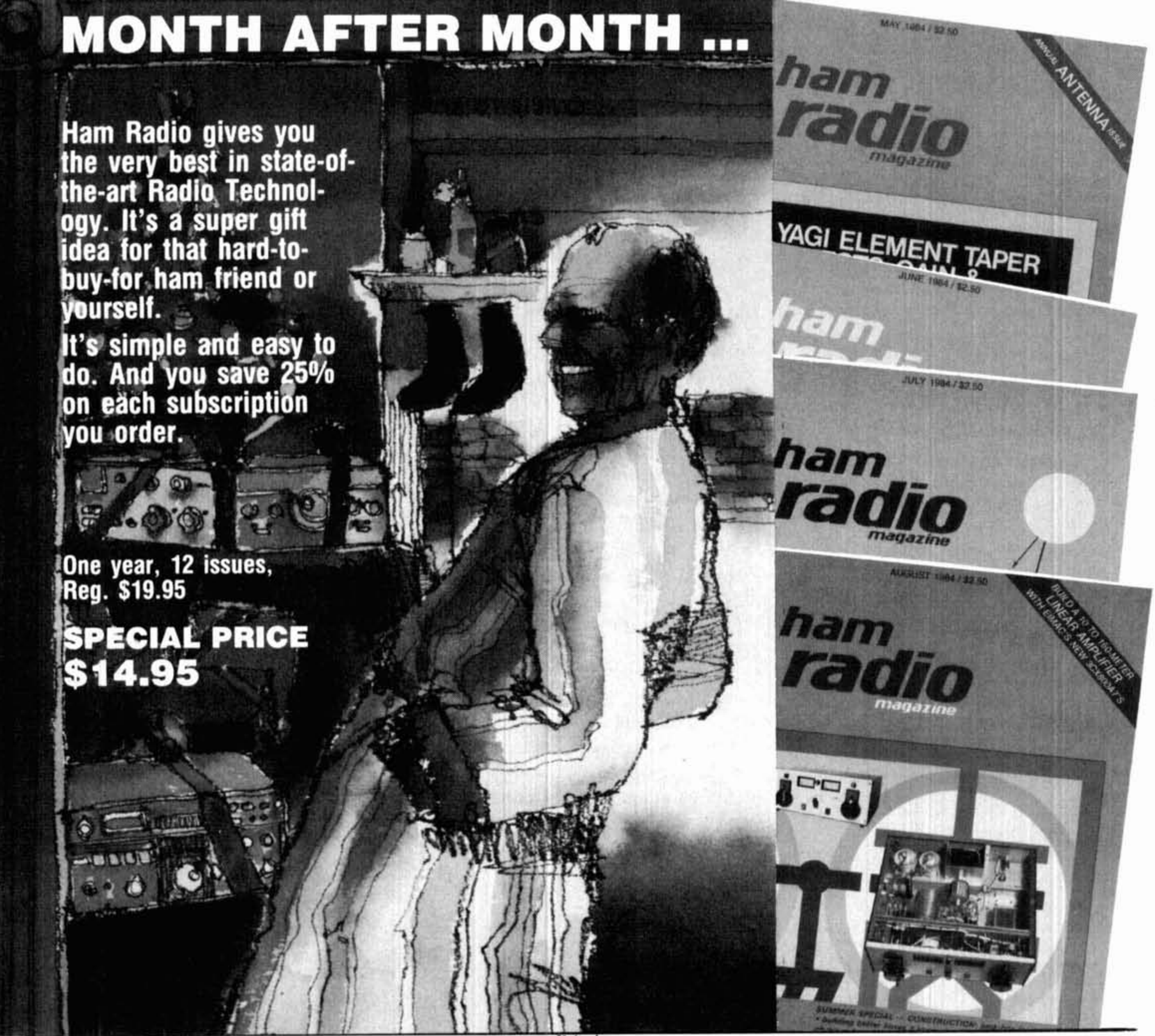

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THE GUERRI REPORT

Ernie Guerri
W6MGI

computer technology: fast, *fast*, *FAST!*

Some of the most dramatic changes in computer technology are taking place in architecture — that is, *in the way computer logic is organized.* This is necessary because electronic speeds are now so fast that the physical distance between circuit elements has become a major limitation. To alleviate this problem, the Cray supercomputer features logic bays arranged in a circle so that all interconnects are not more than one circle diameter away.

NASA has developed a design called the Massively Parallel Computer — and massive it is: over 16,000 identical processors are arranged in parallel. This approach allows an image processing task that takes 8 hours on a conventional large mainframe to be reduced to only 17 minutes! Companies such as Cray, ETA, and Fujitsu are developing computers that will be 10,000 times faster than an IBM PC by 1987 or 1988. The implications of this kind of progress make exciting news for hams. Might it someday be possible to contain a basic HF receiver on a single chip? A complete SSB receiver could actually consist of 3000 individual receivers, each having 1 kHz bandwidth and tuned to a different part of the HF spectrum under computer control. Such a unit could represent the ultimate in interference avoidance and MUF agility!

cooling high-speed circuits

Ever since the beginning of the electronic era, heat has been a problem. The absence of effective ways to remove it at the device level continues to be a major limitation to present large-scale integration. Designers are now examining methods by which an IC substrate can be bonded to a porous metal carrier, with coolant circulated through the porous metal, then evaporated and recovered in a closed system. Using this method, thermal transfer can be improved to a rate 100 times better than with conduction alone, with each LSI circuit containing its own refrigeration system.

Also significant is the interest in running VLSI circuits at the temperature of liquid nitrogen (77 degrees K). At this temperature electron mobilities go up, speed increases, and thermal efficiency improves. With modern techniques, even the cryogenic problems aren't too difficult. Look for examples of this approach to appear in commercial products before too long.

Even more exotic are super-cold devices called Josephson junctions (JJ's). These are thin film devices that operate at 4 degrees K (the temperature of liquid hydrogen), exhibit picosecond (one millionth of a microsecond) speed, and consume nearly zero power because they operate at superconducting temperatures. After spending nearly twenty years developing JJ's for supercomputer applications, IBM recently threw in the towel because of the difficulty of manufacturing the device and its support system. Work continues in Japan, however, with Fujitsu pursuing research and the Ministry of International Trade and Industry (MITI) funding the development of a JJ analog-to-digital converter. With several GHz bandwidths, such a device could make possible digital storage scopes with several hundred MHz capability, or low-noise digital VHF receivers with direct conversion to digital information at the front end. Although very low temperature devices have many attractive characteristics, they may be difficult to put to use —except in space, where the necessary low temperature is free. But the benefits are great enough to warrant a substantial continued effort around the world. We should see some exciting breakthroughs in the near future.

faster antennas, too

Take a look at what's happening in telecommunications. More and more information is being sent over each circuit; system bandwidths are being increased, and most data is now digitized before transmission. All this wideband data eventually goes to an antenna that radiates the signal. This means that the antenna has to have some measure of frequency independence — that is, be broadband. As data rates and information density increase, the pulse response of the system also becomes a consideration.

It is now being observed that many antenna types don't exhibit adequate pulse response for present and contemplated data links. The problem is not an easy one to resolve. In order to radiate, antennas must depart from the distributed characteristics that give transmission lines their good pulse response. Much attention is now being given to measuring the pulse/transient response of various antennas, and the relevant journals abound with complex math as a result. Perhaps all this effort will lead to new antennas that will couple the desired energy into space without acting as if they had all kinds of L & C hung across them.

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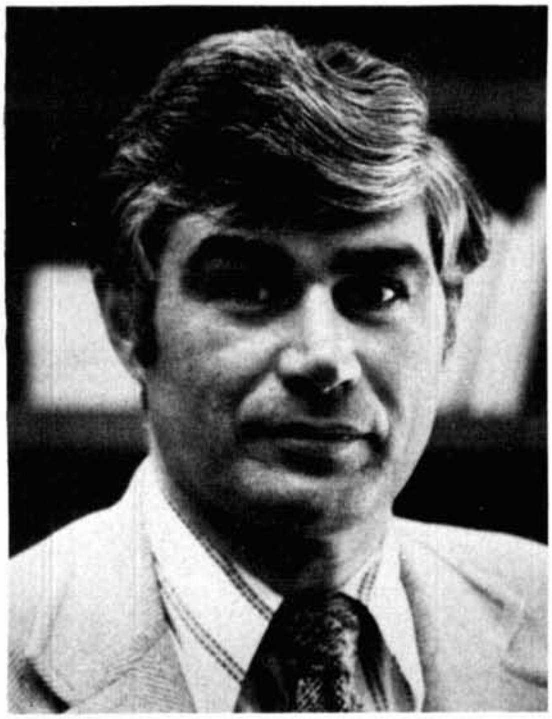

meet Ernie Guerri, W6MGI

Ernie Guerri, W6MGI, comes to the pages of *ham radio* with a background that includes 32 years as a licensed Amateur, and 27 years in the development and management of advanced technology. He is a Senior Member of the IEEE and a life member of ARRL.

Ernie was educated in Physics at the University of Maine, Semiconductor Electronics at the University of California, and in Business at Stanford. He has held engineering and management positions at IBM, Raytheon, and General Dynamics. Each of these positions involved work at the leading edge of aerospace technology, telementry, and deep space communications. Most recently he has been President and General Manager of the Advanced Technology Center of Gould Inc., one of America's large ($1.5B) electronics companies. In October, 1983 he left Gould to form his own technology consulting firm, which he now operates from offices in Chicago.

Ernie will be commenting on technological developments that will shape the equipment of tomorrow. Some will have direct relationship to Amateur Radio; others will hopefully encourage implementation of new technologies in yet unexplored areas of our hobby.

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extending the modular 2-band receiver

Add two boards — and build yourself an HF transceiver for SSB and CW

In the July, 1983, issue of *ham radio* I presented a design for a practical, easy-to-use HF receiver with digital readout.[1] The simple addition of a converter to the front end extends the receiver's frequency coverage to VHF; the addition of just two more boards turns the unit into an SSB/CW transceiver as well, see **fig. 1**.

design concept

Simply stated, an SSB transmitter amplifies voice, mixes it with a carrier frequency in such a way as to balance out the carrier, removes one sideband, and mixes the result up or down in frequency to the desired output frequency.

In this design, see **fig. 2**, the audio from the microphone is amplified by a two-stage speech amplifier and applied to a simple two-diode balanced modulator that removes the carrier, providing a signal that contains two sidebands and a suppressed carrier. The carrier source is the BFO. By selecting either USB or LSB, the operating mode for the transmit signal is also selected. To remove one of the sidebands, the signal is passed through the same crystal filter used for receive. Just as the unwanted sideband is removed on receive, the output from the crystal filter contains only one sideband. Because this sideband signal, however, is too low in level to allow the transmitting mixer to function properly, an IF amplifier must be used to increase the signal to an effective level. The output of this stage is injected into the MC1496 double-balanced mixer IC, where it again mixes with the VFO to produce outputs at 14 and 4 MHz (9 + 5 = 14 and 9 - 5 = 4). The same filters as those used for receive are used here to remove the undesired output. The 20-meter filter removes the 80-meter signal and vice-versa.

The SSB signal present at the output of the band-pass filter is clean but at a very low level. A two-stage broadband amplifier has been designed for an output of about 10 watts (see **fig. 3**). The driver transistor is a 2N3866 which in turn drives a 2N5590 operating in class AB. The output at this power level is "clean" (low spurious/harmonic content) and requires no additional filtering. However, if you wish to drive a much higher output broadband amplifier, I would recommend adding low-pass filters for each band. Several articles on this subject have been published in this and other magazines.

operation

The same mixing scheme used for SSB transmission can be used to generate CW. A twin-T oscillator serves the dual purpose of generating both a sine wave tone, used for monitoring, and the CW signal. When a clean tone of a single frequency is applied to an SSB transmitter, a single output frequency, separated from the removed carrier frequency by the frequency of the applied tone, is produced from the transmitter. For example, if a 1 kHz tone is injected into the SSB transmitter, a CW output offset by 1 kHz is generated. Conversely, if tuning in another station produces a 1 kHz CW tone on receive, your transmitter will be on the exact transmitting frequency of the other station when you answer. (A similar method was used to produce CW in the old Heathkit SB/HW series of transceivers.)

To send CW it's necessary to activate the transmitter by either manually switching to transmit or, more easily, using the included 555 timer circuit. This keeps the transmitter on between the dots and dashes. The twin-T oscillator and the timer circuit are keyed at the same time; only the 555 timer is keyed in SSB. This timer stage switches all stages into transmit for a period determined by the adjustable time constants. In addition, the AGC for the IF amplifier must be disabled while in the transmit mode by grounding the AGC control pin 5 on the MC1350, through a 4.7 kilohm resistor.

By **James J. Forkin, WA3TFS**, 3210 Shadyway Drive, Pittsburgh, Pennsylvania 15227

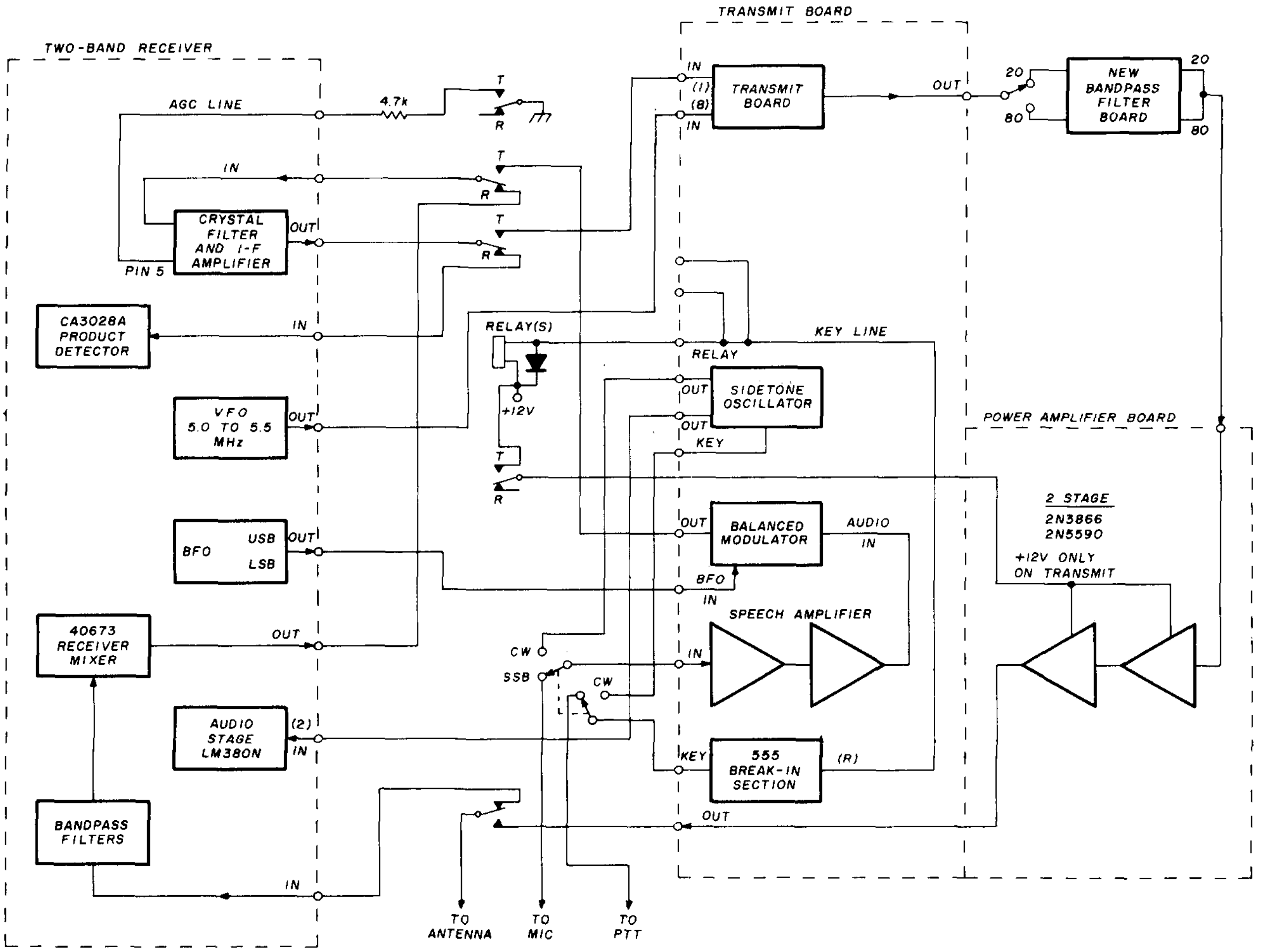


fig. 1. The interconnection diagram. Circuits within the dotted lines are on either the receive board on the left or the transmit board on the right. If more than one relay is used, wire the coil in parallel with the one shown. Use a protection diode as shown on each. Use shielded wiring on all audio circuits and 50-ohm coax on the transmit/receive relay wiring. The new bandpass filter board is identical to the board described in the July, 1983, *ham radio* receiver article.

construction

The transmit modification is accomplished through the addition of two PC boards. One consists of the two-stage amplifier described above. (Component layout and printed circuit board artwork are detailed in **figs. 4**, and **5**, respectively.) The other board, however, is the actual transmit conversion. Shown on the board, (component layout and printed circuit board artwork are detailed in **figs. 6** and **7**),from left to right, is the 1 kHz sidetone oscillator coupled through a panel switch to the speech amplifier stage in the CW mode. Next is the 555 timer stage, which holds the rig in transmit for a period of time adjusted by the trimmer pot. To the right of the timer is the transistor, used to switch the relays used in the various stages of the receiver. Next in line is the two-stage speech amplifier; note that the two stages are coupled by a jumper wire to simplify the addition of audio companders, processors, or other components later. The balanced modulator is next. Be careful to wind the toroid core exactly as shown, keeping all leads as short as possible. It is this stage that determines the ultimate level/amount of carrier suppression the transceiver will offer. The double-balanced mixer completes the board.

switching

It is important to switch the crystal filter and IF amplifier stage when going from receive to transmit. A single-pole, double-throw miniature relay mounted close to the input and output of this stage does this. Use shielded wire to and from the transmit board. (RG-174 miniature coax works well.) The front end filters must also be modified by adding two relays or alternatively, replaced with new filters exactly like those used for receive for the transmitter, thus eliminating the need for relay switching here. The only

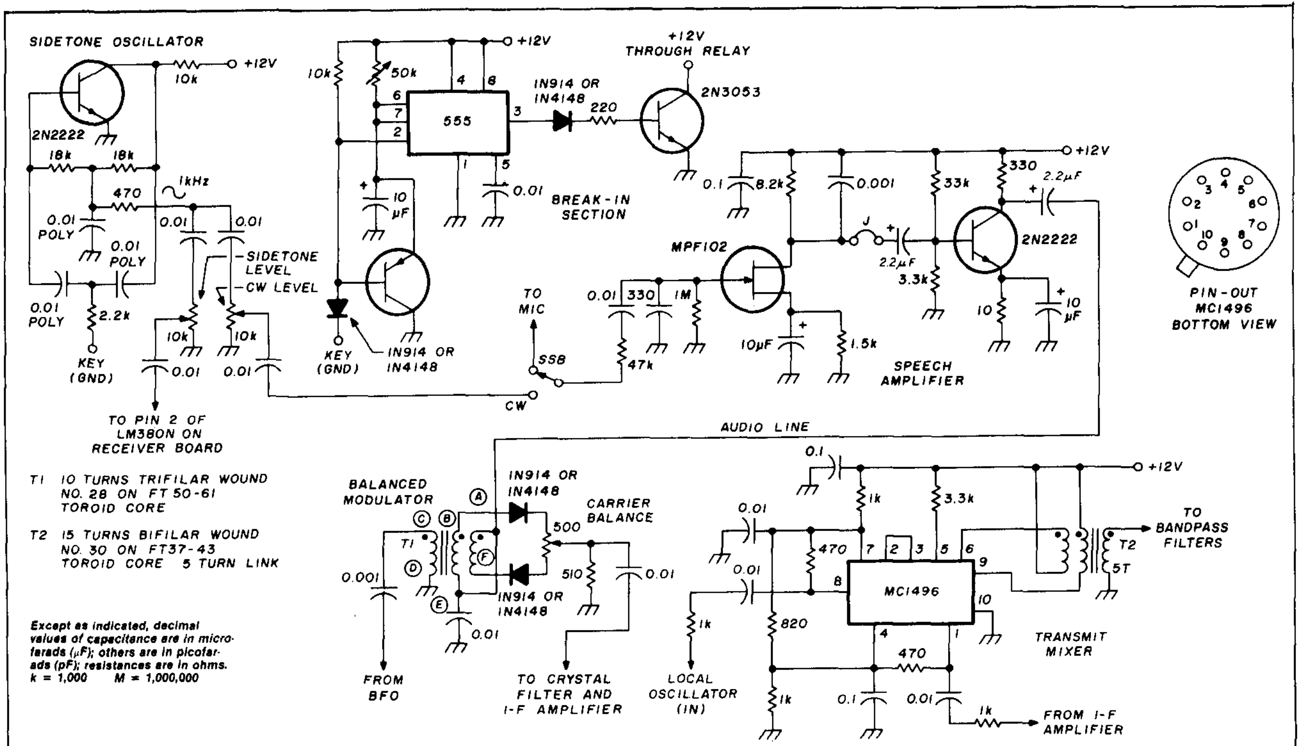


fig. 2. The schematic diagram of the transmitter board.

disadvantage to the latter approach is the need to allow for weight and additional space; despite these disadvantages, replacing the filters rather than adding relays does simplify the modification. Use coax for the filters — keep the RF where it belongs! And don't forget that to pull the AGC voltage below 5 volts for maximum gain, you'll have to use another relay or add an extra set of contacts to one of the other relays to ground pin 5 of the IF amplifier through a 4.7 kilohm resistor. Should you choose to use diode switching instead of relays, you may wish to consult several articles that have been published on the subject for help with the design.

For simplicity's sake, you may decide, as I did, to use relays. Several types of 5- and 12-volt DC relays are available on the surplus market. Use 12-volt relays if you can find them at a reasonable price, or wire three 5-volt relays in series; they'll key reliably on 12 volts. Mount your relays to the board with double-sided foam tape or glue. Place a diode across each relay coil to prevent voltage spikes.

initial adjustments

After completing the modification, make sure that the receiver still works. Realign it and check the BFO frequencies. When you're convinced that the receiver is working as well as it did before the modification, connect a dummy load to the antenna terminal and key the transmitter in the CW position. Check that all

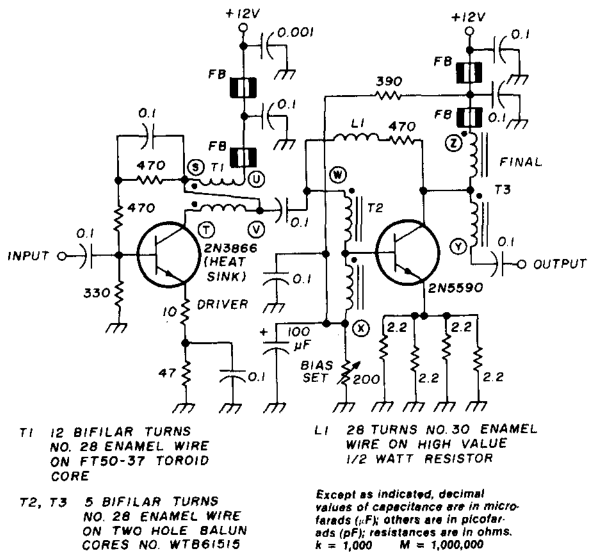
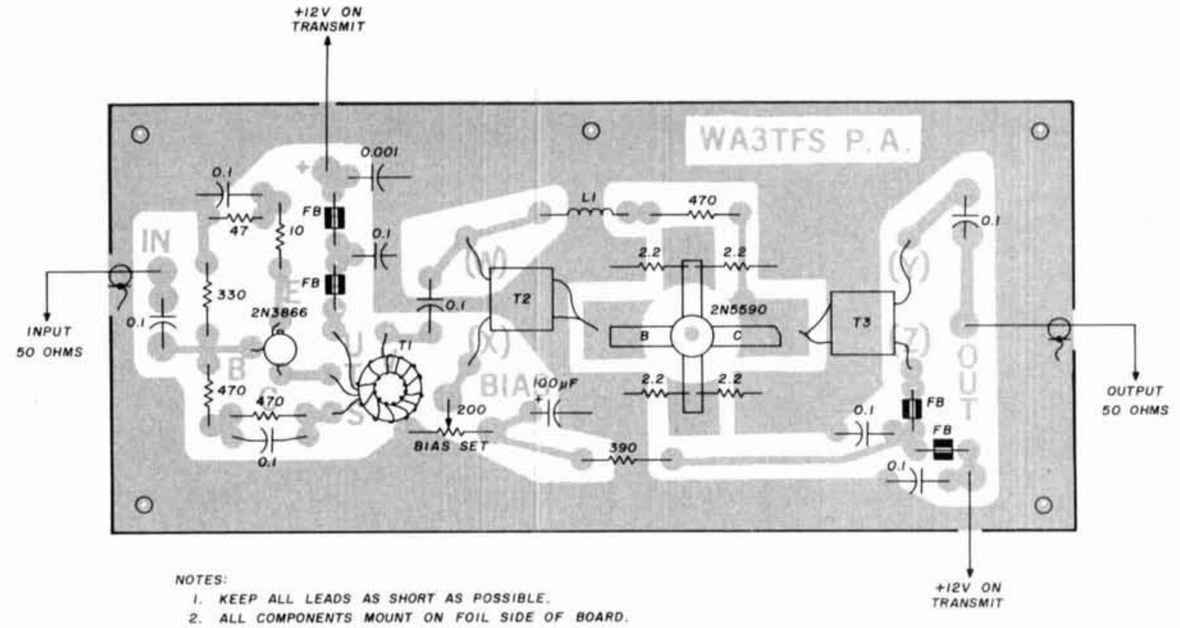


fig. 3. The schematic diagram for the final amplifier board. The driver section is similar to the VFO buffer in the receiver and the final stage is patterned after the original design.

relays switch as they should and adjust the 555 for a hold-in time of about 1 second. Check that the AGC voltage at pin 5 of the MC1350 is in fact dropping below 5 volts on transmit. Put the rig into SSB and

✓ 220

✓ 109



NOTES:
1. KEEP ALL LEADS AS SHORT AS POSSIBLE.
2. ALL COMPONENTS MOUNT ON FOIL SIDE OF BOARD.
3. MOUNT 2N5590 FIRST TO HEATSINK AND THEN SPACE BOARD ABOVE HEATSINK WITH WASHERS SO THERE IS NO STRAIN ON TRANSISTOR.
4. USE 50 OHM COAX FOR INPUT/OUTPUT.
5. ADJUST BIAS FOR 75-80mA WITH NO SIGNAL ON 2N5590.

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

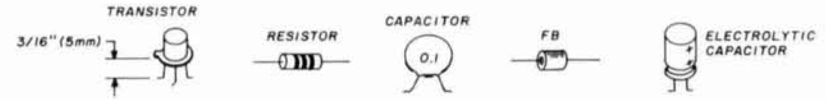


fig. 4. The parts layout for the RF amplifier board. Positive 12 volts is applied only on transmit. Set bias adjustment for an idle current in the final stage of 75 to 80 mA with no drive applied to the board.

adjust the output stage for a resting current of 80 mA. With the rig in SSB there will be no drive to the final stage. Put the switch back in CW. When keying, adjust the trim pot in the oscillator stage for sidetone level. Increase the drive level with the other trim pot until no increases in output level from the transmitter are noted. Back off the adjustment slightly. CW tune-up is completed.

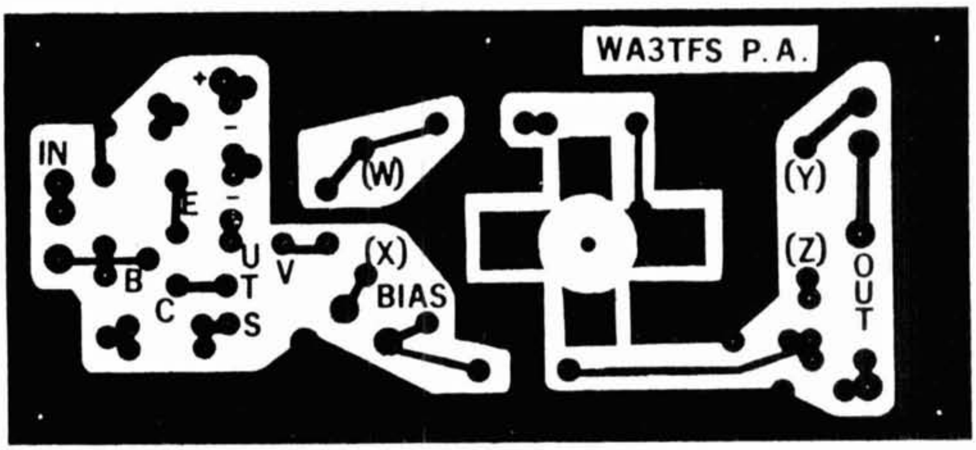
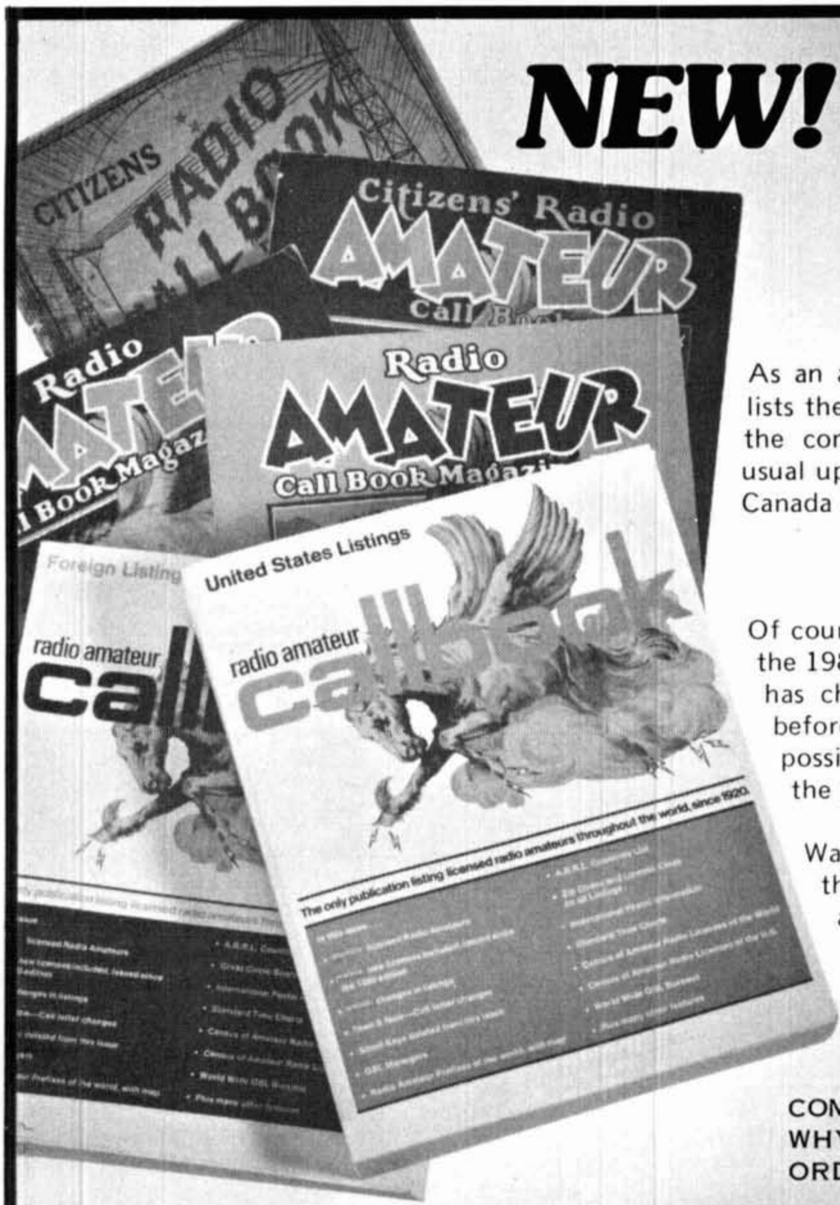


fig. 5. The foil side layout for the RF amplifier board. All parts on this board are mounted on the foil side.



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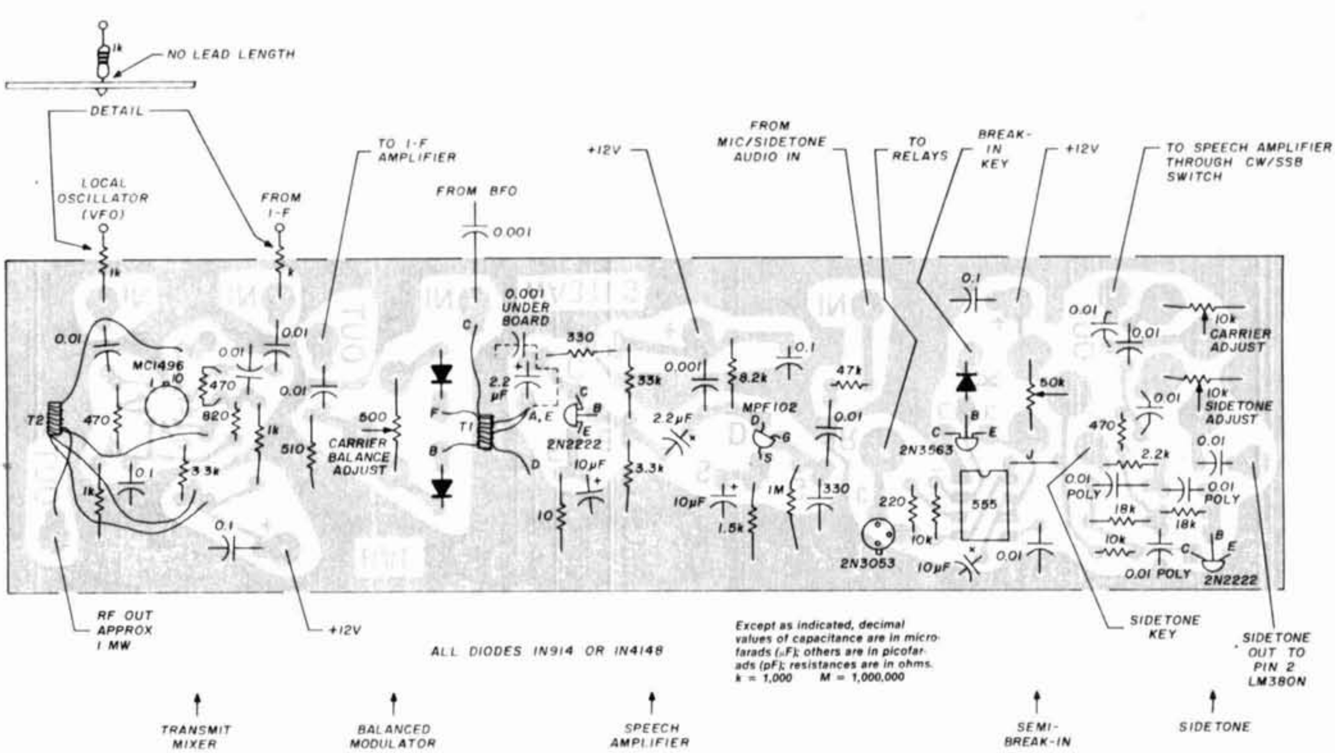

fig. 6. The parts layout for the transmit board. (Note: On this board components are *not* mounted on the foil side.) Use shielded wire for all audio connections, and 50-ohm coax for all RF wiring.

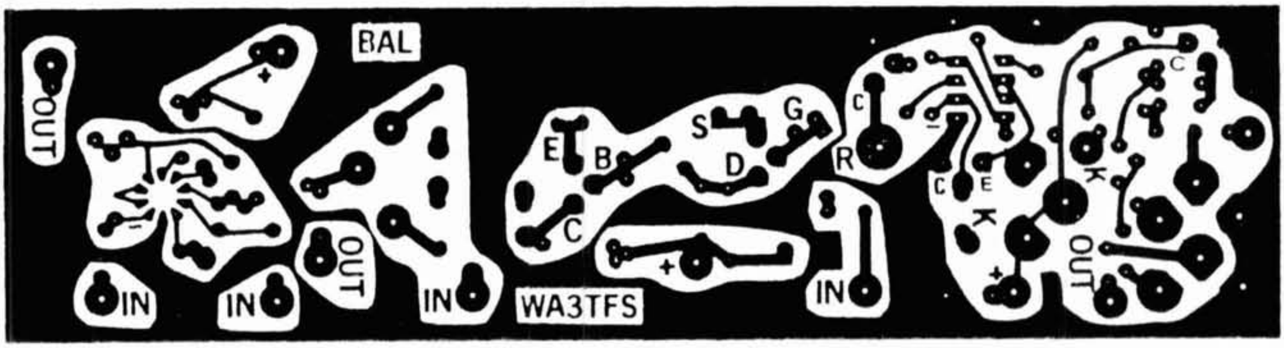

fig. 7. The foil side layout for the transmitter board.

Next attach a microphone to the input. (A CB replacement microphone will be sufficient.) If the CW/SSB switch has been wired correctly, the microphone will be connected to the input of the speech amplifier and the twin-T oscillator will be disconnected. When you press the PTT and whistle into the microphone, the RF output should increase. The level should be about the same as when transmitting CW, but may vary because of different output levels of various microphone elements.

Disconnect the microphone and while still in SSB, key the transmitter with a jumper wire. While checking the output with a meter, or better still, an oscilloscope, adjust the trim pot in the balanced mixer for minimum output and, consequently, maximum carrier suppression. If you can't see any change in the meter reading as you make this adjustment, you'll know you've either wound the coil in the balanced modulator incorrectly or caused some stage to oscillate because of poor wiring layout or failure to ground something. Check your construction step by step. You should see a definite null in output power. If everything appears to be in order, go back and readjust everything until no further change occurs. Your new transceiver is now ready to be connected to the antenna.

conclusion

This complete transceiver will operate reliably and efficiently as long as care is taken to attach a matched 50-ohm load. The output stage will not self-destruct if you have high SWR or forget to attach the antenna, but output power will be low. The rig should run about 8 to 10 watts out into a matched antenna. I have worked all states on 20-meter sideband and find I require no more power from the home station.

A kit is available from the author to make the modification described in this article. For details, please send an SASE to me at the address shown at the beginning of this article.

ham radio

receiving signals from space

How to locate geostationary satellites from your QTH

By Dennis Mitchell, K8UR, 1 Cider Mill Lane, Upton, Massachusetts 01568

With the price of TV-Receive Only (TVRO) terminals on its way down and the availability of channels expanding, interest in geostationary satellites is increasing. This article describes how to locate these satellites from any given latitude and longitude in terms of azimuth, elevation, and range.

Two programs are included — one in BASIC for the TRS-80™ (level II or similar) and the other for the Hewlett-Packard 67 or equivalent. While the mathematics are the same for each program, some minor changes have been made to accommodate the specific programming language used and the functions available on each machine. For example, (cos/sin) is substituted for (cot) because of the absence of the cotangent function on the HP-67 and in TRS-80 Level II BASIC.

celestial mechanics

For a satellite to always appear stationary above a particular point on earth, it must have the same **period** as the Earth — that is, 23 hours, 56 minutes, 4.09 seconds or 86164.09 seconds. In order to have a period that matches that of the earth, the geostationary satellite must be a specific height above earth. This measurement can be found by using the Newtonian *law stating that the square of the velocity of an object* (satellite in this case) is equal to the universal gravitational constant times the mass attracting the object (the Earth), divided by the distance of the object from the center of the mass (Earth).

$$V^2 = \frac{GM_E}{D} \quad (1)$$

where V = velocity of satellite
G = universal gravitational constant $6.673 \times 10^{-11} \frac{\text{Newtons-meter}^2}{\text{kilogram}^2}$
M_E = Mass of Earth = 5.975×10^{24} kg
D = distance from center of earth to satellite
= (R + H) = radius of earth + height of satellite above earth

G and M_E are constants and can be combined:

$$G' = GM_E = 3.987 \times 10^{14}\ meters^3/sec^2$$
$$= 3.987 \times 10^5\ km^3/sec^2 \quad (2)$$

which results in

$$V^2 = \frac{G'}{R + H} = \frac{G'}{D} \quad (3)$$

$$\text{or } V_{SAT} = \sqrt{\frac{G'}{D}} \quad (4)$$

The period of one complete revolution of the satellite is equal to the distance it travels in orbit divided by its velocity or:

$$P_{SAT} = \frac{2\pi D}{V_{SAT}} = \frac{2\pi D}{\sqrt{\frac{G'}{D}}} = 2\pi \sqrt{D^3/G'} \quad (5)$$

But this is equal to 86,164 seconds (approximately 24 hours) for it to be a geo*stationary* satellite as explained above.

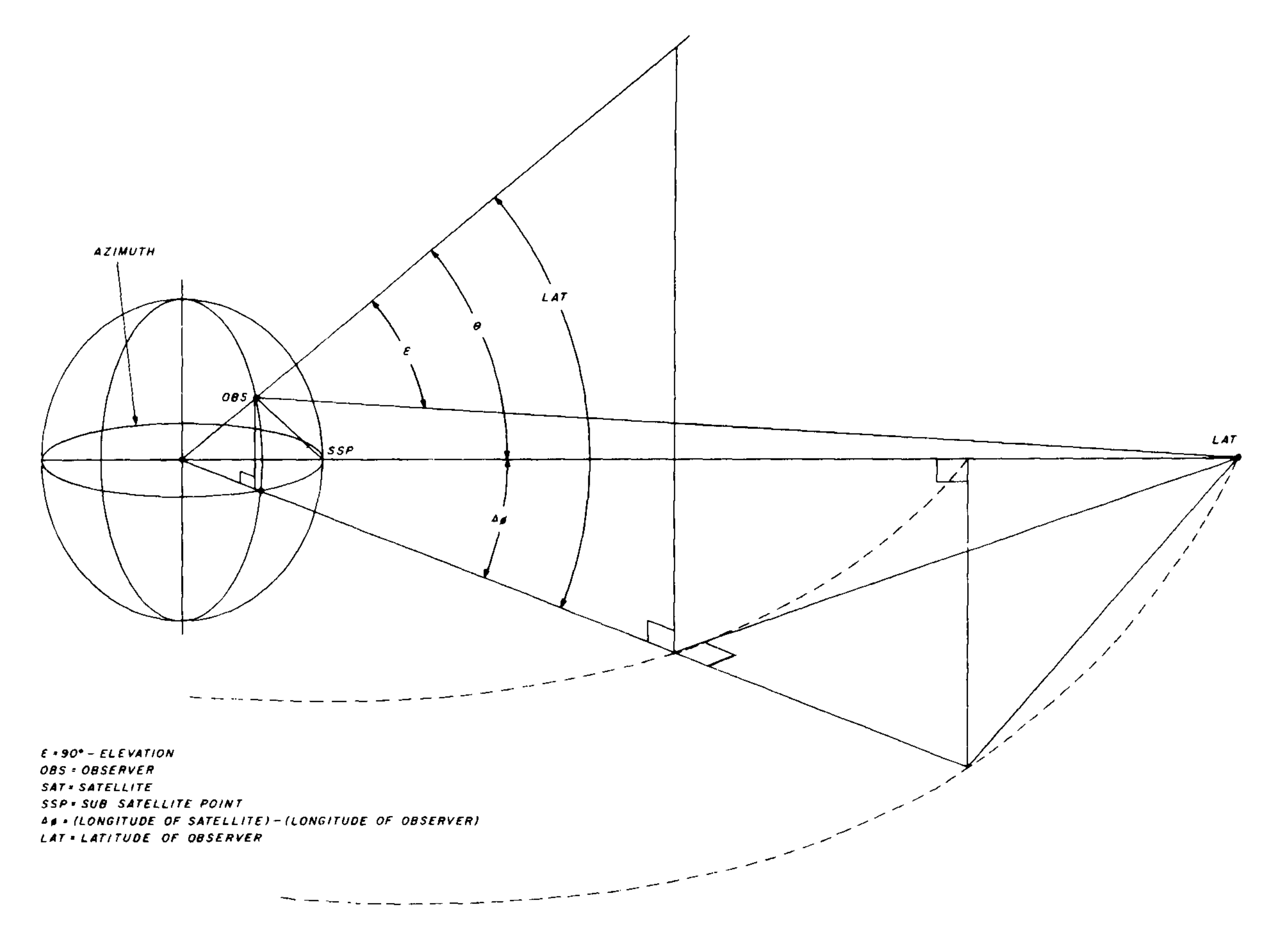

fig. 1. Angles involved in calculating satellite's azimuth, elevation and range.

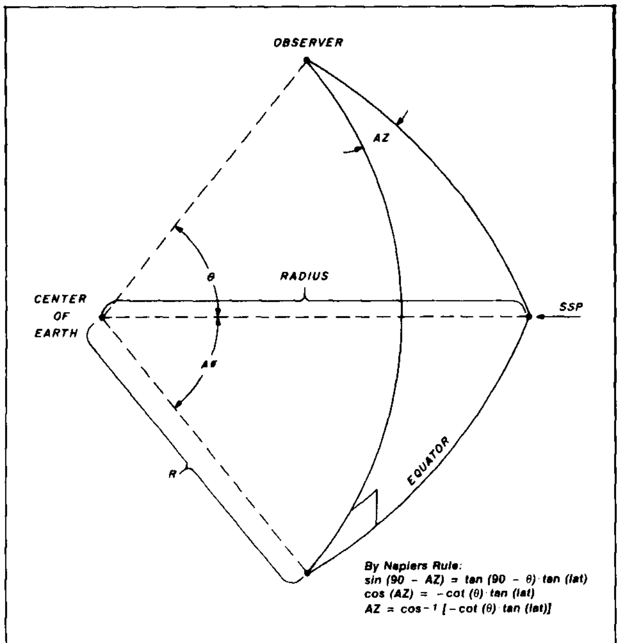

fig. 2. Napier's rule illustrates the relationship between the satellite and the observer.

Rearranging terms and solving for D:

$$D = \sqrt[3]{\left(\frac{P_{SAT}}{2\pi}\right)^2 G'} \qquad \textbf{(6)}$$

$$= 42,168 \ km$$

$$H \text{ (height of satellite above earth)} = 42,168 - 6378$$
$$= 35,790 \ km$$
$$= 22,239 \ miles$$

$$\text{and } V_{SAT} = \sqrt{\frac{G'}{D}} = \frac{398,700}{42,168} = 3.075 \ km/sec$$
$$= 3075 \ meters/sec.$$

finding azimuth, elevation, and range

Fig. 1 shows the angles involved in finding the azimuth, elevation, and range of the satellite.

Fig. 2, an exploded view of a section of the Earth, shows how the locations of the Earth's center, an observer, the Equator, and latitude are related. The difference in longitude and the sub-satellite point are also shown.

By viewing **figs. 1** and **2** and using Napier's rule with

table 1. List of the current C-band (3.7-4.2 GHz) geosynchronous satellites and their longitude.

satellite name	degrees west
AURORA I	143
ANIK B	109
ANIK C2	105
ANIK D	104.5
ANIK III	114
COMSTAR I	128
COMSTAR II	95
COMSTAR III	87
COMSTAR IV	127
GALAXY I	128
GALAXY II	74
SATCOM I-R	139
SATCOM II-R	72
SATCOM III-R	131
SATCOM II	119
SATCOM IV	83
SBS I	100
SBS II	97
SBS III	95
TELSTAR 301	96
USAT I	85
WESTAR I	99
WESTAR II	79
WESTAR III	91
WESTAR IV	99
WESTAR V	123

table 2. Sample calculation of geosynchronous satellite azimuth, elevation, and range for an observer in Upton, Massachusetts.

satellite	longitude	azimuth	elevation	range (km)
AURORA I	143	257	5	41128
ANIK B	109	228	28	38763
ANIK C2	105	224	30	38555
ANIK D	104.5	223	30	38530
ANIK III	114	233	25	39049
COMSTAR I	128	245	15	39979
COMSTAR II	95	212	35	38125
COMSTAR III	87	202	38	37883
COMSTAR IV	127	244	16	39907
GALAXY I	128	245	15	39979
GALAXY II	74	183	41	37700
SATCOM I-R	139	254	7	40811
SATCOM II-R	72	180	41	37695
SATCOM III-R	131	248	13	40199
SATCOM IV	83	196	39	37799
SBS I	100	218	33	38324
SBS II	97	215	34	38201
SBS III	95	212	35	38125
TELSTAR 301	96	213	35	38162
USAT I	85	199	39	37838
WESTAR II	79	190	40	37739
WESTAR III	91	207	37	37993
WESTAR IV	99	217	33	38281
WESTAR V	123	241	19	39627

```
10 CLEAR 100
20 REM**********************************************************
30 REM***  THIS PROGRAM WILL COMPUTE THE AZIMUTH, ELEVATION ***
40 REM***  AND RANGE TO GEOSTATIONARY SATELLITES FOR TVRO   ***
50 REM********* BY D.C.MITCHELL-K8UR/1 1/1/81 ****************
52 REM********* UPDATED SATELLITE LIST 6/27/84 ***************
55 REM******** 1 CIDER MILL LN. UPTON,MA.01568 ***************
60 REM**********************************************************
70 CLS
80 PRINTCHR$(23):PRINT"ENTER YOUR LATITUDE":GOSUB 100:LA=DD
90 CLS:PRINTCHR$(23):PRINT"ENTER YOUR LONGITUDE":GOSUB 100:LO=DD:GOTO 150
100 PRINT@260,"DEGREES-----":PRINT@324,"MINUTES-----":PRINT@388,"SECONDS-----"
110 PRINT@286,"";:INPUTDD:PRINT@324,"MINUTES----- ";:INPUTMM:PRINT@388,"SECONDS----- ";:INPUTSS
120 PI=3.14159
130 LA$=STR$(DD)+STR$(MM)+STR$(SS):DD=(MM*60+SS)/3600+DD:RETURN
140 REM
150 INPUT"ENTER YOUR CITY";C$:INPUT"ENTER YOUR STATE";T$:C$=C$+","+T$
160 CLS
170 LA=PI*LA/180:LO=PI*LO/180
180 PRINT"    SATELLITE ***  AZIMUTH / ELEVATION / RANGE *** LOCATOR"
190 PRINT"                       FOR   "C$:PRINT""
200 PRINT"SATELLITE    LONG      AZIMUTH       ELEVATION     RANGE (KM)"
210 PRINTSTRING$(60,"*")
220 READ S$,S
230 IF S$="END" THEN END
240 S=S*PI/180
250 DEFFN AC=-ATN(X/SQR(-X*X+1))+1.5708
260 X=(COS(LA)*COS(S-LO)):TH=FNAC
270 X=(-TAN(LA)*COS(TH)/SIN(TH)):AZ=FNAC
280 IF SIN(S-LO)>0 THEN AZ=6.28-AZ
290 EL=ATN((COS(TH)-.151046)/SIN(TH))
300 RA=SQR((1.81854E9-(5.37111E8*COS(TH))))
310 AZ=AZ*180/PI:EL=EL*180/PI:S=S*180/PI
320 DATA"AURORA I",143,"ANIK B",109,"ANIK C2",105,"ANIK D",104.5
321 DATA"ANIK III",114,"COMSTAR I",128,"COMSTAR II",95,"COMSTAR III",87
322 DATA"COMSTAR IV",127,"GALAXY I",128,"GALAXY II",74,"SATCOM I-R",139
323 DATA"SATCOM II-R",72,"SATCOM III-R",131,"SATCOM IV",83,"SBS I",100
324 DATA"SBS II",97,"SBS III",95,"TELSTAR 301",96,"USAT I",85,"WESTAR IV",99
325 DATA"WESTAR II",79,"WESTAR III",91,"WESTAR V",123,"END",999
326 AZ=FIX(AZ):EL=FIX(EL):RA=FIX(RA)
330 PRINTS$;TAB(13)S;TAB(24)AZ;TAB(39)EL;TAB(53)RA
340 GOTO 220
```

fig. 3. BASIC language program listing used to determine geosynchronous satellite azimuth, elevation angle, and range from your QTH.

spherical trigonometry and some trigonometric identities, the equations for azimuth, elevation, and range are:

$$\theta = cos^{-1}[cos\ (lat) \cdot cos(\Delta\ long)] \quad \textbf{(7)}$$

$$Az = cos^{-1}\ [-tan(lat)\ cos\ (\theta)/sin\ (\theta)] \quad \textbf{(8)}$$

If sin (Δlong) > 0 then $Az = 360 - Az$
and the

$$elevation\ angle = tan^{-1}[cos(\theta) - 0.151)/sin(\theta)] \quad \textbf{(9)}$$

where: $R/(R + H) = 6378/42168 = 0.151$
and:

$$Range = \sqrt{(R + H)^2 + R^2 - 2 \cdot (R + H) \cdot R\ cos\ \theta}$$

program hints

In the HP-67 program, the observer's latitude and longitude are replaced in decimal form. (Latitude is replaced in lines 3 through 7; longitude in lines 9 through 13.) Don't forget to use your own numbers — not mine — in these steps. The only other entry is the satellite longitude taken from **table 1**; after entry, hit key (A). Outputs are elevation, azimuth, and range in that order.

The BASIC program, which should need no explanation, prompts the user for all inputs. As shown in **table 2**, outputs provide satellite name, azimuth, elevation, and range in kilometers.

Locating the geostationary satellite you're looking for among the many orbiting the Earth in the crowded "satellite belt" is getting more difficult, but with a computer program such as this and some good microwave gear, they *can* be found.

STEP	KEY ENTRY	KEY CODE	COMMENTS
001	f LBL A	31 25 11	
	STO A	33 11	
	4	04	
	2	02	USERS OWN
	•	83	LATITUDE
	4	04	
	8	08	
	STO 0	33 00	LATITUDE
	7	07	
010	1	01	USERS OWN
	•	83	LONGITUDE
	2	02	
	2	02	
	STO 1	33 01	LONGITUDE
	RCL A	34 11	
	RCL 1	34 01	
	-	51	
	STO B	33 12	Δ LONGITUDE
	f COS	31 63	
020	RCL 0	34 00	
	f COS	31 63	
	x	71	
	g COS^{-1}	32 63	
	STO 2	33 02	θ
	f COS	31 63	
	RCL 0	34 00	
	f TAN	31 64	
	CHS	42	
	x	71	
030	RCL 2	34 02	
	f SIN	31 62	
	÷	81	
	g COS^{-1}	32 63	
	STO C	33 13	AZIMUTH
	f x > 0	31 81	
	GTO 0	22 00	
	f LBL B	31 25 12	
	RCL 2	34 02	
	f COS	31 63	
040	•	83	
	1	01	
	5	05	
	-	51	
	RCL 2	34 02	
	f SIN	31 62	
	÷	81	
	g TAN^{-1}	32 64	
	STO D	33 14	ELEVATION
	f - x -	31 84	
050	RCL C	34 13	
	f - x -	31 84	
	GTO 1	22 01	
	f LBL 0	31 25 00	
	3	03	
	6	06	
	0	00	
	h x <> y	35 52	
	-	51	
	STO C	33 13	
060	GTO B	22 12	
	f LBL 1	31 25 01	
	1	01	
	•	83	
	8	08	
	1	01	
	8	08	
	EEX	43	
	9	09	
	ENTER	41	
070	5	05	
	3	03	
	7	07	
	•	83	
	1	01	
	4	04	
	1	01	
	EEX	43	
	6	06	
	RCL 2	34 02	
080	f COS	31 63	
	x	71	
	-	51	
	f √	31 54	
	R/S	84	RANGE

REGISTERS

0	1	2	3	4	5	6	7	8	9
LAT	LONG	θ							
S0	S1	S2	S3	S4	S5	S6	S7	S8	S9

A	B	C	D	E	I
SAT LONG.	Δ LONG.	AZIMUTH	ELEVATION		

fig. 4. HP-67 program listing for locating geosynchronous satellites.

ham radio


6 STORE BUYING POWER
NEW!!
TH-21AT/
TH-41AT
TR-2600A
TM-211A/TM-411A
HERE'S WHERE
YOU GET THOSE
LOW, LOW PRICES ON
KENWOOD
PLUS FREE SHIPMENT
MOST ITEMS UPS SURFACE (Continental U.S.A.)
CALL NOW OR
DROP INTO ANY OF
OUR SIX LOCATIONS
ESTABLISHED FAVORITES
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ham radio TECHNIQUES

Bill Orr W6SAI

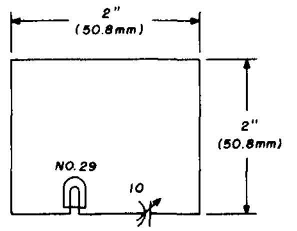
more about radials

Antenna experimentation is one field in which the enthusiast doesn't need an advanced degree in electronics and a room full of expensive test equipment. Sometimes a twenty-five cent "instrument" can provide meaningful results for the investigator.

A case in point: the experiments of Ralph, W8HXC, and Don, AF8B, which were designed to determine the effectiveness of quarter-wave radials on various 2-meter vertical antennas. The tests, conducted intermittently over a period of 5 years, pointed out some interesting aspects of radials that help to remove some of the mystery of VHF antennas.

The original investigation was designed to determine the best way to decouple the shield of a coaxial feedline from the field of the VHF antenna. The goal was to make the antenna do all the work, and to prevent the feedine from becoming part of the antenna. Only by making the feedline "inert" to the field of the antenna could the antenna do its job of laying down a low-angle signal.

To determine the degree of RF on the outer surface of the coaxial line, the simple "RF-sniffer" shown in **fig. 1** was built. It was used to detect current loops on the antenna elements, the feedline, and supporting mast and structure. Made out of junk box parts, the simple device worked well with transmitter powers as low as 7 watts.



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. 1. The RF "Sniffer" for 144 MHz is built on a cut-down ping-pong paddle. Capacitor is adjusted for greatest lamp illumination at the test frequency.

The "sniffer" consisted of a 144-MHz resonant circuit with a pilot lamp indicator, all mounted on a wooden handle. The capacitor was adjusted for maximum glow of the lamp (resonance) when held near the RF source used in the experiments.

The first experiments conducted were on a homebrew 1/4-wave groundplane antenna. It was found that the outside of the coax line, which dropped down beneath the groundplane antenna, was "hot" and exhibited a standing wave of energy along it that could be detected with the "sniffer." Excellent feedline isolation was achieved by simply wrapping the RG-58/U feedline into a two-turn coil 1-1/2 inches in diameter directly below the antenna. This little RF choke decoupled the feedline so that it was isolated from the antenna.

The next experiment was with an *extended half-wave vertical antenna.* RF was found on the feedline, and adding the choke in the feedline accomplished little. The outside of the line was still coupled to the antenna. Four quarter-wave radial rods were added to the antenna immediately below the matching coil (**fig. 2**). It was necessary to readjust the antenna for best SWR; however, the feedline isolation was *not* improved, and the radials did not seem "hot"with RF energy.

The last experiment, which was more meaningful, used a 5/8-wavelength antenna (48-inch long radiator) and a two-turn base matching coil (**fig. 3**). The feedline was carried down inside the metal supporting mast and a set of four quarter-wave radials with a clamping arrangement that allowed the radials to be placed anywhere on the mast was added.

Initially, the radials were positioned at the base of the antenna loading coil and the coil feedpoint was adjusted for best SWR indication. When a near-perfect match was achieved, the RF "sniffer" was used to examine the feedline. Unhappily, the feedline and mast indicated pronounced current loops over the entire length! The feedline and support pipe had become part of the antenna in spite of the radials, which were supposed to isolate the antenna from the feedline. In addition it was discovered that there was very little RF in the radials, a sure indication they were not doing their job.

Further experimentation proved that moving the radials down the mast, *away* from the antenna base, changed the SWR reading and required feedpoint readjustment. By cut-and-try a combination of feedpoint adjustment and radial position yielded excellent SWR, radials "hot" with RF and no detectable current loops on either the feedline or the supporting mast below the radials. Measurement placed this optimum radial position 3/8-wavelength below the base of the antenna. The radial angle was finally set at 45 degrees to the horizontal for best SWR.

Further tests with this antenna and with a car-mounted antenna of the same general type led to the interesting discovery that 5/8-wavelength long radials attached at the base of the 5/8-wavelength antenna provided the same excellent feedline isolation as did 1/4-wavelength radials attached 3/8-wavelength down the structure. A final experiment showed that radials could be attached to the mast at any point up to 3/8-wavelength beneath the antenna base provided that the sum of radial length and distance from the antenna base totalled 5/8-wavelength.

Don, AF8B, points out that the 5/8-wavelength vertical antenna plus the 5/8-wavelength long radial system is the same overall electrical length as an extended double-Zepp antenna.

The conclusion of the experiments is that radial length cannot be taken for granted and, in the case of an extended antenna, may not be 1/4-wavelength long. The important dimension is the overall length of antenna plus radial. The test to determine radial length is to use a "Sniffer" to make sure the RF remains in the radials and not on the outside of the coaxial feedline. (Thanks to Don, AF8B, for supplying data on the W8HXC and AF8B experiments.)

the Australian wideband dipole

Reader interest has been aroused by my description of the so-called "Australian dipole" wideband antenna (January, 1983, page 67). It seems that there is a whole family of wideband HF antennas and other related products of this type manufactured by *Antenna Engineering Australia PTY. Ltd.*, Box 191, Croydon, Victoria 3136, Australia. Contact Ian R.H. Wade, Sales Manager, for further information. The correct name of the antenna described in my January column is *Model 632 Travelling Wave Dipole.*

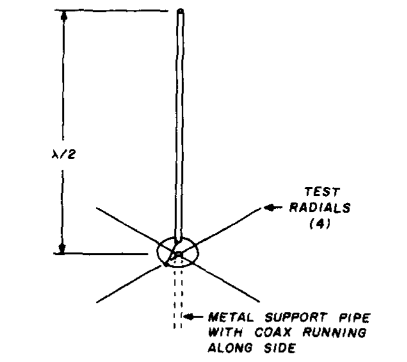


fig. 2. Test radials were added to extended 1/2 wave antenna.

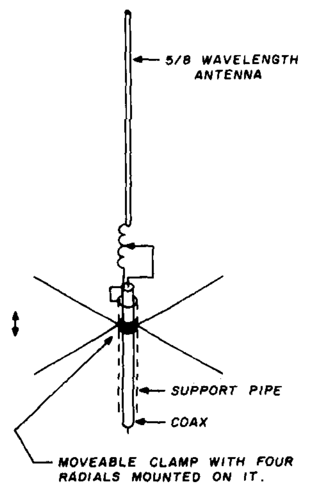


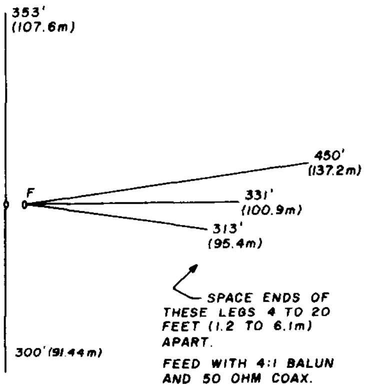
fig. 3. 5/8-wavelength "gain" antenna was mounted to support pipe and feedline passed down inside the pipe. A set of 1/4-wave radials with a mounting clamp was placed on the pipe. Radials could be moved up and down with reference to the base of the antenna.

the K4EF "all-band" antenna

Several years ago Ev Brown, K4EF, described a wire antenna that would cover all HF Amateur bands between 80 and 10 meters (*ham radio*, May 1977, page 10). Since then he's done a lot of work on his design and has devised a new configuration that has several advantages over the old one. The new antenna covers the 160-meter band, uses four support points instead of five, and occupies less space. In addition, because the elements are arranged in a V-configuration, it provides some signal gain on the higher frequency bands.

A plan view of the new antenna design is shown in **fig. 4**. The array consists of five long wires arranged in a semicircle. The antenna is fed at points F-F with a 4-to-1 balun and a 50-ohm transmission line. In actual use, one of the two elements at the left of the illustration is used with one of the three wires at the right. The wires can be selected from the operating position with a remote switch. For example, if the 353-foot wire is added to the 313-foot wire, an element 666 feet long is produced. An odd number of half waves is required to produce approximately 200 ohms feedpoint impedance at or near the element center. The chart of **table 1** shows the odd-halfwave resonances in this combination. As can be seen, the bandwidth coverage is enormous (see column 3), and when you consider that the 666-foot combination is merely one of six, the complete configuration provides wide spectrum coverage with very low SWR. A simple computer program could calculate all of the resonances and bandwidths for all elements. The results could then be combined to determine what frequency gaps (if any) exist in the complete array coverage.

As Ev says, ". . . it is difficult to convey to a ham who has never used an all-band, broadband antenna just how convenient it is. During contests, changing bands is accomplished by flipping the bandswitch. Checking band conditions is done in an instant.



element length feet	Amateur band resonant frequencies 160	80	40	30	20	18	15	12	10
613		3970	7180	10.4		18.4		24.8	28.0
631		3850	6950	10.1		17.9	21.0	24.1	28.8
666		3650			14.0		21.4		30.2
684		3585							29.4
750	1940								
803	1820								

fig. 4. Two legs of antenna comprise an element. For example, the 353-foot leg plus the 313-foot leg form an element 666-feet long. The chart shows resonance at 3650 kHz, 14.0 MHz, 21.4 MHz, and 30.2 MHz. Other combinations provide additional resonant frequencies. (Top view of antenna shown.)

My FOC friends frequently ask to get credit for another band and find me waiting for them. Perhaps the most important aspect of the idea is that it encourages the operator to use the whole spectrum available."

the W2TBZ quad-loop beam antenna

I had not seen Sid, W2TBZ, for over 15 years and our QSOs on the air were few and far between. "Keep in touch," I had said, and just recently I heard from him — with a new antenna idea that he was using with great success on 15 and 20 meters.

To stay in touch with his friends, Sid needed an inexpensive wire beam that could be easily erected and would provide a modest amount of gain and a low angle of radiation. Various antennas were tried, and the final version, a 2-loop Quad beam is shown in **fig. 5**. Estimated gain of this bidirectional array is about 4.5 dB over a dipole.

The antenna consists of two side-by-side Quad loops, horizontally polarized and driven in phase. The feed system consists of two equal lengths of 300-ohm TV line and a 1-to-1 balun. The feedpoint impedance of a single loop in this configuration runs about 120 ohms, so parallel-connected loops provide a terminal impedance close to 60 ohms. This provides a good match to a 50-ohm transmission line system.

The 40-foot masts support the antenna. The figure-8 radiation pattern is at right angles to the plane of the array. The pattern is sharper than that of a dipole, being about 60 degrees between the half-power (−3 dB) points.

table 1. Odd halfwave resonances in 666 feet of wire.

band meters	electrical length halfwaves	resonant frequency (MHz)	bandwidth to 2:1 SWR points
80	3	2.179	2.142 to 2.216
	5	3.657	3.597 to 3.717
	7	5.134	5.057 to 5.211
	9	6.612	6.512 to 6.712
	11	8.089	7.969 to 8.209
	13	9.567	9.427 to 9.707
	15	11.044	10.879 to 11.184
	17	12.522	12.342 to 12.702
20	19	13.999	13.789 to 14.209
	21	15.477	15.244 to 15.709
	23	16.954	16.999 to 17.208
	25	18.432	18.155 to 18.709
	27	19.909	19.610 to 20.207
15	29	21.386	21.065 to 21.706
	31	22.864	22.521 to 23.206
	33	24.341	23.975 to 24.706
	35	25.819	25.431 to 26.206
	37	27.296	26.886 to 27.705
10	39	28.774	28.342 to 29.205
	41	30.251	29.797 to 30.704

Note: The 666-foot element (summarized above) is only one of six element combinations. Single element switch will provide enormous coverage of HF spectrum with low SWR.

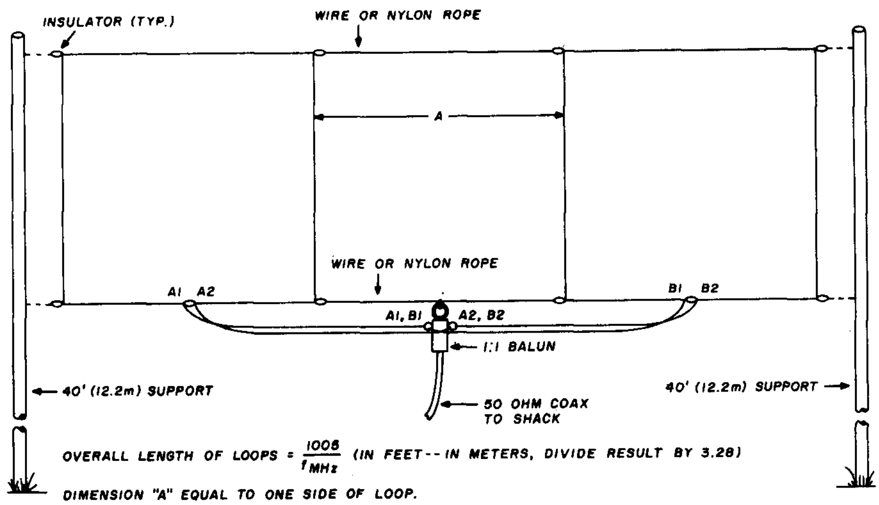


fig. 5. The W2TBZ loop beam. The phasing lines are made of 300-ohm TV line. Length of phasing line = $\frac{468}{f_{MHz}}$ **× velocity factor of 300-ohm line. Balun is supported at center of lower nylon rope.**

interesting reading!

From time to time I like to recommend interesting books or periodicals that provide information that otherwise may be unobtainable, and that are of general interest to Radio Amateurs.

This month's recommendation is *The Monitoring Times*, published monthly by Grove Enterprises, Inc., 140 Dog Branch Road, Brasstown, North Carolina 28902. The subscription rate is $10.50 for one year.

The Monitoring Times is full of timely information about what's going on in the HF/VHF spectrum. The editor and publisher is Bob Grove, WA4PYQ. This newspaper covers items of interest not generally found in Amateur publications. I look forward with interest to each issue! The latest information on the mysterious "beacon" and "numbers" stations may be found in this publication, as well as up-to-date information and interesting stories of other aspects of radio communication.

Some of the columns in *Monitoring Times* are "High Seas Radio," "Signals from Space," "Utility Intrigue," "RTTY/FAX," and "Pirate Radio." There's also a good review of some of the new communications receivers in the present issue of this interesting publication.

Good luck, Bob — you have a winner!

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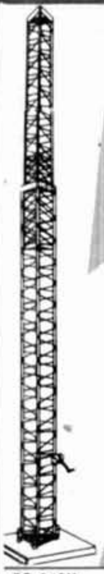

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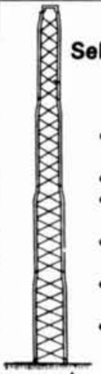

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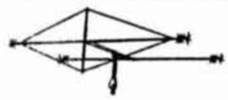

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make your own PC boards
using silk screen techniques

A step-by-step guide to easy, inexpensive duplication of simple circuits

At least two silk-screen approaches to PC board *duplication will work for reproducing relatively simple* circuits. One is based on the use of printing film, and the other is based on the use of a photo-sensitizing material that can be applied directly to a silk screen or to a board. I have found both of these techniques to be quite satisfactory, and I consider them to be superior to the usual board photo-sensitizing approach for all but the most sophisticated circuit configurations. The screen-sensitization technique can be used with "LIFT-IT"™ patterns or by applying sensitizer directly on the board if one is certain replication will not be necessary. The printing-film approach is suitable for relatively simple circuits such as those used for RF voltmeter probes.

In order to produce a conductive pattern on a circuit board it is necessary to transfer a drawn pattern to the copper on a board. This requires the application of material that will protect the desired conductor area from an etchant. Of the various methods available, silk-screen techniques are probably the least expensive and most convenient solution to the typical multi-board problems encountered by hams. (Where only single boards are required, the photo-sensitizing method can be applied directly to the board.)

board preparation

I buy my copper-clad material (copper one side) at hamfests, usually for less than one cent per square inch, far less than the 20 cents or more charged for sensitized boards.

First I cut the board to size with a bandsaw. Metal shears or a pair of tinsnips can also be used. The board may also be scored with a linoleum knife and separated. After the boards have been cut to size, the edges and corners should be deburred to avoid cutting the silk in the process of inking. If the board is badly corroded the copper surface should be scoured with 600 grit emery, and finally with a cleaning powder such as "Old Dutch Cleanser," one that is free of chlorides and phosphates. An all-over clean copper lustre is required to assure efficient etching.

mounting frame preparation

Two kinds of mounting frames can be used. Because the boards I use are seldom larger than 3 by 5 inches, I purchased some 3/4-inch square wood strips and cut them into 6- and 8-inch lengths. Using picture frame clamps, I assembled these pieces into frames having outer dimensions of about 7 by 9 inches, gluing the pieces of wood together with white glue and inserting 2-inch long wood screws through the joints and reinforcing the joints with flat L brackets measuring 1-1/2 inches on each side (see **fig. 1**). After assembly, the forms should be protected with shellac to improve their resistance to water. I use these frames for board applications having continuous use, such as power supply configurations.

It is difficult to get enough tension on the screen to minimize under-flow with this arrangement. I have found it convenient to attach 7-inch pieces of flat aluminum stock about 3/4 of an inch wide on the inside of the long sides. These can be used to stretch the screen tightly. Beware of sharp corners on the tensioning bars; any burrs or sharp edges or corners will cut the silk. I cut slots in the bars and use screws to hold them in place. Much less underflow results.

By Keats A. Pullen, Jr., W3QOM, 2807 Jerusalem Road, Kingsville, Maryland 21087

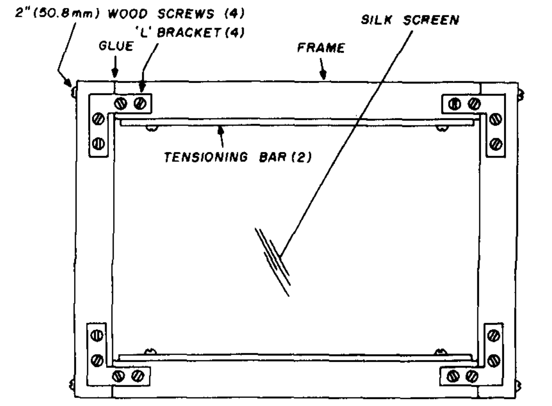

fig. 1. Wood silk-screen printing frame with pattern. The *irregular outer edge of the wood frame is caused by the* silk. I didn't use the cardboard reinforcement with this frame.

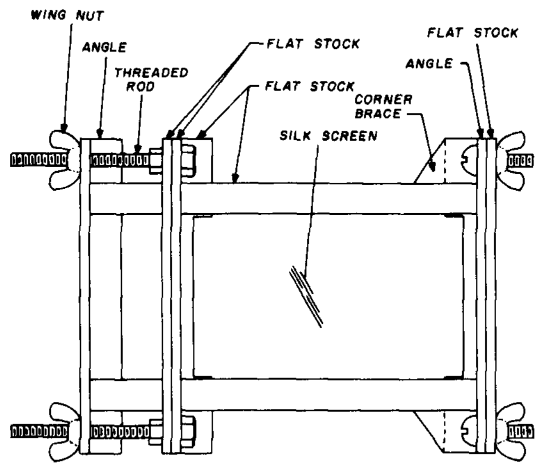

fig. 2. The metal screen printing frame. The cardboard reinforcement is used here.

The second kind of frame can be made from ordinary aluminum stock available in most hardware stores. I use 3/4-inch angle and 3/4-inch flat stock. One clamping surface for holding the screen is fixed; the other is moveable. There are two fixed elements, the second being used for application of the required tension. One of the fixed angle pieces is reinforced to the flat bars with corner braces for additional stiffness. The moveable angle is coupled to the second angle piece with 1/4 x 20 inch threaded rod; wing nuts are used for adjusting tension on the silk.

The one fixed angle element and the moveable one are arranged so that the two ends of the screen, supported by cardboard as explained in the next section, can be clamped tightly to the two members. In this way, ample tension can be applied to the screen for use in printing (see **fig. 2**).

screen preparation

The silk screen is prepared by washing, again with the cleanser, and thorough rinsing. A monofilament nylon screen material of the finest possible mesh is best and will give the finest resolution and minimum problem from etch-through resulting from blockage of ink penetration by the screen material itself. The screen must be stretched as tightly as possible when used, since only then can sufficient contact of the pattern and the copper be achieved, minimizing "rununder."

To protect twisting the thread pattern of the screen material, use cardboard bracing strips on each tension edge, leaving enough silk to wrap around the strip. The silk can then be stapled to the cardboard strip and the combination tacked on the frame or clamped as required. This way the stress can be distributed uniformly on the silk.

using printing film

Since there are two possible ways the screen master can be used, each method is considered separately. I have found orange printing film to be useful and easy to prepare for simple circuits. In using it, one simply marks off and removes narrow ribbons of film to form conductors, lifting them from the backing material. The material removed represents a current path. Care should be taken to minimize the cutting of the backing, a plastic, nylon-type material, as the transfer of the film to the silk is most easily accomplished if the film has been cut through without scoring the backing.

I have made some simple tools for preparing the film. One type, for cutting conductor paths, consists of two halves of a double-edged razor blade mounted on opposite sides of a piece of used copper-clad (see **fig. 3**). This will cut both sides of a conductor path at one time, and help in making sure that the length of the cut is correct. These cuts can extend to about a hundredth of an inch into an adjacent pad or across an intersecting path to simplify the removal of the material. This ribbon is then picked up with an Xacto™ knife or a pin and removed. Pads can be cut with a tool made by taking a short length of 1/4-inch rod, center-drilling it on a lathe or drill press, and cutting down its outside diameter to the size of the pad required.

When the pattern has been prepared, it may be attached to the silk screen material by stretching the screen tightly over the pattern and patting the screen with a piece of cheesecloth wetted with lacquer thinner (use a gentle push, not a sliding motion). You will be able to see where the attachment is satisfactory.

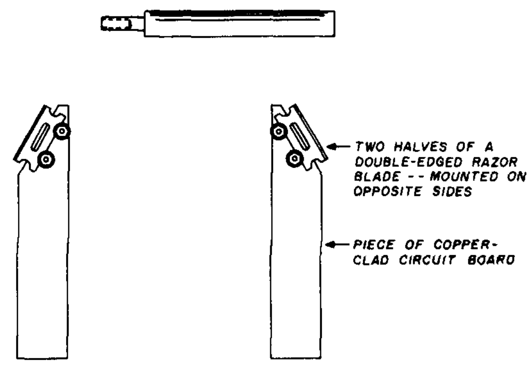

fig. 3. Special tools for use with printing film.

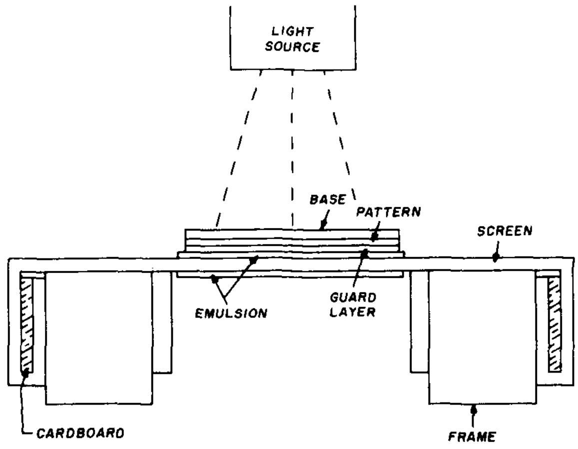

fig. 4. Arrangement for preparing a screen master from a Lift-it of a photo copy of a circuit.

You will want to go back and redo any imperfectly imbedded areas. When the combination has dried completely, carefully peel off the backing, resticking if required.

photo-sensitized silk screen

The silk can also have the required pattern applied to it by the use of a photographic sensitizing technique. The sensitizer I have used is the Hunt Manufacturing Company Printing Photo Emulsion Kit No. 4533. This contains two components which are mixed just prior to use. Instructions are provided with the package. A leaflet on screen printing is also available.

To prepare the photo screen, mount the screen material on the frame you have chosen and apply the mixed sensitizer in a thin, smooth layer on both sides of the screen. (You can expedite drying by blowing *cool* air from a hair dryer on to the screen.) After mixing, handle the coated screen in semi-darkness only. If the image to be transferred is closest to the back of your image master, you expose with that surface adjacent to the sensitive surface and expose through it. (Lift-it masters are exposed from the top, whereas drafted masters will be exposed from the bottom; see **figs. 4** and **5**). The master should be between the light and the sensitive layer, and the image as viewed from the top should be as required. A transparent cover should be placed on top of the master and weighted so as to assure close contact between the master and the screen. (I use a No. 2 photoflood in a reflector about 14 inches from the work, for about six minutes.) The exposed silk is then washed and rinsed immediately.

inking

After the circuit board has been scoured and prepared for use, and the screen with the appropriate pattern is in tight contact with it, the inking can be begun. The ink must be reasonably thick, yet it must spread through the open areas of the screen. At the

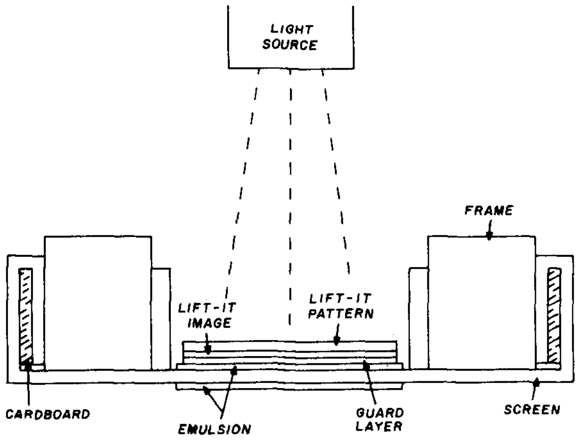

fig. 5. Arrangement for preparing a screen master from a board drawing positive master.

same time it must be able to be completely removed from the stencil screen without leaving residues or damaging the screen. It must dry "hard" — that is, it must, after drying, be resistant to the etchant.

A bead of ink is spread along the short length of the circuit to be printed, and then spread along the image of the circuit. I use a piece of Plexiglass™ or other transparent acrylic as a spreading tool. It should be an inch wide or wider, and can be wide enough to cover the entire width of small boards. (All burrs and sharp edges should be removed from the spreader prior to use. To re-use, simply peel off the dried ink.) Acrylic inks such as the Hunt Permanent Acrylic Screen Printing Ink or the Liquitex Permanent Acrylic Ink are suitable.

After the printed board is dry, the image can be touched up by usng pin or a needle to repair breaks, or an Xacto knife to scrape away any run-unders that may have occurred. I usually use a hair dryer *with heat* to speed the drying in this phase of production.

I print as many boards as I need in rapid succession and then wash out the screen master with a thorough spray of water. (Printing inks are soluble in water until they dry; after drying, they become impervious to water but can be peeled or scraped off.)

initial artwork preparation

When the photographic screen method is used, it is necessary to work from some kind of master. These masters may be those printed in a magazine (either positive or negative) or some you have prepared from any of the various commercial materials. Each approach is discussed here.

A complex circuit or one available as a circuit pattern in an article can be made into a screen by combining a photocopy of the layout with the silk-screen process. The photocopy can provide increased contrast, if necessary, and eliminate the need to cut the magazine. If the original is positive, make a Lift-it from the photocopy and use it to expose the screen. If the original is a negative, make the Lift-it copy and then print it directly onto a piece of high-contrast 4 x 5 inch cut film. This will give a positive that can be used to sensitize the screen again. (With a negative, the print may be made directly from the Lift-it to the board using the Hunt preparation if you prefer. This works, particularly if the hardener is used as described later.)

I make some of my masters on tracing vellum using extremely thin transfer materials such as those made by Vector. Ruled India ink lines are suitable for conductors. Transfers used for pads and IC sockets finish the circuit layout. The result is a simple, direct step-by-step process.

washing and etching

Washing must be done at several points in this process. The boards should be washed thoroughly and carefully after scouring.The screen material should be washed thoroughly from both sides to remove any sizing and acrylic ink. With the photo-sensitization process, it is spray-washed from both sides to remove the filler from the pattern.

With both the exposed photo-screen and the inked screen, I use a discarded spray bottle for washing, which must be done immediately after completion. A fine but fairly hard spray is best.

I generally use ferric chloride as an etchant. Either plastic or glass trays may be used with it; I use Pryex™ glass trays so I can heat the etchant and thereby speed the operation. My heater is an electric plate warmer with two switches added to the line cord, one with a diode connected across its points for the convenience of two heat levels.

After etching, the board should be washed thoroughly. You'll find the ink softened enough to peel off, leaving the copper with the dull appearance of cuprous oxide. If you wish to apply tinning solution, the copper must be made bright once again by the use of 600 grit paper used lightly as needed and scouring. Hardened ink can be dissolved in lacquer thinner.

hardening

The photo-emulsion image on the screen can be hardened by treating it with Hunt's "Permanizer"™ No. 4529. The developed and dried image on the silk screen is painted with this material, and the combination dried with cool air. The use of a water spray wash with cold water once again opens the mesh where the pattern is.

acknowledgement

I am deeply indebted to A. L. Spizzo of Hunt Manufacturing Company for his assistance in solving various technical problems I have encountered.

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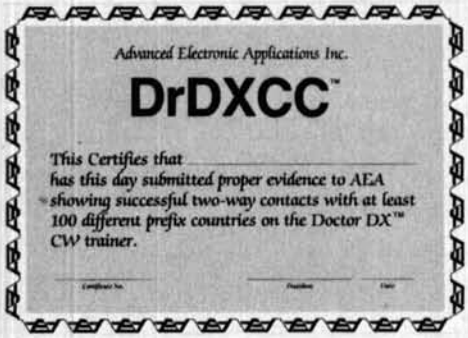

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winter DX

The winter DX season is here. One characteristic of winter is a steep rise in the daily MUF peak followed by an early decline to a deeper predawn minimum. This makes for shorter daytime DX operating time in the higher HF bands, but for more nighttime DX on the lower frequency bands. Signal strengths are higher because of lower absorption of energy and less propagated or local atmospheric noise (by this time of year, thunderstorms are fewer and more distant).

Absorption is a result of the loss of energy from the signal as it collides with ions on its path through the D region (about 100-120 miles, or 60-80 km, above the earth). How much energy is absorbed per transit of the D region depends on the location of the sun, and is a function of cosine X, the zenith angle to the sun. Maximum absorption occurs at the subsolar point (directly under the sun); absorption decreases as the signal transit moves away from the subsolar point in any direction. In our winter the subsolar point moves down to 23 degrees south latitude, resulting in less absorption. At the same time the earth is closer to the sun by 2 percent. The net result is still less absorption in winter. The degree of absorption is related to and follows the changes in the ultraviolet output of the sun. (It takes slightly over 8 minutes for a change on the sun to begin affecting our ionosphere.) A measure of this is the daily solar flux at 2800 MHz recorded in Ottawa, Canada, and broadcast at 18 minutes after the hour by radio station WWV. Another source of absorption, caused by increased particle influx during geomagnetic storms, occurs on propagation paths through or along the auroral zone (60-80 degrees latitude). An indication of this cause is an increase in the geomagnetic K (greater than 4) and A (greater than 30) indices, also broadcast from WWV.

On any propagation path, absorption increases with the number of transits of the D region and also varies inversely with frequency. Therefore in working DX it pays to use the higher frequency bands to obtain more distance per hop (resulting in fewer transits) and less signal loss. This is why we generally think of 6, 10, or 15 meters for DXing. But in winter, we have the opportunity to work DX on the lower frequency bands with less QRN and lower signal loss than at any other time of the year.

Lower signal loss is something to look forward to, but you can't count on it. Sometimes in winter, signals are poor for several days at a time. This is caused by anomalous absorption, which will be discussed in next month's column.

last-minute forecast

The low HF bands, 160 through 30 meters, are expected to be the favored bands of operation during the first two weeks of November, with higher bands providing the best DX during the last two weeks of the month. The solar radio flux should be about the same as last year's values, yet higher than it's been in the last month or two. Some possibility of recurrent geomagnetic storms still exists, with greatest probability of occurrence on November 4, 9, 14, 18, and 28. Remember: even though disturbances affect signal strength and produce fading conditions for some paths, conditions on other paths may actually improve.

November is the month during which numerous meteor showers occur. Shower activity should begin on October 26 and last until November 22. A shower maximum of ten per hour is expected during the Taurids meteor shower from the 3rd through the 10th. Lunar perigee is on the 20th; full moon is the 8th.

A total eclipse of the sun will occur on November 22 and 23 in the south Pacific, starting at 2013 UT in the Philippines and New Zealand, traveling east to Antarctica, and ending at 0133 UT. You might want to schedule some contacts with ZL and KC4 land for some unusual DX.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of high solar flux and a disturbed geomagnetic field. Absorption effects are not too noticeable.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters. Skip distances and signal strength may decrease during midday on those days that coincide with high solar flux values. Nighttime DX will be good except after days of very high MUF conditions and the winter anomaly. The usable distance is expected to be somewhat greater than that achieved on 80 at night.

Eighty and one-sixty meters are the nighttime DXer's bands. The bands open just before sunset and last until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values don't affect these bands much. The anomaly will affect day and night signal strength on some days.

ham radio

NOVEMBER

WESTERN USA

GMT	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	
PST	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	
ASIA FAR EAST	30	30	30	40	40	40	40	40	40	40	30	30	30	30	30	30	30	20	20	20	20	20	20	20	N
EUROPE	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	40	40*	30	30	30	30	30	30	NE
S. AFRICA	15	15	15	15	15	15	15	15	15	20	20	20	30*	20	20	20	20	20	20	20	20	20	15	15	E
S. AMERICA	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	10	10	10	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	S
NEW ZEALAND	10	10	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20*	15	15	15	10	10	SW
OCEANIA AUSTRALIA	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	10	10	W
JAPAN	20	20	20	20	30	30	30	40	40	40	40	30	30	30	30	30	30	30	20	20	20	20	20	20	NW

MID USA

MST	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	
ASIA FAR EAST	30	30	30	30	30	30	20	20	20	20	20	20	30	30	30	40	40	40	40	40	40	40	30	30	N
EUROPE	20	20	20	20	20	20	20	30	30	30	30	40	40	40	40	40	40	40	30	30	30	30	30	30	NE
S. AFRICA	15	15*	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	E
S. AMERICA	15	15	10	10	10	10	10	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	S
NEW ZEALAND	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	10	SW
OCEANIA AUSTRALIA	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	W
JAPAN	20	20	20	20	30	30	30	40	40	40	40	40	30	30	30	30	30	30	30	20	20	20	20	20	NW
CST	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	

EASTERN USA

EST	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	
ASIA FAR EAST	30	30	30	30	30	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	40	40	40	30	N
EUROPE	20	20	20	20	20	20	20	20	30	30	30	30	40	40	40	40	40	40	40*	30	30	30	30	30	NE
S. AFRICA	15	15*	10	10	15*	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	E
CARIBBEAN S. AMERICA	15	15	10	10	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	S
NEW ZEALAND	10	10	15*	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	SW
OCEANIA AUSTRALIA	15*	15	15	15	15	15	15	20	20	20	30	20	20	20	20	20	20	20	20	15	15	15	15	15*	W
JAPAN	20	20	20	20	30	30	30	40	40	40	40	40	40	30	30	30	30	30	30	30	20	20	20	20	NW

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

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Joe's done it again — a column of interest to HF'ers and VHF'ers alike. — Ed.

VHF/UHF WORLD

Joe Reisert, W1JR

high dynamic range receivers

Mention high dynamic range and you'll really get a discussion going. Everyone has a story and a solution. We'd all like to believe that our receivers or transmitters are always clean, and that any splatter or other obnoxious noise has to be coming from somebody else's poor receiver or dirty or overdriven linear amplifier. Because all aspects of the situation, both on the transmit and receive side, are seldom separated, the problem is rarely resolved.

A few years ago, after lots of arm-twisting by Jim Stitt, WA8ONQ, I tackled this dilemma. The main goal was to improve Jim's 6-meter receiver sufficiently so he could be sure it wasn't the culprit in these situations. Then his station could be more competitive in the VHF contests and he'd be able to operate alongside the strong local transmitters — assuming *they* were also clean (more on this later).

The end product was a high dynamic range 6-meter receive converter with an extra transmitter LO (local oscillator) output. Since this month's *ham radio* emphasizes receivers, I decided to discuss this subject in some detail and examine some of the problems, limitations, and solutions for such a design. Typical recommended circuits for a 6-meter receive converter will then be shown.

high dynamic range

What is high dynamic range? One answer is that it describes transmitter or receiver design that allows copying weak DX signals in close proximity to strong signals. It sounds simple enough, but how strong is that local? Let's assume that a big 6-meter station is only 1 mile (1.6 km) distant and runs the legal limit of 1.5 kW PEP output through a feedline with a 1 dB loss to a 10 dB gain antenna pointed at your receiving antenna. If you're using a dipole receiving antenna broadside to this signal and have no feedline loss, the signal at the input to your receiver will be approximately 0 dBm or 1 milliwatt, about 130 dB above the noise floor in a typical VHF receiver! If you also have a 10 dB gain antenna, the signal received (when aimed at this source) will be +10 dBm (10 milliwatts) — more power than is used for the LO in most Amateur receivers! If the distance between stations is doubled, the signal will drop by 6 dB but still be quite respectable.

Recently I tested the dynamic range of a well-designed Amateur 6-meter converter that uses a single JFET preamplifier and a standard level (+7 dBm or 5 milliwatts) DBM (doubly-balanced mixer). When two equal signals of −20 dBm (23 millivolts or 10 microwatts) were present at the input to the converter, spurious signals or IMD (intermodulation distortion) were generated and only 30 dB below the desired outputs. This is hardly high dynamic range! These spurs or IMD products usually appear as sidebands or additional signals spaced equally above and below the normal signals by the difference between the input frequencies (see **fig. 1**). When the IMD gets worse, additional spurs appear alongside the first sidebands as is also seen in the photo in **fig. 1**.

All is not lost. The antennas can be part of the solution. If you use a directional antenna with a clean pattern, moving your antenna back and forth can place a null on a strong signal.[1] If overload is still present, one solution is to place an attenuator at your receiver input (more on this later); on VHF, especially on 6 meters, where the local ambient noise is usually high, the weak signal will still be good copy while the local (interfering signal) will have been "knocked" down.

is the receiver at fault?

Before proceeding, it may be worth mentioning the transmitter. Typically speaking, it would be desirable for all Amateurs to transmit a clean signal. But what is a clean transmitted signal? Typical Amateur linears call for IMD products to be at least 30 dB below the desired signals. However, if the received signal is 130 dB above the noise, 30 dB IMD isn't going to be much help 3 kHz from the other station's frequency. You'll just have to QSY further away.

Therefore, before you accuse the other station of "hitting it too hard," perform one simple test. *First observe the splatter several kHz away* on a relative basis or on your receiver "S" meter. Next, place an appropriate attenuator (10 dB recommended, see **fig. 2**) at the input to your receiver or converter (or use the internal attenuator if one is an integral part of your receiver) and then recheck the splatter level. If the received level drops by approximately the insertion loss of the attenuator, the transmitter is the culprit and the transmitting station is

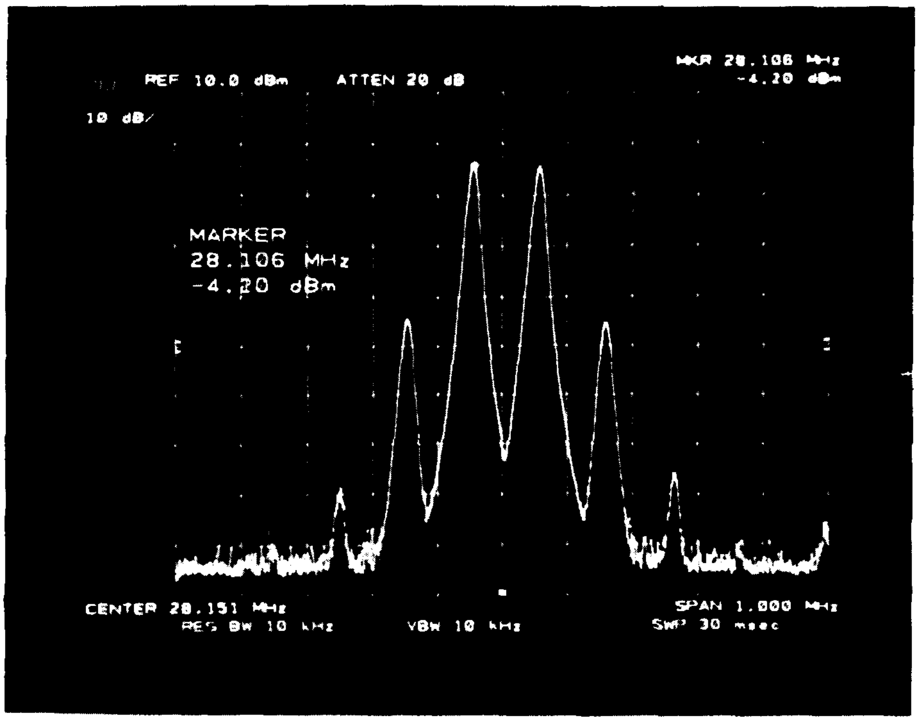


fig. 1. A spectrum analyzer shows the 28 MHz IF output of a typical 6-meter converter as described in the text. Input signals are approximately 50.1 and 50.2 MHz at − 20 dBm each.

either overdriving its equipment or your frequency is just too close for comfort. However, if the splatter drops by more than the attenuator value (it could be up to 3 times less!), your receiver is surely part of the problem.

Assuming that the problem is the receiver (maybe in some future column we'll examine transmitter and power amplifier requirements in greater detail), there are design approaches that will enhance receiver performance.

general receiver design requirements

The old saw "If you can't hear them, you can't work them" still applies to high dynamic range receivers. Low noise figure, sufficient RF selectivity to reject images and undesired out of band signals plus a clean local oscillator are still required.

In order to obtain low noise figure, a preamplifier is usually required ahead of the first mixer. Herein lies the problem. Any gain ahead of the mixer will *decrease the dynamic range*. Therefore, the preamplifier gain must be kept as low as possible, consistent

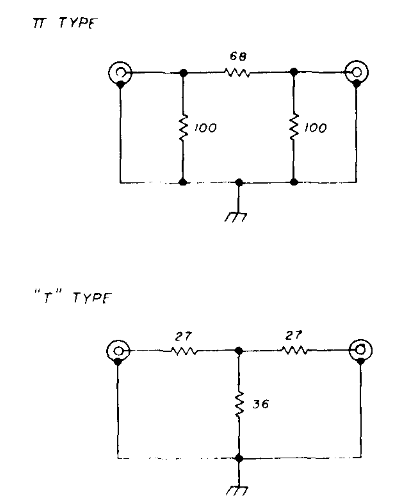


fig. 2. Typical 10 dB attenuator pads for testing or improving the dynamic range of a receiver.

with obtaining the desired noise figure. The desired or total system noise figure can be determined from:

$$NF\,(total) = NF1 + (NF2 - 1)/G1 + (NF3 - 1)/G1G2,\ etc. \quad \textbf{(1)}$$

where G is gain and NF is noise factor (a numeric-not dB) for each stage in succession.

For example, a 3 dB noise figure preamplifier with 9-10 dB of gain feeding a mixer with a 10 dB noise figure will yield an overall system noise figure of less than 5 dB. Fortunately, we can live with lower sensitivity receivers, especially on 6 meters, where a noise figure less than 5 to 10 dB is usually wasted since the typical ambient noise is usually very high.* At 2 meters the problem is more acute, but the level and number of signals are usually somewhat less of a problem. For the computer-minded Amateur, a computer program is available for **eq. 1** so that you can quickly iterate various combinations of gain and noise figure to determine your own optimum case.[2] Before leaving the subject, remember that a preamplifier must have high output power capability in order not to distort the signals prior to the first mixer (more on this later).

For good performance you need adequate RF selectivity ahead of the mixer, which means additional losses that further increase the noise figure. Again, these losses can be handled (as we shall soon see) by the proper choice of RF filter characteristics and by optimizing the location of the filters in the receiver chain.

Let's not forget the choice of IF and its effect on selectivity, images and spurious responses. For 6 meters I personally favor a 28 MHz IF with a 22 MHz LO, rather than a 14 MHz IF with the 36 MHz LO used on some of the older converters. Spurious signal analysis reveals that a 28 MHz IF is slightly less susceptible to "birdies."[3] Also, a 28 MHz IF is far less likely to respond to IF breakthrough. The latter term refers to leakage at the IF frequency that permits normal signals in this spectrum to also be received. The 20-meter band is a good example since propagation yields signals of high intensity, especially during the days when 6 meters is hot. Although 10-meter IF breakthrough can be a

*Everything's relative — Joe, who also operates 160 meters, would probably agree that in contrast, 6 meter ambient noise is low — **Editor** problem, the number and strength of stations present is usually less, especially below 28.3 MHz.

High dynamic range mixers that require moderate LO power (10-100 milliwatts) are usually required. Also the LO must be very clean with low phase noise (more on this later) and should be followed by an amplifier to boost the level high enough to adequately drive the mixer.

Finally, if the overall system noise figure is to be realized, the mixer must usually be followed by a low noise figure postamplifier with a high dynamic range. The IF receiver should also have high dynamic range and a moderate (10 dB or so) noise figure.

preamplifiers

Surely the preamplifier is one of the most important aspects of a good receiver. However, obtaining high dynamic range and low noise figure simultaneously and with a reasonable input and output VSWR is difficult. Devices (transistor, FET, etc.) with low inherent noise figure are common. However, increasing preamplifier dynamic range usually requires increased device current or a device with greater current-carrying capacity. This, in turn, usually increases the noise figure and the overall gain, the exact opposite of the desired effect!

Before discussing different preamplifiers in detail, it may be well to mention the subject of the linearity in an active device. Just because an amplifier is operated in class "A" doesn't mean it is free from distortion. *Every amplifier, regardless of its type and power, has a point beyond which the output signal will no longer be an exact replica of the input signal.* Hence distortion will occur.

Over the years various methods have been devised to measure distortion. The most frequently used test is for 1 dB compression. This is defined as the CW power level where the output signal increases 9 dB for an input power increase of 10 dB. Most class "A" amplifiers can only increase output power by 2-6 dB beyond this level, as shown in **fig. 3**. Amplifiers often are

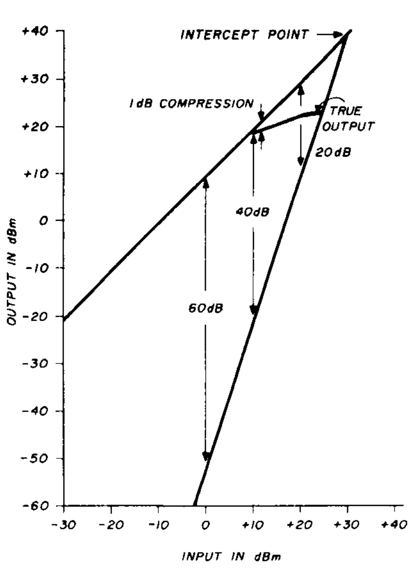


fig. 3. A high dynamic range preamplifier (fig. 6) input versus output level response with typical IMD levels and compression levels.

heavily distorting a signal 5-10 dB before it reaches compression levels; consequently, this is not a good point for referencing distortion. Furthermore, some devices are more nonlinear than others, especially when approaching the compression point.

In 1967, McVay wrote his classic reference paper on the third-order intercept point, a new method of measuring dynamic range.[4] Basically, what this method does is to determine distortion based on a two-signal IMD test performed in a similar manner to that used to specify single sideband linear amplifiers. The third-order intercept point is then determined either mathematically or by use of nomograph (see **fig. 3**). The distortion can then be calculated or read off the nomograph for any power level on any device if the third-order intercept point is known. Suitable nomographs are available in reference 4 and from most commercial amplifier manufacturers.

In most of the work I have done on high dynamic range, I have used the intercept point test method. Several things are immediately apparent. The IMD products increase at three times the rate of the desired output signal level change. Hence, the ratio between the output signal level and distortion will change on a 2 for 1 basis. For example, if the IMD products from two equal level signals are 60 dB below the desired output signal level and the signal level is increased by 1 dB, the IMD products will now be only 58 dB below the desired output level. A 10 dB input signal increase will decrease the IMD difference by 20 dB. This can be seen graphically in **fig. 3**. Therefore, once IMD becomes apparent, it will usually degrade very rapidly, perhaps even on a greater than 2-to-1 basis, with increased signal level! This is common on many active devices whenever the IMD is less than 60 dB below the output levels.

A search was launched for the ideal preamplifier. First a low-gain (12-13 dB) grounded-gate J310 JFET preamplifier was designed (after all, FETs are supposed to have such great dynamic range and low noise figures). The results were fair. On a typical circuit the output compression point was +14 dBm (250 milliwatts). The IMD was down 60 dB for −3.5 dBm (0.45 milliwatts) output per signal for an output third-order intercept point of +26.5 dBm (450 milliwatts). However, the noise figure was over 4 dB with no input matching. When input noise figure matching was added, the gain increased and the input impedance match degraded — both detrimental to the desired results. Also, the overall selectivity for this preamplifier was inadequate for the final converter.

Before proceeding with the next preamplifier design, some re-examination was in order. Previous experience with modular circuits led to the conclusion that in a high dynamic range receiver all circuits should have good input and output VSWR at a common impedance such as 50 ohms.[5] This would allow easy interchange between filters, amplifiers, mixers, and LOs, as well as facilitate any future improvements or changes, especially when new or improved devices became available.

With this in mind, the search for a low-noise high dynamic range pre-

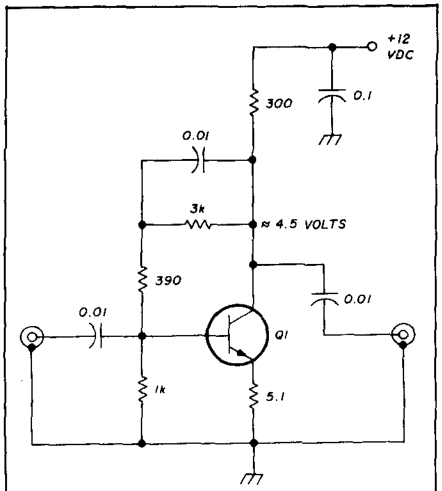


fig. 4. A high dynamic range 14 dB gain preamplifier using RF feedback for wide bandwidth. (See text for complete specifications.)

for −60 dB IMD were −4 dBm (0.4 milliwatts) for an output intercept point of +26 dBm (400 milliwatts). Higher output power could be obtained with still higher I_C. Unfortunately, when high feedback and high I_C are used, the noise figure also increases. In this case the noise figure was already about 4 dB for an I_C of 25 mA. Adding more current or a 4:1 output transformer would have resulted in an undesirable increased noise figure and equally undesirable increased gain.

Not being totally content with this amplifier, I tried one of the less expensive (approximately $8.25 each) broadband hybrid amplifiers, a Motorola MWA 130, which exhibits a +19 dBm (95 milliwatts) compression point. For 60 dB IMD, the outputs were +4.77

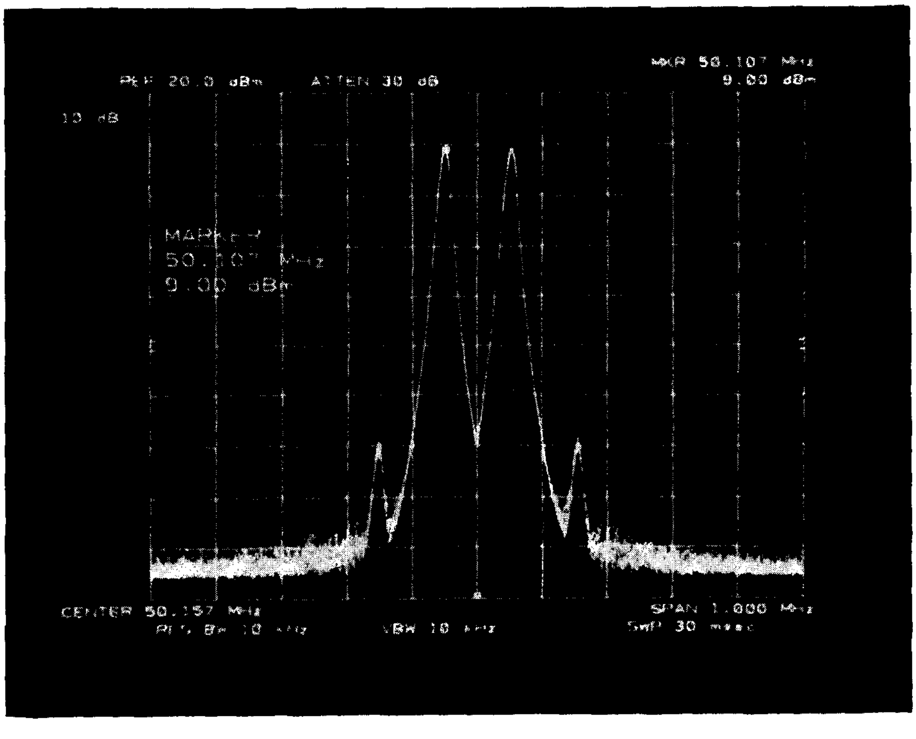


fig. 5. A spectrum analyzer photograph of the output of the high dynamic range amplifier (fig. 6.) with two equal level input signals at approximately 50.1 and 50.2 MHz at 0 dBm each.

amplifier with good input and output VSWR began. A preamplifier was designed around the 2N5109 transistor, a CATV favorite, using shunt and series feedback to obtain a matched input and output impedance (**fig. 4**). The VSWR was less than 1.5:1 from 1 to 70 MHz range while the IMD was

acceptable but only with high (25 mA) I_C (collector current). Typical outputs dBm (3 milliwatts) for an output intercept point of +34.8 dBm (3 watts), a substantial improvement over the home-brew circuit. However, the current drain was 60 mA and the noise figure was about 6.5 dB at 50 MHz,

similar to the performance of the 2N5109 circuit just described, when its current was raised to the same level. Also, the gain — over 15 dB — was too high for this application.

I finally tested one of my favorite preamplifiers, a single transformer lossless feedback type using a common base circuit similar to the one designed by Norton.[6] Although it is more complex to construct, the results are well worth the effort. Using a medium gain (9 dB) configuration, the output power and IMD were outstanding, provided the emitter current was moderate (17 mA). Output compression was typically +20 dBm (100 milliwatts). IMD was down 60 dB for +9 dBm (8 milliwatts) output, for an output intercept point of +39 dBm (8 watts)! The typical IMD versus input and output for this circuit is shown in **fig. 3** and a typical two-tone spectrum display is shown in **fig. 5**. As a bonus, if the preamplifier is properly constructed, the bandwidth is greater than 1.8 to 200 MHz with a 2:1 maximum VSWR and 10-150 MHz for a 1.2:1 VSWR! Truly this was the circuit I was searching for (**fig. 6**).

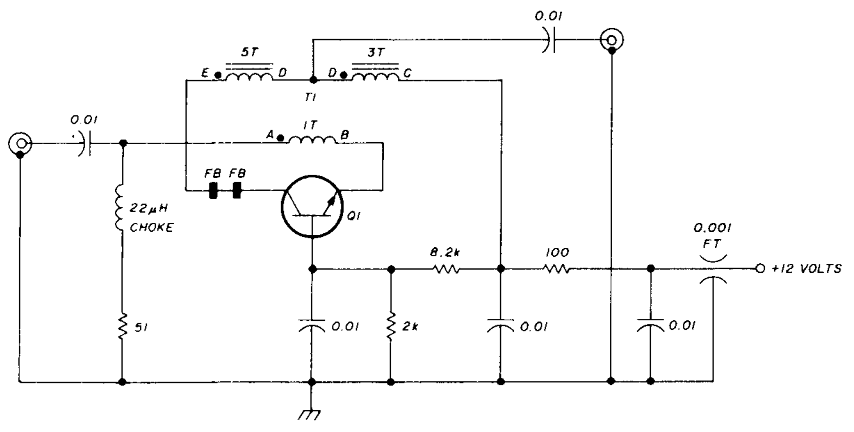
A big key to the success of a high dynamic range preamplifier is the type of transistor chosen. Many RF devices will work well but not always have the same noise figure, bandwidth, or IMD. In the lossless feedback case, the noise figure was typically 1.5 to 2 dB maximum when using the NEC NE41632B transistor, but a 2N5109 had a noise figure of 2.5 to 3 dB in the same circuit. In addition, previous work showed that the most linear transistors were those which were specifically designed for CATV and class "A" linear operation with a *very constant DC current gain (h_{fe}) over a wide range of collector current*. In the CATV business, which is particularly interested in IMD, these devices are frequently referred to as large area multiple emitter structures. The NE41632B and the 2N5109 are both included in this category. (For those who do not have easy access to the NE41632B transistor or the balun core shown in **fig. 6B**, I have made arrangements for PROTO-FAB, 74 Wedgemere Drive, Lowell, Massachusetts 01852, to provide them at a nominal cost. Write them for price and delivery information.)

Caution: This circuit has been modified and has a higher dynamic range than the original Norton circuit. However, his original circuit is patented (U.S. Patent No. 3,891,934, issued June 24, 1975, to David E. Norton and Allen F. Podell). Therefore, any attempt to duplicate this circuit for profit may violate the rights held by the Anzac Division of Adams Russell, Inc.

RF filtering

It goes without saying that high dynamic range cannot be obtained if spurious frequencies or high power out-of-band signals are present in the receiver. Hence RF filtering is very important. It was pointed out in a prior article that the type of input filtering chosen can lessen the chances of destruction from HF signals or lightning entering the first preamplifier of a receiver.[5]

In my August column I discussed the problems of multiple pole filtering such as VSWR distortion and increased losses. Hence it was decided to use a simple low-loss single pole bandpass filter with a pseudo-highpass response at the input to the receiver.[7] In this case a 5 MHz bandwidth was chosen because it would allow reception of 48 MHz European video carriers as well as 52 MHz VK/ZL DX with little degradation at either frequency, but still reject other services. This filter has a nominal insertion loss of 0.75 dB, less than a multi-section type. The schematic is shown in **fig. 7**, with its typical frequency response in **fig. 8**. The input filter chosen doesn't have sufficient out-of-band rejection by itself. Hence a post filter (**fig. 9**) with the same bandwidth (5 MHz) but higher insertion loss (2 dB typical) is required. Its frequency response is shown in **fig. 10**. The filter topography may be somewhat new; it was developed by this author and William K. Talley while at the Mitre Corporation in an effort to obtain a symmetrical attenuation versus frequency response.[7] This filter design is available in computer-aided design form in reference 2. Since this filter is placed after the preamplifier, the loss has a minimal impact on overall system noise figure. As a bonus, the extra insertion loss of the filter will improve the overall system dynamic range accordingly.



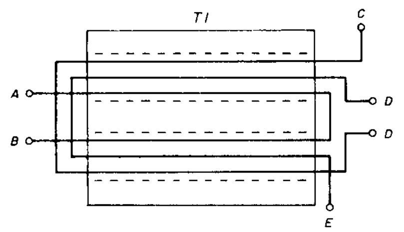
item	description
FB	ferrite bead. Type not critical
Q1	NEC NE41632B or equivalent. (See text for supplier.)
T1	1:3:5 turns ratio transformer wound with No. 32 AWG enameled wire on a Fair-Rite Products type 2843002402 balun core. (See text for supplier.) (Patent warning here.)

fig. 6A. A low-noise wide bandwidth high dynamic range preamplifier using transformer coupled lossless feedback. (See text for further specifications.)

mixers

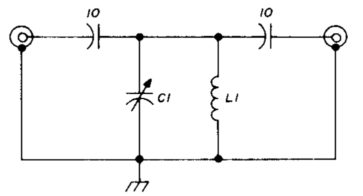
In prior work I had experienced poor dynamic range with the more common Amateur type of mixers such as dual-gate MOSFETs and single-ended bipolar and JFET mixers. However, despite conversion loss, DBMs have always performed very well in my circuit designs.[5,8,9] Hence, when striving for high dynamic range, I decided from my prior experience that DBMs "are the only way to fly." They are easy to drive with a reasonable 50-ohm impedance match at all ports, a goal stated earlier. Also, because of their balanced structure, they tend to cancel any AM present on the local oscillator, a problem which is particularly prevalent if phase-locked LOs are used.

However, when striving for high dynamic range, DBMs must be treated properly. Attenuator pads on the RF and LO ports are a must to terminate undesired mixer generated products and the LO as well as to terminate any



Construction details.
1. First wind one complete turn of No. 32 AWG enameled wire through the balun core from point A to B.
2. Wind three complete turns of the same wire from point C to D. To aid in identification when the transformer is completed, place a small knot in the wire at starting point C.
3. Wind 5 complete turns of the same wire from points D to E.
4. Strip end points D, twist together, and solder.
5. Strip and tin the remaining wires and connect them to the proper point in the circuit. To prevent connecting the transformer incorrectly, leave the small knot on the B+ lead intact until the final wires are soldered in place.

fig. 6B. Balun core is available from PROTO-FAB, 74 Wedgemere Drive, Lowell, Massachusetts 01852, at nominal cost.



item	description
C1	2-20 pF variable capacitor.
L1	approximately 0.28 µH. 9 turns No. 24 AWG enameled wire on Micrometals type T50-10 or equivalent toroid.

fig. 7. A recommended input filter for a 6-meter high dynamic range receive converter. (See text for specifications.)

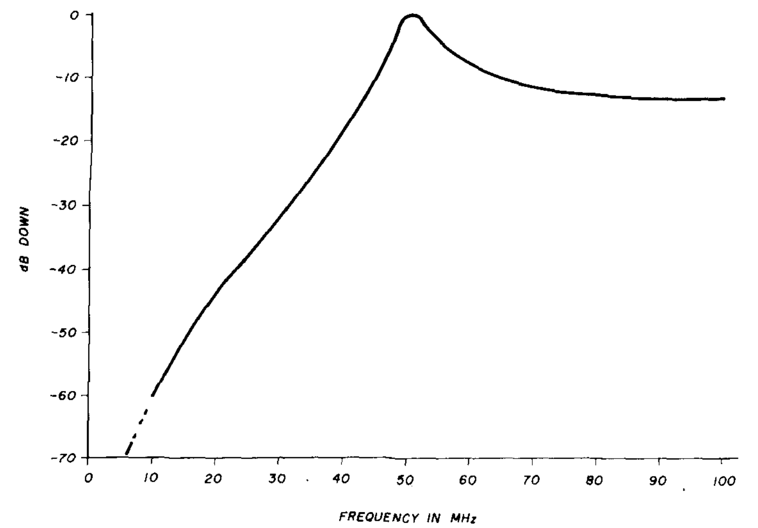
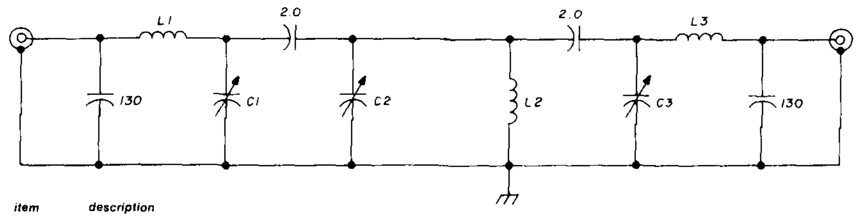


fig. 8. The attenuation-versus-frequency response of the bandpass filter shown in fig. 7.



item	description
C1,2,3	2-20 pF trimmer capacitor.
L1,2,3	approximately 0.42 μH inductor. 11 turns of No. 24 AWG enameled wire on Micrometals type T50-10 or equivalent toroid.

fig. 9. A recommended three-section bandpass filter for a 6-meter receive converter. (See text for specifications.)

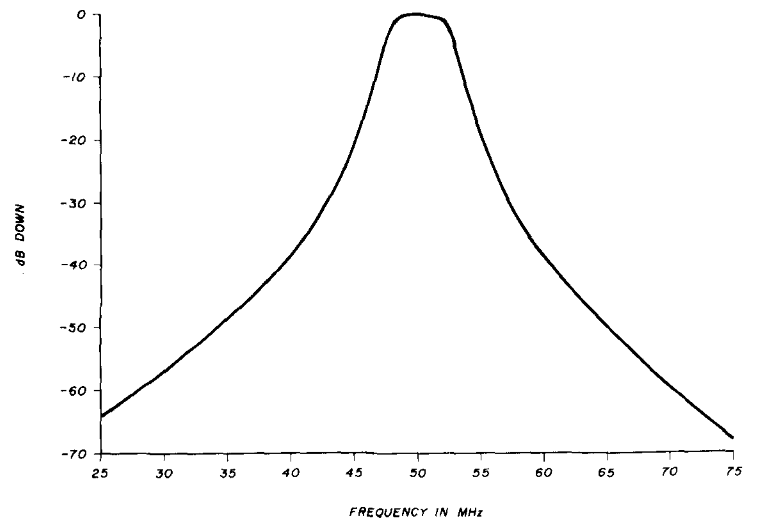


fig. 10. The attenuation-versus frequency response of the 6-meter bandpass filter shown in fig. 9.

in-line filters in their proper impedances.[8] Likewise, a diplexer should be added to the IF output port if low IMD is to be maintained.[8,10]

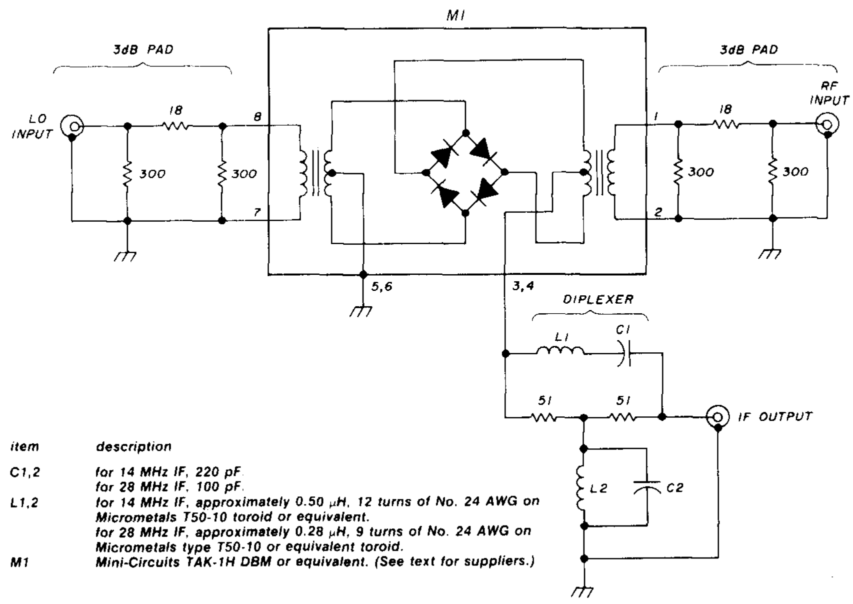
Many DBMs were tested, with the Mini-Circuits Labs TAK-1H selected as the best overall mixer on a cost-versus-performance basis. For comparison, some of the data taken on this and some other popular DBMs are summed up on **table 1**. The final circuit using the TAK-1H is shown in **fig. 11**. This mixer requires higher LO power (+17 dBm or 50 milliwatts) than the more common DBMs usually seen in Amateur equipment, but this is a definite need if high dynamic range is to be obtained. Due to the 3 dB pad on the LO port of the DBM, the LO power required by the overall circuit in **fig. 11** is +20 dBm (100 milliwatts). The RF and LO bandwidth are 2 to 500 MHz. Hence, this circuit has considerably more capability than meets the eye.

A few final remarks on DBMs are in order. Although the so-called high dynamic range mixers (those specified for use with +17 dBm or 50 milliwatts LO), are recommended, the typical DBMs specified with a +7 dBm (5 milliwatts) LO can be used, but with 5 to 10 dB lower LO and dynamic range. Sometimes DBMs are not readily available in single quantity. This can often be handled by getting a group of persons together to buy the minimum quantity. PROTO-FAB, as mentioned earlier, has also agreed to make the TAK-1H DBM available at a reasonable price. Finally, many DBMs are now showing up at flea markets at some pretty good prices, so shop around. You may not find the exact DBM desired but you may be willing to accept slightly lower performance as a compromise. If you adopt the modular approach suggested, it will be easy to upgrade performance at a later date.

local oscillators

My favorite crystal oscillator is the overtone Colpitts.[5,8] The frequency of the LO is determined by the IF chosen, as discussed above. It has great stability and low phase noise, a requirement table 1. Typical measured data on commonly used DBMs. Input signals are at 50, LO at 22, and IF at 28 MHz.

type	LO (dBm)	input compression (dBm)	output intercept (dBm)	approximate cost	quantity
SBL-1	+7	+2	+14	4.50	(10-49)
MD-108	+7	+4	+16	14.00	(1-5)
SRA-1	+7	+4	+16	11.95	(1-49)
SRA-1H	+17	+12	+22.5	17.95	(5-24)
MHP-106	+17	+13	+22.5	45.00	(1-5)
TAK-1H	+17	+16	+28	19.95	(5-24)
MD-139	+20	+17.5	+29	115.00	(1-5)
RAY-3	+23	+16.5	+24	34.95	(4-9)
SAY-1	+23	+20	+32.5	54.95	(1-9)
VAY-1	+27	+24	+36.5	74.95	(1-9)



item	description
C1,2	for 14 MHz IF, 220 pF for 28 MHz IF, 100 pF
L1,2	for 14 MHz IF, approximately 0.50 μH, 12 turns of No. 24 AWG on Micrometals T50-10 toroid or equivalent. for 28 MHz IF, approximately 0.28 μH, 9 turns of No. 24 AWG on Micrometals type T50-10 or equivalent toroid.
M1	Mini-Circuits TAK-1H DBM or equivalent. (See text for suppliers.)

fig. 11. Recommended DBM circuit for a high dynamic range receiving converter with a 14 or 28 MHz IF. (See text for full capabilities.)

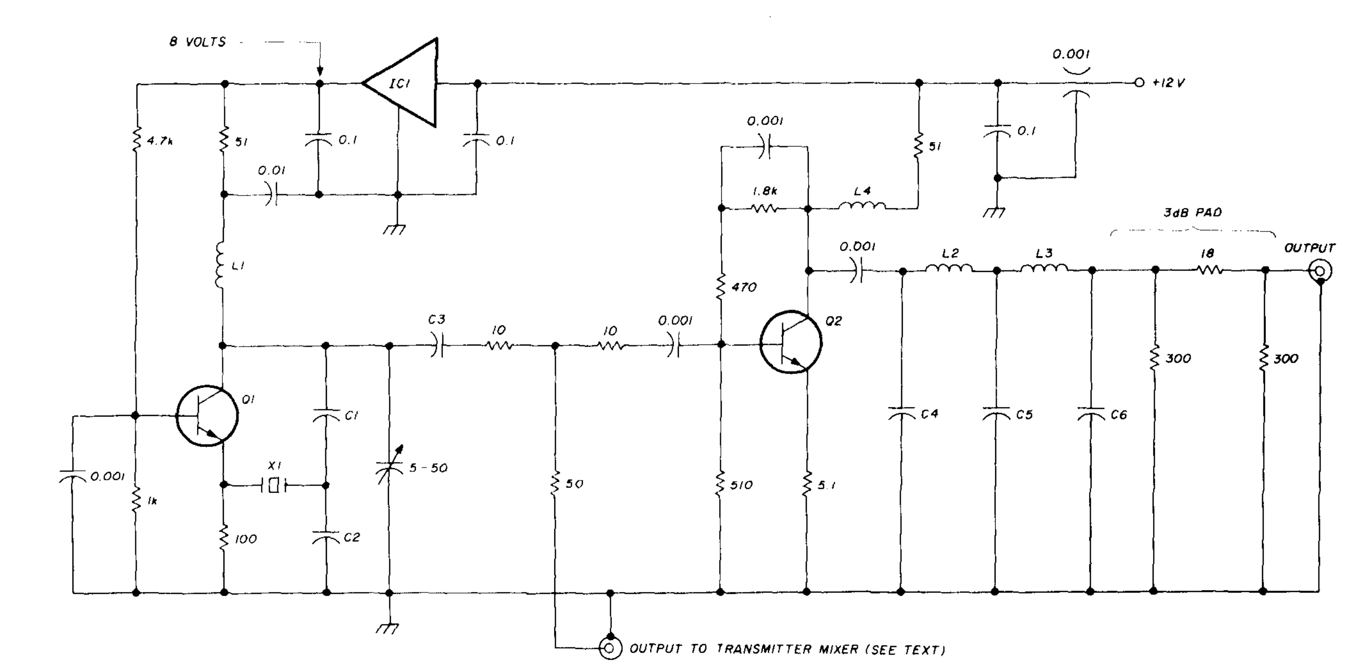
of any high-performance receiver. Phase noise, caused by poor design in the phase lock loops employed, is typically poor in many of the transceivers presently available. Phase noise generates noise on the LO, which, in turn, causes strong signals to be heard several kHz away.

As shown in reference 8, the output of this LO is only about +10 dBm (10 milliwatts). Therefore, an amplifier is required if a high level DBM is used. It was decided not to fight the class "C" type of amplifier, but to go class "A" because there would be improved linearity and less possibility of generating LO noise. Since design of the 2N5109 feedback amplifier had already been completed (**fig. 4**), the bias values were modified slightly for use as the LO amplifier. A simple 1/2 λ low-pass filter followed this amplifier to keep any harmonics from reaching the mixer circuitry.[7]

When the preliminary design was completed for WA8ONQ, it was decided to take the LO output through a two-way power splitter for use on both the receive and transmit mixer. This was more power than required by the transmit mixer, and also dropped the output power substantially on the receive side. So a unique connection was made at the output of the oscillator in conjunction with the attenuator usually used at this point.[8] The result is a secondary output sufficient to drive a standard level DBM (+7 dBm) in a transverter similar to those designs in reference 9. If this output is not needed, terminate it with 51 ohms for possible future use and to insure that the oscillator is seeing the proper match. The final LO schematic is shown in **fig. 12** and delivers +20 dBm (100 milliwatts) at the output connector, the power required by the mixer circuit in **fig. 11**.

postamplification

As already stated, the DBM type of mixer has conversion loss and, therefore, must be followed by a low-noise postamplifier if the system noise figure is to be preserved. The signal levels at this point are about equal to those present at the input to the preamplifier. Therefore, the preamplifier already used is an excellent candidate for this amplifier since it has a good impedance match, low noise figure and high dynamic range. This is also a recommended circuit for HF, 10-meter OSCAR reception, interface with existing VHF/UHF converters, or other applications where moderate gain, low noise figure and high dynamic range are required over a broad bandwidth. In my shack, I have a DPDT coax switch which allows me to bypass the postamplifier when strong signals are present, thereby increasing dynamic range.



item	22 MHz (28 MHz IF)	36 MHz (14 MHz IF)
C1	50 pF	24 pF
C2	200 pF	100 pF
C3	33 pF	20 pF
C4,C6	150 pF	100 pF
C5	300 pF	200 pF
L1	11 turns No. 24 on T50-10 toroid	9 turns No. 24 on T50-10 toroid
L2,L3	10 turns No. 24 on T50-10 toroid	8 turns No. 24 on T50-10 toroid
X1	22 MHz third overtone series mode crystal	36 MHz third overtone series mode crystal
IC1	8 volt, 3 terminal regulator, 78L08 or equivalent	
L4	6.8 μH RF choke	
Q1	2N2222 transistor	
Q2	2N5109 transistor	

(Use a small gear-type heat sink on Q2)

fig. 12. Recommended 22 or 36 MHz LO for a 6-meter high dynamic range receive converter. (See text for output levels.)

IF requirements

At this point it must be obvious that the converter presented has to be followed by a high dynamic range IF receiver if the true capabilities are to be realized. Bypassing the postamplifier or inserting an adjustable attenuator after the receive converter can also help improve dynamic range, but the ultimate limits will probably be limited by the IF receiver. Fortunately, the commercial manufacturers are improving the HF gear that is now being marketed. Furthermore, a converter of the type just described, in conjunction with one of the more modern HF receivers, will definitely outperform any presently available equipment that is solely devoted to the VHF or UHF spectrum.

construction

The 6-meter converter was designed using a modular approach.[5] The final block diagram is shown in **fig. 13.** Each circuit was placed in a cast aluminum box such as the Pomona 2417, Bud CU 123/124 or equivalent with BNC type input/output connectors and feedthrough-type capacitors on the power supply lines. Each box has a piece of double-clad PC board attached to the cover for point-to-point wiring and grounded, as previously discussed in reference 8. The final

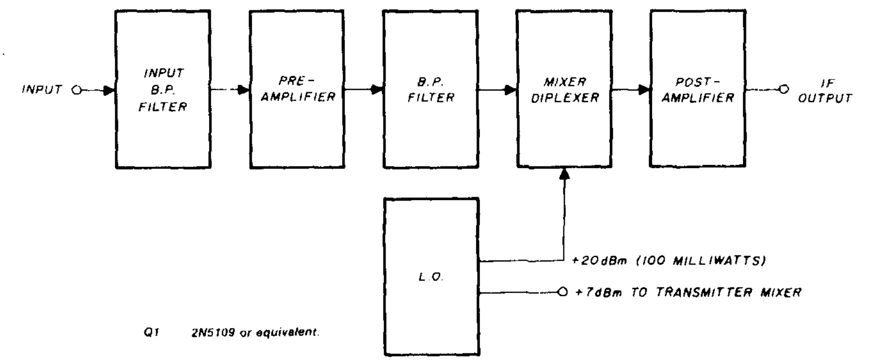


fig. 13. Overall block diagram of a recommended high dynamic range 6-meter converter.

result is a versatile unit with no apparent RF pickup or interaction between modules.

Construction of the circuits is quite straightforward. Leads should be kept short, especially on the filters. High-quality trimmers with low lead inductance such as the ceramic, mylar, or teflon types should be used. The transformer construction hints (**fig. 6B**) for the preamplifier should be followed carefully and the leads on the capacitors, especially on the base bypass, should be kept as short as possible. If a different DBM is used, check the pin designations as some manufacturers use different pin-outs.

tune-up and performance

Very little tuning is required. The input filter can be easily aligned by tuning for maximum signal at 50.1 MHz. The second filter may require more effort. It is best tuned with a sweep setup. However, on the tests I conducted, it was very close to nominal if all capacitors are tuned for maximum signal at 50 MHz in a matched test setup and then inserted in the chain. All the LO requires is to be tuned for maximum output. Properly aligned, the converter described has typically a 5 MHz bandwidth, a gain of about 4 to 6 dB and a noise figure less than 6 dB.

future designs

The state of the art is constantly improving. If a modular approach is used on this receive converter, new or improved circuits can be easily added or changed as prices decrease or parts become available. If you don't change the IF frequency, you'll probably never have to build another LO. If a lower LO power is required, just add an appropriate attenuator on the output. Higher dynamic range mixers are slowly decreasing in price while increasing in performance. **Table 1** can be used as a guide to selection of DBMs.

Finally, our IF receivers must be improved, especially on dynamic range and phase noise. Ultimately, I think that the best receiver will be one that uses a high dynamic range converter directly feeding a narrow bandwidth crystal filter. However, this will require a variable LO and some additional design to prevent phase noise and birdies.

Again, I have rambled on and written a more lengthy column than I intended. However, I feel that the material presented is broad enough in scope and should be worthwhile regardless of frequency. As stated earlier, even using more conventional circuits such as standard level DBMs and JFETs, a substantial improvement can be made over most existing receive converters. After all, the principles discussed are usable at any frequency if time and money are no object!

I hope this material will encourage you and others to try to improve receiver dynamic range and thereby make life more enjoyable. The cost to build such a high dynamic range receive converter is really not that much more than that of a conventional converter. If the dynamic range of the receive circuits is improved, transmitters can be evaluated more effectively. Who knows, you too may find a way to improve these circuits! (Is there any interest in designs for higher bands?)

acknowledgements

I would like to thank Dr. David Norton for his advice and reference material on the lossless feedback amplifier and Jim Reisert, AD1C, for his constructive comments and suggestions on this column. I'd also like to thank Jim Stitt, WA8ONQ, for his encouragement and comments on the performance of the converter circuits as they evolved.

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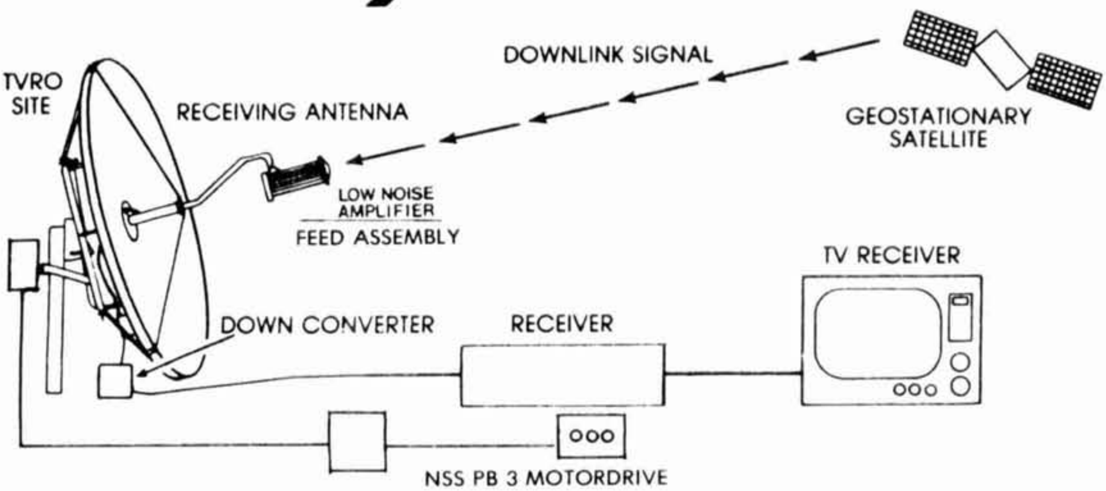

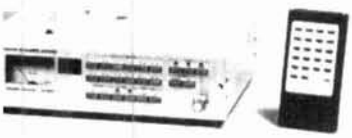

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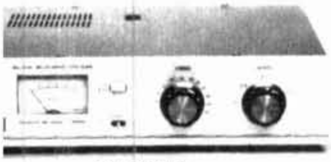

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the half-wave transmission line in bridge measurements

With the increasing availability of inexpensive, accurate RF impedance bridges, Amateurs are discovering how useful such a bridge can be in building and understanding antenna systems. However, unless bridge measurements are taken directly at the antenna, which is generally impractical, the effect of line length has to be taken into consideration when making calculations to determine the actual antenna impedance. This can be accomplished rigorously (by exact mathematical solution)[1] or in this case, simplified as in **eq. 1**. The transmission line is considered lossless, which for short lines is a valid assumption.

$$Z_{ANT} = \frac{Z_{in} - jZ_o \tan \phi}{Z_o - jZ_{in} \tan \phi} \bullet Z_o \quad (1)$$

where Z_{ANT} = antenna impedance
Z_{in} = transmission line impedance measured at bridge terminals
Z_o = characteristic transmission line impedance
ϕ = electrical length of transmission line in degrees

Now as everyone knows, regardless of the antenna or transmission line impedance, measurements made in multiple half-wavelengths repeat — i.e. the same value is seen regardless of whether measurements are taken directly at the antenna terminals or at an electrical half-wavelength along a transmission line. This is where I got into trouble. I measured the antenna impedance through a transmission line that was exactly one-half wavelength (180 degrees) at 3.75 MHz. Then I performed the same measurement (of the antenna) at 4.0 MHz *without* changing the transmission line length. How much difference could that make?

First off, my 180 degree half wavelength line at 3.75 MHz is actually *(4.0)(180)/3.75 = 192 degrees electrical length at 4.0 MHz*. And my measured impedance at the end of the line at 4 MHz was 175 + j 100 ohms impedance.

Substituting the values in **eq. 1**, I found:

$$Z_{ANT} = \frac{175 + j100 - (j)(50)(0.213)}{50 - (j)(175 + j100)(0.213)} \bullet (50)$$

$$Z_{ANT} = 70.8 + j\ 99.7 \text{ ohms impedance}$$

Well, my bridge had measured 175 + j 100 ohms, and my true calculated antenna impedance actually was 70.8 + j 99.7 ohms! That's really a large difference, and yet the electrical length of the line was only 12 degrees longer, or 6.7 percent longer.

So don't be fooled as I was into thinking that if you use an electrical half wavelength line at mid-band, your band end measurements will be close unless you actually correct for the few degrees as I first neglected to do.

Naturally, all antenna measurement calculations could have been done using a Smith chart, but to me the equation shows the impedance relationships involved more clearly.

Although the example was given for the 80-meter band, the same equation can be used for other bands, either for a band center half wavelength line, or as a general equation for use with any length of coax line, as long as you know the electrical length. And after I applied the technique just described, my previously measured data was much more meaningful.

reference

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106	118	130	142	154	166	178	190	202	214	226	238	250	262	274	286	298	310	322	334	346
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109	121	133	145	157	169	181	193	205	217	229	241	253	265	277	289	301	313	325	337	349
110	122	134	146	158	170	182	194	206	218	230	242	254	266	278	290	302	314	326	338	350
111	123	135	147	159	171	183	195	207	219	231	243	255	267	279	291	303	315	327	339	
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IC-751 9-band xcvr/.1-30 MHz rcvr	$1399.00	1199
PS-35 Internal power supply	160.00	144⁹⁵
FL-32 500 Hz CW filter (1st IF)	59.50	
FL-63 250 Hz CW filter (1st IF)	48.50	
FL-52A 500 Hz CW filter (2nd IF)	96.50	89⁹⁵
FL-53A 250 Hz CW filter (2nd IF)	96.50	89⁹⁵
FL-33 AM filter	31.50	
FL-70 2.8 Khz wide SSB filter	46.50	
HM-12 Hand microphone	39.50	
SM-6 Desk microphone	39.00	
CR-64 High stability reference xtal	56.00	
RC-10 External frequency controller	35.00	
MB-18 Mobile mount	19.50	

Options: 720/730/740/745/751	Regular	SALE
PS-15 20A external power supply	$149.00	134⁹⁵
EX-144 Adaptor for CF-1/PS-15	6.50	

ICOM

Options - continued	Regular	SALE
CF-1 Cooling fan for PS-15	45.00	
PS-20 20A switching ps w/speaker	229.00	199⁹⁵
CC-1 Adapt. cable; HF radio/PS-20	10.00	
CF-1 Cooling fan for PS-20	45.00	
EX-310 Voice synth for 751, R-71A	39.95	
SP-3 External base station speaker	49.50	
Speaker/Phone patch - specify radio	139.00	129⁹⁵
BC-10A Memory back-up	8.50	
EX-2 Relay box with marker	34.00	
AT-100 100w 8-band automatic ant tuner	349.00	314⁹⁵
AT-500 500w 9-band automatic ant tuner	449.00	399⁹⁵
MT-100 Manual antenna tuner	249.00	224⁹⁵
AH-1 5-band mobile antenna w/tuner	289.00	259⁹⁵
PS-30 Systems p/s w/cord, 6-pin plug	259.95	233⁹⁵
GC-4 World clock	99.95	94⁹⁵

HF linear amplifier	Regular	SALE
IC-2KL w/ps 160-15m solid state amp	1795.00	1299

VHF/UHF base multi-modes	Regular	SALE
IC-251A* 2m FM/SSB/CW transceiver	$749.00	549⁹⁵

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	Regular	SALE
IC-551D 80 Watt 6m transceiver	$699.00	599⁹⁵
PS-20 20A switching ps w/speaker	229.00	199⁹⁵
EX-106 FM option	125.00	112⁹⁵
BC-10A Memory back-up	8.50	
SM-2 Electret desk microphone	39.00	
IC-271H 100w 2m FM/SSB/CW xcvr	899.00	799⁹⁵
IC-471H 75w 430-450 SSB/CW/FM xcvr	1099.00	Call
PS-35 Internal power supply	160.00	144⁹⁵
PS-15 20A power supply	149.00	134⁹⁵
IC-271A 25w 2m FM/SSB/CW xcvr	699.00	619⁹⁵
AG-20/EX-338 2m preamplifier	56.95	
IC-471A 25w 430-450 SSB/CW/FM xcvr	799.00	699⁹⁵
PS-25 Internal power supply	99.00	89⁹⁵
EX-310 Voice synthesizer	39.95	
HM-12 Hand microphone	39.50	
SM-6 Desk microphone	39.00	

VHF/UHF mobile multi-modes	Regular	SALE
IC-290H 25w 2m SSB/FM xcvr, TTP mic	549.00	489⁹⁵
IC-490A 10w 430-440 SSB/FM/CW xcvr	649.00	579⁹⁵

VHF/UHF/1.2 GHz FM	Regular	SALE
IC-22U 10w 2m FM non-digital xcvr	299.00	249⁹⁵
EX-199 Remote frequency selector	35.00	

Closeout item	Regular	NOW
IC-25H 45w, 2m FM w/up-dn TTP mic	389.00	299⁹⁵
BU-1H Memory back-up	38.50	†10⁰⁰

† BU-1H **$10** purchased with IC-25H, otherwise **$38.50**

	Regular	SALE
IC-27A Compact 25w 2m FM w/TTP mic	369.00	329⁹⁵
IC-27H Compact 45w 2m FM w/TTP mic	409.00	369⁹⁵
IC-37A Compact 25w 220 FM, TTP mic	449.00	399⁹⁵
IC-47A Compact 25w 440 FM, TTP mic	469.00	419⁹⁵
UT-16/EX-388 Voice synthesizer	29.95	
IC-120 1w 1.2 GHz FM transceiver	499.00	449⁹⁵

6m portable	Regular	SALE
IC-505 3/10w 6m port. SSB/CW xcvr	$449.00	399⁹⁵
BP-10 Internal Nicad battery pack	79.50	
BP-15 AC charger	12.50	
EX-248 FM unit	49.50	
LC-10 Leather case	34.95	

Hand-held Transceivers

Deluxe models	Regular	SALE
IC-02A for 2 meters	$ 319.00	289⁹⁵
IC-02AT w/DTMF	349.00	299⁹⁵
IC-04A for 440 MHz	TBA	
IC-04AT w/DTMF	379.00	339⁹⁵

Standard models	Regular	SALE
IC-2A for 2 meters	$ 239.50	214⁹⁵
IC-2AT with TTP	269.50	219⁹⁵
IC-3A for 220 MHz	269.95	234⁹⁵
IC-3AT with TTP	299.95	239⁹⁵
IC-4A for 440 MHz	269.95	234⁹⁵
IC-4AT with TTP	299.95	239⁹⁵

Accessories for Deluxe models	Regular
BP-7 800mah/13.2V Nicad Pak - use BC-35	67.50
BP-8 800mah/8.4V Nicad Pak - use BC-35	62.50
BC-35 Drop in desk charger - all batteries	69.00
BC-16A Wall charger - BP7/BP8	10.00

Accessories for both models	Regular
BP-2 425mah/7.2V Nicad Pak - use BC35	39.50
BP-3 Extra Std. 250 mah/8.4V Nicad Pak	29.50
BP-4 Alkaline battery case	12.50
BP-5 425mah/10.8V Nicad Pak - use BC35	49.50
CP-1 Cig. lighter plug/cord - BP3 or Dlx	9.50
DC-1 DC operation pak for standard models	17.50
LC-2AT Leather case for standard models	34.95
LC-14 Soft case for Deluxe models	17.95
HM-9 Speaker microphone	34.50
HS10 Boom microphone/headset	19.50
HS-10SA Vox unit for HS-10 (dlx only)	19.50
HS-10SB PTT unit for HS-10	19.50
ML-1 2m 2.3w in/10w out amplifier	SALE 79.95
ML-25 2m 2.3w in 20w out amplifier	SALE 179.95
3A-TTN Optional TT Pad - 2A/3A/4A	39.50
SS-32M Commspec 32-tone encoder	29.95

Shortwave receivers	Regular	SALE
R-71A 100 Khz-30 Mhz digital receiver	$799.00	689⁹⁵
FL-32 500 Hz CW filter	59.50	
EX-310 Voice synthesizer	39.95	
RC-11 Wireless remote controller	59.95	
CR-64 High stability oscillator xtal	56.00	
R-70 100 Khz-30 Mhz digital receiver	749.00	599⁹⁵
EX-257 FM unit	38.00	
IC-7072 Transceive interface, 720A	112.50	
FL-44A SSB filter (2nd IF)	159.00	144⁹⁵
FL-63 250 Hz CW filter (1st IF)	48.50	
SP-3 External speaker	49.50	
CK-70 (EX-299) 12v DC option	9.95	
MB-12 Mobile mount	19.50	

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design superhet coilsets
with a microcomputer

Improve tracking with a one-minute interactive program

The design of superhet coilsets to ensure tracking for correct simultaneous tuning of preselector and oscillator circuits has taken on an undeserved air of mystery. Some receiver designers have avoided the problem by resorting to pre-peaked narrow-band circuits (not exactly ideal) or separate tuning of the preselector circuits, a throwback to the 1920s.

The project that led to this article was a receiver for 150-1560 kHz and 2.5-20 MHz in 6 bands with an IF of 1650 kHz. All the parts, including a zero-temperature coefficient (Invar) three-section tuning capacitor, were available. The specifications for the coilset — involving 18 coils, 18 trimmers, and 6 padders — had to be calculated. But how?

the one-minute solution

Years ago I found a set of formulae in the literature that seemed practical to use, although their derivation was not entirely clear to me. Using them, I wound and trimmed coilsets for my home-brew receivers. Because the receivers had worked well I hoped they would also track reasonably well. The problem was that calculating a single set took a whole day using a slide-rule and even after the advent of the pocket calculators, several hours. Half the time was spent in making mistakes and a quarter of the time in wondering if I had discovered all of them. I decided to write a design program based on the existing formulae to produce faster and more reliable results. Written in BASIC, the program reduced the chore to less than a minute!

Thereafter I added a subprogram for plotting the actual tracking curve on the screen. Such a curve shows the difference between the sum of signal and intermediate frequency on the one hand and the oscillator frequency on the other, for an entire tuning range. Ideally, it should be a straight line of zero error value. Without the aid of a computer, this would have taken me many hours to do. With the computer, the task took only a few seconds. As I watched, aghast, the errors ran off the screen as the program kept crashing for all types of designs.

After much thought, I concluded that the algorithms (published in a reliable journal) did not include the effect of stray capacitances such as coil winding capacitances. Even a few pF caused substantial tracking errors, up to 200 kHz or more.

An exhaustive search of the international literature led me to believe that tracking equations are avoided like the plague. Perhaps the subject is thought to be too boring and too difficult for Amateurs. I did find, however, two articles that were *not* over my head, but neither one could pass the computer test! A third article used complex math such as Vieta's theorem and

By Frithjof A.S. Sterrenburg, Westerstraat 47, Sijbekarspel, The Netherlands

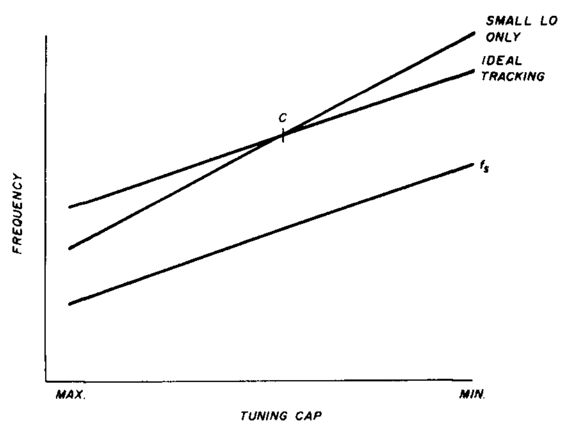

fig. 1. Why superhets require measures to ensure tracking.

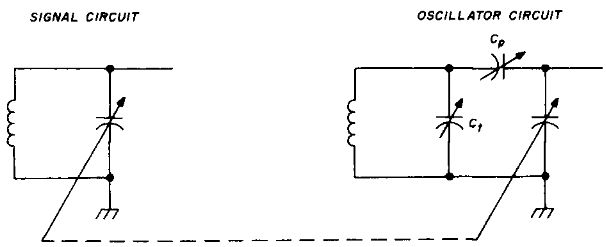

fig. 2. Runge's Remedy: trimmer and padder.

higher order polynomials, but the values given in the examples *did not check* with the formulae supplied! For instance, a trimmer of 46.36 pF in parallel with a tuning capacitor of 20-500 pF was said to give a tuning range of 1.5-5.0 MHz, whereas the actual value would be 1.5-4.3040 MHz. I almost gave up, but finally — in the absence of any information I could both understand and trust — I decided to try to work out the whole thing from the beginning. This is the result, but as you can imagine, I sometimes wonder how many published superhet designs really track.

the principle

In a superhet all signals (f_s) are mixed with a variable oscillator frequency (f_o) to produce a fixed intermediate frequency (f_i). The better solution is to choose an LO frequency higher than the incoming signals $(f_o = f_s + f_i)$ as this reduces spurious responses due to oscillator harmonics. We will consider this case only; the alternative $(f_o < f_s)$ is analagous with the proviso that the signal and oscillator coils are exchanged in the equations. The tuning capacitor will have identical sections; this is the reason tricks are necessary — with different sections for the tuning capacitor good tracking can be obtained, but such components are not readily available. The general equation for resonance is:

$$f^2 = \frac{25330}{L \cdot C} \text{ (MHz, pF and } \mu H) \qquad \textbf{(1A)}$$

and from this it is clear that tuning is not a linear function. If the oscillator coil (LO) were given less inductance than the signal coil L_s to obtain a high f_o, the rate of change of f_s and f_o would never yield a constant difference f_i. **Fig. 1** shows the ideal tracking curve and the error resulting from using a smaller L_o only. If the receiver is made to track near the center of the tuning range (point C), f_o will be too high at the high end and too low at the low end. This could mean attenuation of the signal by 30 dB if you were using potted inductors in the 150-1600 kHz range.

A technique that solves this tracking problem, patented as far back as 1924 by W.T. Runge, is shown in **fig. 2**; a trimmer, C_t curtails the tuning capacitor and has the greatest effect at the high frequency end of the tuning range while the padder capacitor, C_p, does the same thing for the low end. This you know if you have ever aligned a receiver, but it is less widely known that adjustment of C_t and C_p only is not enough. **Fig. 3** shows a situation in which perfect tracking is obtained at both ends of the tuning range with nevertheless a substantial error in the middle, in this case because L_o is too small. Tracking, therefore, requires the determination and adjustment of C_t, L_o, and C_p at *three* frequencies and the curve will then approach a straight line with zero error at three points. Residual errors can be further reduced by shifting the outer two tracking points to the middle which will then result in a curve similar to that in **fig. 4**.

The circuits shown in **fig. 2** represent an idealized case; **fig. 5** represents the actual situation including stray capacitances. As indicated by the vertical lines, a clear distinction should be made between the left half (all circuit elements associated with the coils) and the right half (all capacitances associated with the tuning capacitors and the receiver — for example, input capacitance of the active elements).

circuit elements

Both for practical and theoretical considerations all right half receiver sections must be identical. One important case to be considered is a Colpitts-type oscillator with capacitive divider, indicated by C_x and C_y in **fig. 5**. An equivalent capacitor, C_e, is therefore added to the signal sections, where

$$C_e = (C_x \cdot C_y)/(C_x + C_y) \qquad \textbf{(2)}$$

The winding capacitance of the coils, C_w, is a problem because it is largely unpredictable (depending on

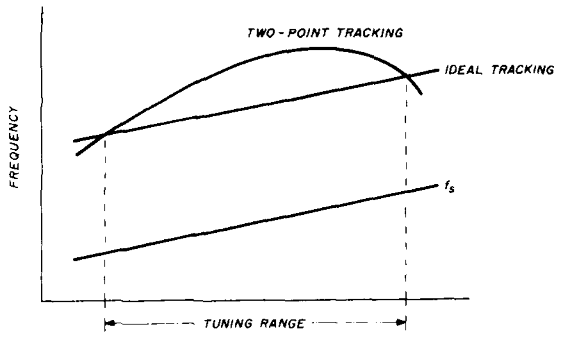

fig. 3. Two-point tracking is not enough!

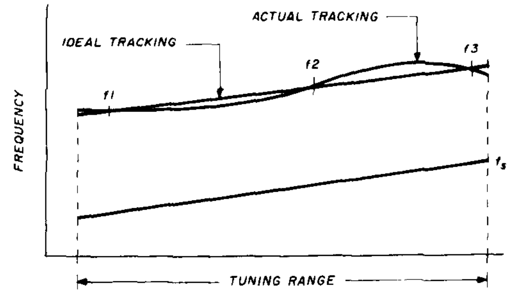

fig. 4. Three-point tracking.

winding technique) and is different for L_s and L_o. I solved the problem by combining C_w with C_t to form "coil capacitances" C_{cs} and C_{co}. The computer will ask you whether that value seems realistic; you can find out by simply winding a trial coil that resonates at the desired frequency with the tuning capacitor used. Then find its self-resonant frequency (without the tuning capacitor) with a dipper and from this derive its C_w. The same applies to the minimum and maximum capacitance of the tuning capacitor, C_{min} and C_{max}, and the wiring stray capacitance C_s. These are the known and unknown values of the circuit elements we shall work with.

calculating the values for the signal coils

These are easy to calculate. They are tuned by the combination on the right, which varies with tuning from a high to a low value:

$$C_H = C_{max} + C_s + C_e \quad (3)$$

$$C_L = C_{min} + C_s + C_e \quad (4)$$

Using the general resonance formula, it is seen that if a tuning range starts at f_L low frequency, an idealized coil without C_{cs} would tune to maximum frequency.

$$f_m = \sqrt{(C_H/C_L)} \cdot f_L \quad (5)$$

Let the desired top frequency by f_H and let $R = f_H/f_L$ then:

$$R = \frac{25330/\sqrt{L_s \cdot (C_L + C_{cs})}}{25330/\sqrt{L_s \cdot (C_H + C_{cs})}} \quad (6A)$$

Therefore: $R^2 = (C_H + C_{cs})/(C_L + C_{cs})$ **(6B)**

or: $C_H + C_{cs} = R^2 \cdot (C_L + C_{cs})$ **(6C)**

and from this it follows that the maximum allowed total capacitance across the signal coil (trimmer plus winding capacitance) can be

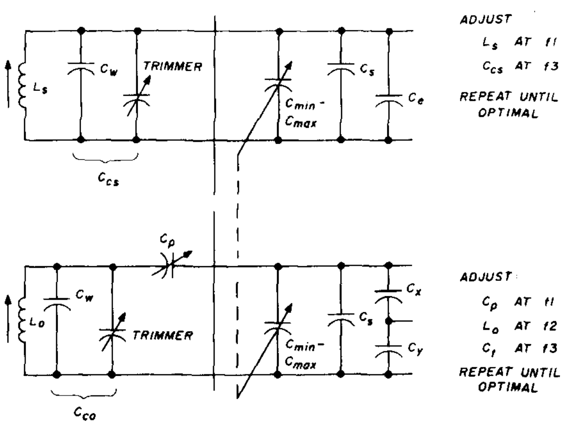

fig. 5. The actual tuned circuits.

$$C_{cs} = (R^2 \cdot C_L - C_H)/(1-R^2) \quad (6D)$$

Allowing a reasonable value for the trimmer, you'll have to determine whether this leaves enough for the winding capacitance, by employing the procedure mentioned above — using a trial coil and dipper — if you're not sure.

example

Design a coil set for the medium wave band that tunes from 520-1620 kHz and incorporates a non-Colpitts oscillator. The tuning capacitor to be used provides from 15-500 pF capacitance and the stray capacitance is 15 pF. C_{cs} when evaluated turns out to be 25.7 pF which is a low value for this range. The medium wave band, with a frequency ratio of over 1:3 requires an approximate 1:10 capacitance ratio and this leaves little leeway for strays.

The value of the signal coil is:

$$L_s = 25330/f_L^2 \cdot C_H \quad (1B)$$

and in this example $L_s = 173.25\ \mu H$

The signal circuits are now completely defined; additional preselector circuits must be identical.

determining the tracking points

One tracking point must be at the center of the range. This can be found by taking the geometric means of the two end frequencies: $\sqrt{f_L \cdot f_H}$. The outer tracking points must be shifted slightly toward the center. The amount will vary with the tuning range. Empirically, the following factor gives good results:

$$Q = 1 + (f_H/40 \cdot f_L) \quad \textbf{(7)}$$

You can easily change this factor in the program and observe the result on the display. The greatest deviation will occur at the high end, which is acceptable because the bandwidth of the signal coils is also a maximum there. The tracking (and trim-) points are:

$$f_1 = f_L \cdot Q \text{ (low end)} \quad \textbf{(8)}$$

$$f_2 = \sqrt{f_L \cdot f_H} \text{ (center)} \quad \textbf{(9)}$$

$$f_3 = f_H/Q \text{ (top end)} \quad \textbf{(10)}$$

which calculates to be 560, 918, and 1503 kHz, respectively in the example outlined earlier. As the operator tunes to these frequencies, he will set the tuning capacitor to values that can be calculated from the resonance formula after subtracting C_{cs}. These values (for the right half of the signal sections) are:

$$C1 = \frac{25330}{f_1^2 \cdot L_s} - C_{cs} \quad \textbf{(11)}$$

$$C2 = \frac{25330}{f_2^2 \cdot L_s} - C_{cs} \quad \textbf{(12)}$$

$$C3 = \frac{25330}{f_3^2 \cdot L_s} - C_{cs} \quad \textbf{(13)}$$

In the oscillator section, the total capacitance to the right will be the same as above (C_s and C_e are identical), so with the three capacitances $C1$, $C2$, and $C3$ the oscillator must tune to $f_1 + f_i$, $f_2 + f_i$ and $f_3 + f_i$.

determining the oscillator circuit elements

Taking into account the total coil capacitance for the oscillator circuit and the padder, gives the following three equations:

$$f_1 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_1}{C_p + C_1} \cdot L_o} \quad \textbf{(14)}$$

$$f_2 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_2}{C_p + C_2} \cdot L_o} \quad \textbf{(15)}$$

$$f_3 + f_i = 25330/\sqrt{C_{co} + \frac{C_p \cdot C_3}{C_p + C_3} \cdot L_o} \quad \textbf{(16)}$$

The solution to these equations means that the tracking error is indeed made equal to zero at these points.

Let's tackle the padder first. Define the ratio $(f_1 + f_i)/(f_3 + f_i)$ as "A" and call $(f_2 + f_i)/(f_3 + f_i)$ as "B," then by dividing **eq. 14/eq. 16** we find that:

$$C_{co} = \frac{A^2 \cdot C_p \cdot C_1/(C_p + C_1) - (C_p \cdot C_3)/(C_p + C_3)}{1 - A^2} \quad \textbf{(17)}$$

Similarly, division of **eq. 15** by **eq. 16** yields:

$$C_{co} = \frac{B^2 \cdot C_p \cdot C_2/(C_p + C_2) - (C_p \cdot C_3)/(C_p + C_3)}{1 - B^2} \quad \textbf{18)}$$

These two equations can be used to solve for C_p:

$$C_p = \frac{C_1C_2(B^2 - A^2) + C_1C_3(A^2B^2 - B^2) + C_2C_3(A^2 - A^2B^2)}{C_1(A^2 - A^2B^2) + C_2(A^2B^2 - B^2) + C_3(B^2 - A^2)}$$

(This looks neater when the terms in parentheses are called X, Y, and Z respectively, as in the program.) C_{co} is found by entering C_p in **eq. 17** or **18** and then:

$$L_o = \frac{25330}{(f_2 + f_i)^2 \cdot \left(C_{co} + \frac{C_p \cdot C_2}{C_p + C_2}\right)} \quad \textbf{(19)}$$

This completes the coilset design. In this example: $C_p = 585.7\ pF$, $C_{co} = 41.9\ pF$, $L_o = 84.66\ \mu H$ for $f_i = 450\ kHz$.

tracking curve

Plotting the tracking curve on the screen depends on the graphic capability of your microcomputer. High resolution plotting in assembly language is definitely not necessary: what you want to see is the general trend and the peak errors, not an accurate graph. The method used in plotting the tracking curve is described below.

First the tuning range is divided into as many equal sections as the micro has columns. Because of rounding off, the range must be extended above f_H; otherwise, the plot sometimes won't reach this value. For a 40-column machine and half a column extra margin the command would be:

for f = f_L to f_H + (f_H − f_L)/80 step (f_H − f_L)/40

For each f calculate the total capacitance across the signal coil:

$$C = 25330/(L_s \cdot f^2) \quad \textbf{(1C)}$$

So the capacitance to the right of the lines is equal to $C_v = C - C_{cs}$. From this we can calculate the oscillator frequency correspnding to f:

$$f_o = \sqrt{\frac{25330}{L_o\left(C_{co} + \frac{C_p \cdot C_v}{C_p + C_v}\right)}} \quad \textbf{(20)}$$

Then print $f_o - f - f_i$ together with f for a tracking table or use $f_o - f - f_i$ as the row- and f as the

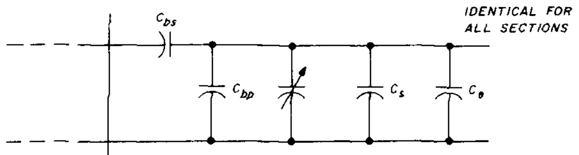

fig. 6. Series and parallel bandspreading.

column parameter. To avoid crashing from incorrect row data, first display the table and examine the size of the error values, then enter a corrective multiplication factor via an INPUT command if necessary.

For this example the errors (rounded off) were −3.2, +4.26, −6.97 and +8.5 kHz. This indicates excellent tracking and implies that everything is correct — an assurance I never had before.

the program

The program listing in **table 1** was written for the TI-99/4A which is suitable for this application because of its upper and lower case capability, which makes for easy reading, its wide choice of variables, and high degree of precision in mathematics. Note that when the program runs, very small values are subtracted from other very small values so that you may have to use double precision. I have left out everything that is not essential and used the simplest BASIC commands for easy translation. One version I have draws all coils and capacitors in HiRes and inserts their values next to them for added finese. The listing takes less than 4k of memory, and can still be adapted for your particular machine. A printer is not necessary; the program warns you to "NOTE" the values displayed.

bandspreading

As supplied, the program already includes parallel bandspreading, because parallel capacitance is added in the form of C_{cs} by your independent choice of f_H. But let's look at bandspreading in more detail (**fig. 6**).

Parallel bandspreading with C_{bp} improves the tuning curve. Ordinary tuning capacitors provide logarithmic coverage, with compression at the high end. For straight-line frequency coverage you need a tuning capacitor with pointed plates (the BC-221 has one), but these are exceptionally rare. Adding C_{bp} reduces the top end compression. The disadvantage of adding parallel capacitance is that the L/C ratio of the tuned circuit is lowered and Q is reduced, especially at the high end. You may assume that the effect on the Q is acceptable if C_{bp} is smaller than:

$2 \cdot \{C_L + \textit{reasonable value for trimmer (e.g. 15 pF)} + \textit{value found for } C_w\}$

Series bandspreading C_{bs} raises the L/C ratio, but

table 1. Program listing for superhet coilset design.

```
100 CALL CLEAR
110 PRINT "  SUPERHETERODYNE COILSETS": : : : : :"   for Ham Radio Magazine":
: :"by Frithjof A.S. Sterrenburg": : : :"          ** 1984 **": : : : : : :
120 FOR DELAY=1 TO 1000
130 NEXT DELAY
140 CALL CLEAR
145 REM for colpitts type oscillator
150 INPUT "does the oscillator have a  capacitive divider? y/n":A$
160 IF A$="y" THEN 170 ELSE 230
170 INPUT "enter Cx,Cy ":CX,CY
180 CE=(CX*CY)/(CX+CY)
190 CALL CLEAR
200 PRINT "an extra parallel condenser Ce of";CE,"pF will be specified for thesi
gnal sections in the finalprint-out"
210 FOR DELAY=1 TO 1000
220 NEXT DELAY
230 CALL CLEAR
235 REM define tuning range
240 INPUT "enter minimum capacitance oftuning condenser (pF) ":CMI
250 PRINT : :
260 INPUT "enter maximum capacitance oftuning condenser (pF) ":CMA
270 PRINT : :
280 INPUT "enter estimated stray capa- citance in pF ":CS
290 PRINT :"Ce calculated earlier will  be printed at the end if it was required
": :
300 CL=CMI+CS+CE
310 CH=CMA+CS+CE
320 INPUT "enter lower frequency of tu-ning range in MHz ":FL
330 PRINT : :
340 FM=SQR(CH/CL)*FL
350 PRINT "the tuning range would be";FL;"to";FM;"MHz without trimmer or wind-in
g capacitance": :
360 INPUT "enter desired top frequency of tuning range ":FH
370 R=FH/FL
380 CCS=(CH-(R^2*CL))/(R^2-1)
390 CALL CLEAR
400 PRINT "                NOTE:": :"total desired coil capacitor(trimmer + coil capa
citance)=";CCS;"pF": :
410 INPUT "does this seem feasible? y/n ":B$
420 IF B$="n" THEN 360
430 CALL CLEAR
440 CL=CL+CCS
450 CH=CH+CCS
460 LS=25330/(FL^2*(CH))
465 REM total specification of signal coils
470 PRINT "                NOTE:": :"Ls=";LS;" microH": :"assumed stray capacitance=
";CS;"pF": :"trimmer + coil capacitance= ";CCS;"pF": :
480 PRINT "for all preselector sectionsadd extra parallel capacitorof";CE;"pF(Ce
)": :
490 INPUT "enter c to continue ":C$
500 IF C$="c" THEN 510 ELSE 490
510 Q=1+(FH/FL)*0.025
520 F1=FL*Q
530 F2=SQR(FL*FH)
540 F3=FH/Q
550 CALL CLEAR
555 REM determine the tracking points and oscillator components
560 PRINT "                NOTE:": :"low trimpoint=";F1;" MHz":"mid trimpoint=";F2;"
MHz":"high trimpoint=";F3;" MHz": : : :
570 C1=(25330/(F1^2*LS))-CCS
580 C2=(25330/(F2^2*LS))-CCS
590 C3=(25330/(F3^2*LS))-CCS
600 INPUT "enter intermediate frequencyin MHz ":FI
610 LET A=((F1+FI)/(F3+FI))
620 LET B=((F2+FI)/(F3+FI))
630 X=B^2-A^2
640 Y=(A^2*B^2)-B^2
650 Z=A^2-(A^2*B^2)
660 CP=((X*C1*C2)+(Y*C1*C3)+(Z*C2*C3))/((Z*C1)+(X*C3)+(Y*C2))
670 CCO=(A^2*(CP*C1)/(CP+C1)-(CP*C3)/(CP+C3))/(1-A^2)
680 LO=25330/((F3+FI)^2*(CCO+((CP*C3)/(CP+C3))))
690 CALL CLEAR
695 REM total specification oscillator components
700 PRINT "                NOTE:": :"padder=";CP;"pF": :"trimmer including winding
capacitance=";CCO;"pF":

710 PRINT "(for signal coil";CCS;"pF was allowed)": :"Lo=";LO;"microH": :"capaci
tive divider";CX;"and";CY;"pF": :
720 INPUT "enter 1 for next set,2 for  tracking table,3 for graph, 4 to end ":K
730 ON K GOTO 140,740,820,990
740 CALL CLEAR
745 REM   printout of tracking errors for range,for 30 frequencies
750 FOR F=FL TO FH+(FH-FL)/60 STEP (FH-FL)/30
760 C=25330/(LS*F^2)
770 CV=C-CCS
780 FO=SQR(25330/(LO*(CCO+((CP*CV)/(CP+CV)))))
790 PRINT F*1000,(F+FI-FO)*1000
800 NEXT F
810 GOTO 720
820 CALL CLEAR
825 REM plot of tracking curve for 24 lines/30 columns
830 INPUT "for error plot enter multi- plication factor,maximum de-viation=10 ":
M
840 CALL CLEAR
845 REM   draw plotting grid (for 99/4A)
850 CALL CHAR(128,"101010101FF101010")
860 PRINT TAB(16);"factor=";M: :
870 CALL HCHAR(13,1,128,31)
880 CALL VCHAR(1,16,128,24)
890 Y=1
895 REM  for 30-columnn machine this is identical to 750-780 inclusive and can b
e a gosub
900 FOR F=FL TO FH+(FH-FL)/60 STEP (FH-FL)/30
910 C=25330/(LS*F^2)
920 CV=C-CCS
930 FO=SQR(25330/(LO*(CCO+((CP*CV)/(CP+CV)))))
940 X=(F+FI-FO)*1000*M
945 REM plot asterisks
950 CALL HCHAR(12+X,Y,42)
960 Y=Y+1
970 NEXT F
980 GOTO 720
990 END
```

tuning at the top end is even more compressed. Series and parallel bandspread techniques can, of course, be combined. For series bandspreading, a series capacitor C_{bs} is added to the right hand side and this is identical for all tuning capacitor sections. In fact, we construct a new tuning capacitor with different C_{min} and C_{max} and then run the program as normally.

To determine C_{bs} for a desired range we define new constants. Let $C_{cs} = j$, desired range $k = f_H/f_L$, $m = C_H$, $n = C_L$ and the unknown bandspread capacitor = b. Then starting with the equation for the tuning range:

$$k^2 = \frac{j + \dfrac{b \cdot m}{b + m}}{j + \dfrac{b \cdot n}{b + n}} \qquad \textbf{(21A)}$$

you will arrive at this:

$$(b \cdot m) \cdot (b + n) - (b + m) \cdot [\{k^2 \cdot n + (k^2 - 1) \cdot j\} \cdot b + (k^2 - 1) \cdot j \cdot n] = 0 \qquad \textbf{(21B)}$$

Now let's call $k^2 \cdot n + (k^2 - 1) \cdot j = p$ and $(k^2 - 1) \cdot j \cdot n = q$, then:

$$b^2 \cdot m + b \cdot m \cdot n - (q \cdot b^2 + p \cdot b + q \cdot m \cdot b + p \cdot m) = 0 \qquad \textbf{(21C)}$$

$$\text{or } b^2 + \frac{m \cdot n - p - q \cdot m}{(m - q)} \cdot b - \frac{p \cdot m}{(m - q)} = 0 \qquad \textbf{(21D)}$$

If we call the two fractions "r" and "s" respectively, this is the common equation $b^2 + r \cdot b - s = o$. One root is negative, the other is the value of the bandspread capacitor:

$$C_{bs} = \frac{-r + \sqrt{r^2 + 4 \cdot s}}{2} \qquad \textbf{(22)}$$

Although all bandspread capacitors have been lumped with the right hand half of the circuits in this derivation, they are switched for other ranges and will therefore be *physically present in the left hand half* (in the coil cans, for instance). In the end, you can properly combine all the capacitances calculated.

conclusion

Although any coilset can now be designed in a few minutes, the main value of this project was something else. For the first time, I could plot tracking curves, and these showed that all the literature I could lay hands on gave unreliable formulas! The microcomputer freed me from the nightmare of calculating, so I could begin to *think* and rediscover a bit of neglected theory for myself. A microcomputer may have its practical uses, but above all, it's a powerful tool for learning!

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LS13	.50	LS193	.80
LS14	.50	LS194	.65
LS15	.32	LS195	.65
LS20	.27	LS196	.75
LS21	.28	LS197	.85
LS27	.28	LS221	.80
LS30	.30	LS240	1.00
LS32	.38	LS241	1.00
LS33	.50	LS242	1.00
LS37	.33	LS243	1.00
LS38	.34	LS244	1.25
LS42	.50	LS245	1.50
LS51	.24	LS251	.50
LS54	.25	LS253	.55
LS55	.27	LS257	.55
LS73	.35	LS258	.55
LS74	.45	LS259	2.50
LS85	.75	LS260	.50
LS86	.39	LS266	.50
LS90	.50	LS273	1.25
LS93	.55	LS279	.45
LS97	2.00	LS280	1.50
LS107	.37	LS283	.60
LS109	.39	LS290	.85
LS112	.39	LS293	.85
LS122	.45	LS298	.89
LS123	.75	LS299	1.60
LS124	2.75	LS323	2.60
LS125	.50	LS348	.75
LS126	.49	LS366	.45
LS132	.50	LS367	.50
LS133	.55	LS368	.40
LS138	.65	LS373	1.30
LS139	.65	LS374	1.35
LS151	.50	LS375	1.00
LS153	.50	LS377	1.49
LS154	1.75	LS378	.85
LS155	.50	LS390	1.19
LS156	.55	LS393	1.19
LS157	.60	LS399	1.25
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7428	.25	74162	.60
7432	.27	74163	.60
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7451	.20	74173	.75
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7474	.40	74175	.80
7483	.45	74185	1.70
7485	.55	74192	.70
7486	.30	74193	.80
7490	.35	74195	.80
7493	.33	74273	1.75
7496	.40	74367	.60
74107	.28	74390	1.40
74109	.45		

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11.088	1.59
12.000	2.75
13.440	1.00
14.31818	2.49
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16.00000	1.50
16.5888	1.49
17.430	2.49
18.2259	1.00
20.000	3.75
21.87108	1.00
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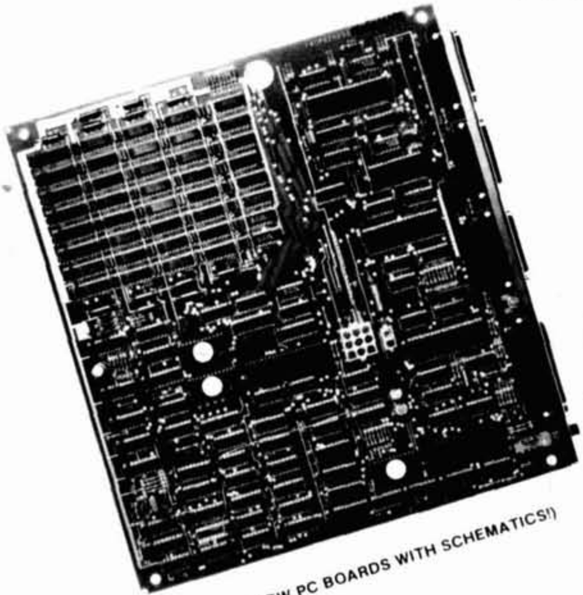

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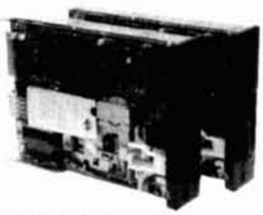

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receiver sweep alignment system

No sweep generator? Try these handy throwaway circuits

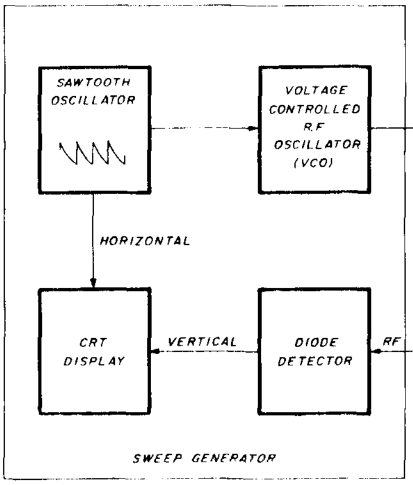

fig. 1. Sweep generator; the CRT (Cathode Ray Tube) displays relative amplitude of the swept filter output.

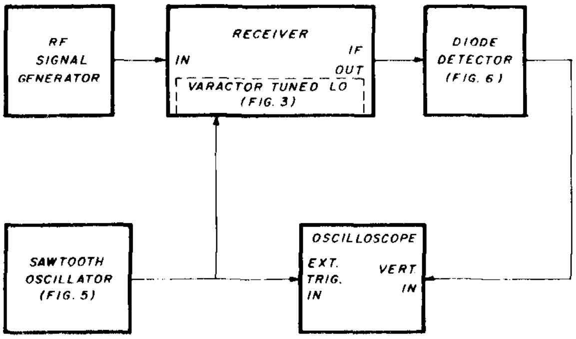

fig. 2. Minimal configuration sweep system uses the receiver LO (Local Oscillator) as the VCO (Voltage Controlled Oscillator).

When a circuit just "doesn't sound right," the obvious solution is to use a sweep generator to evaluate and align the filters. But a sweep generator may not be available; for most of us, it's too expensive a piece of equipment for the occasional use it receives.

This article describes how I approached this problem in the development of an SSB receiver by adding a few extra circuits during construction, then removing and discarding them after use. A separate signal generator and oscilloscope were also required.

sweep measurement basics

Fig. 1 illustrates a typical sweep measurement system. The sawtooth oscillator generates a voltage "ramp" which tunes the voltage controlled oscillator frequency across the filter passband. For small frequency changes, the voltage controlled oscillator often uses a varactor ("varicap") diode to change circuit capacitance. The amplitude of the RF/IF signal coming through the filter varies with (and helps define — **Ed.**) the filter's frequency response, and is detected by the diode detector. The detector output is displayed by the oscilloscope vertical channel while the sawtooth oscillator drives the oscilloscope horizontal channel in step. The resulting display plots the filter's amplitude versus frequency response. This display is used to align filters since it gives an instant indication of circuit adjustment results. This is very handy when a large number of interrelated circuit manipulations must be made.

The proposed minimal sweep alignment system uses the existing receiver local oscillator as the voltage controlled oscillator. This is similar to a panoramic receiver with a much smaller sweep range. **Fig. 2** shows this scheme as implemented in an HF SSB receiver project. Three new circuits are added; a varactor tuning diode, a diode AM detector, and a sawtooth oscillator circuit.

detailed circuit description

Because the three circuits were designed to be dis-

By Cliff Klinert, WB6BIH, 1126 Division Street, National City, California 92050

carded after use, careful consideration was given to parts availability.

The familiar 1N4000-series of silicon rectifiers make good "varactor" diodes when biased in the linear region. **Fig. 3** shows the varactor tuning circuit using a 1N4007 connected to a typical 5 MHz oscillator tank circuit. The circuit was tested by applying an adjustable DC bias voltage and measuring the corresponding frequency with a counter. **Fig. 4** shows the results of this experiment for DC bias voltages from 14 to almost 28 volts. *Note that the curve is almost linear.* Good linearity throughout the sweep system is required to provide an undistorted picture of the filter's passband response.

The sawtooth oscillator, the heart of the system, generates a periodic linearly increasing voltage ramp waveform. If the oscilloscope used in this project had provided a sweep output connection, the external oscillator might not have been required. The varactor circuit requires a *linear sawtooth voltage providing a 14* to 24 volt ramp per frequency sweep. Simple sawtooth

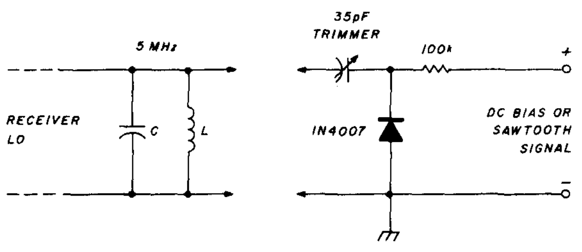

fig. 3. The varactor diode acting as a voltage variable capacitor changes the resonant frequency of an LC circuit in step with a varying DC bias.

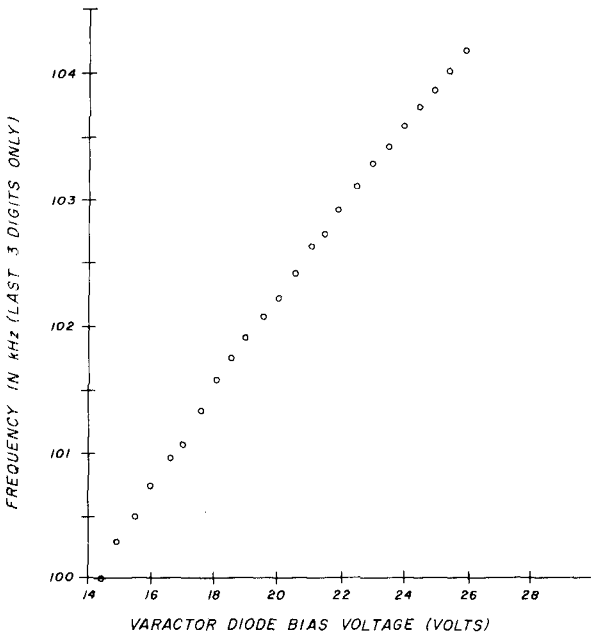

fig. 4. Varactor diode tuning curve is a plot of actual data taken from the circuit of fig. 3. The range of the swept frequency is adequate for the SSB filter tested.

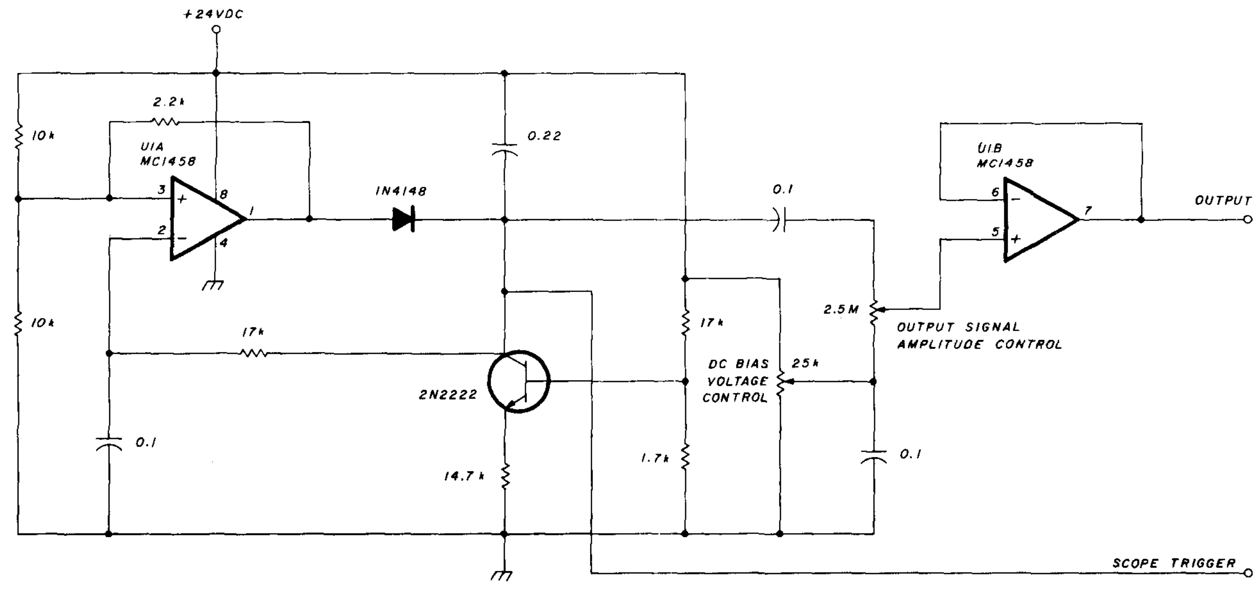

fig. 5. The sawtooth oscillator uses an inexpensive IC (MC1458) to both generate the ramp waveform and buffer its output. Mylar capacitors are used for best results.

oscillator circuits are normally designed around unijunction transistors and in this case, to keep costs down, a less expensive IC, an MC1458, (U1 of **fig. 5**), was used. The 0.22 μF capacitor is charged by a 2N2222 constant current source until the MC1458 "triggers" and briefly shorts the capacitor back to the positive supply voltage line. The cycle is then immediately repeated. Since the 2N2222 collector current is determined by its base and emitter bias circuit, it is nearly constant while charging the 0.22 μF capacitor. This constant current ensures a linear capacitor voltage rise with time. The other half of the MC1458, U1B, is used as a voltage follower and provides a low impedance, higher current version of the sawtooth voltage at its output. The 2.5 Megohm potentiometer controls the amount of frequency excursion, while the 25 kilohm potentiometer tunes the voltage controlled oscillator frequency.

The diode detector provides a DC voltage that is proportional to the RF swept signal output amplitude. A hot carrier diode such as a 1N5711, shown in **fig. 6A**, can be used. If a hot carrier diode is not available, an inexpensive germanium diode that is slightly forward biased can be used in its place as shown in **fig. 6B**. A DC return is required for the *biased* diode, and the diode impedance decreases as the current in-

fig. 7. Oscilloscope photographs. Unless otherwise noted, the sweep speed is 5 milliseconds per division and vertical sensitivity is 0.1 volt per division. The unit under test is a 9.0 MHz SSB IF amplifier using an MC1350 and surplus crystal filter.

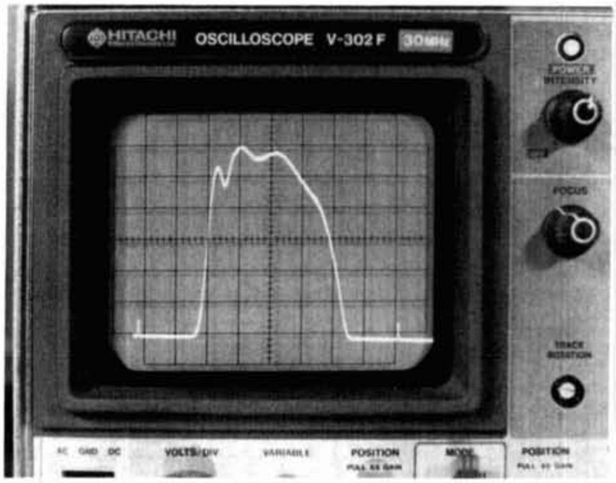

fig. 7A. This shows a typical frequency sweep about 3.5 kHz wide.

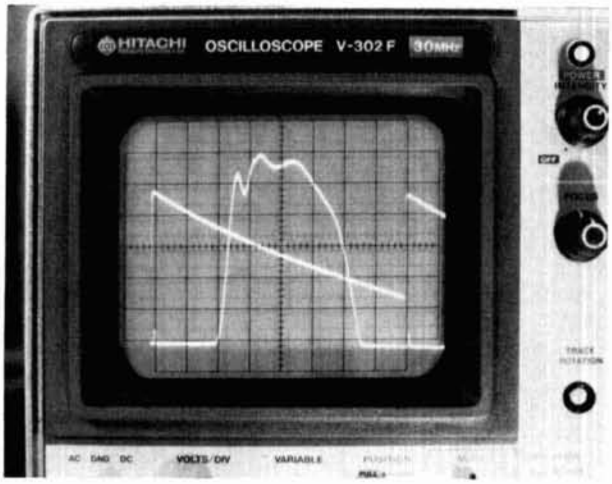

fig. 7B. The second channel is used to display the sawtooth signal showing its relationship to the frequency sweep.

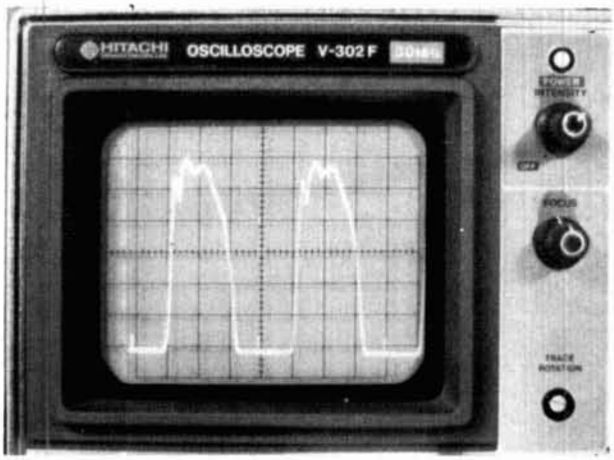

fig. 7C. The sweep speed is too low (10 milliseconds per division). This indicates that the speed should be adjusted to provide the desired display.

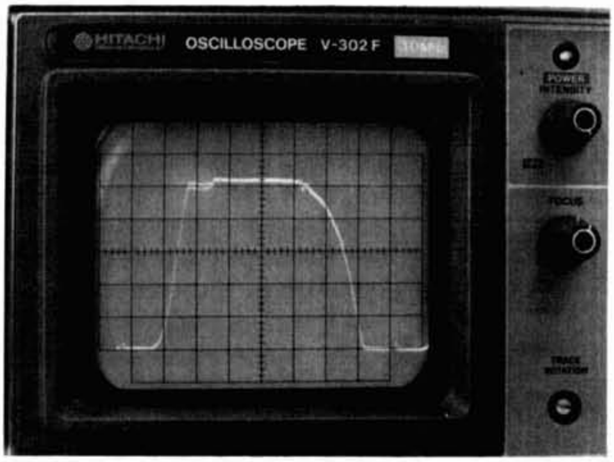

fig. 7D. The IF amplifier is "flat topping." The gain or input signal must be reduced to remove the distortion at the top of the display.

creases. A high impedance tuned circuit will be "shorted out" by the biased diode. Adjusting the diode bias control will produce a sharp peak in the detected signal output. Reducing the control's resistance near zero will, of course, destroy the diode. A silicon diode such as a 1N914 works very poorly as a detector.

construction and installation proceed smoothly

The circuit requires 24 volts DC which can be provided by either a bench power supply or batteries. Connect the varactor circuit (**fig. 3**) to the receiver local oscillator. Apply a variable DC bias voltage and adjust the varactor-tuned circuit to achieve results similar to those in **fig. 4**. Make a notation of the voltage variation required to obtain the full frequency sweep.

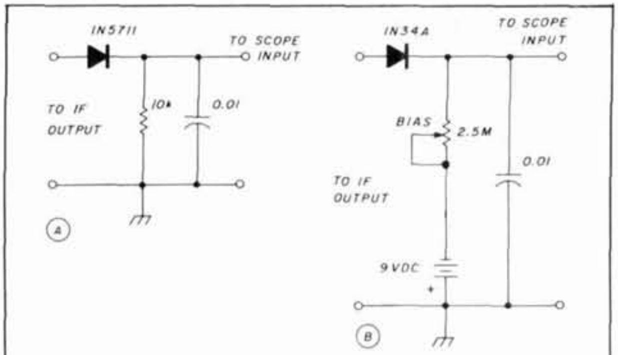

fig. 6. The diode detector is connected between the IF or filter output coupling link and the oscilloscope vertical input. The detector can also be used at the audio output.

Assemble the sawtooth oscillator (**fig. 5**) and check the results with the oscilloscope. The output voltage should be capable of a swing nearly equal to the power supply limits with a period of approximately 50 milliseconds. The circuit can be assembled on a circuit board, but for temporary use, just solder the components together by their leads on the bench. Stray noise pickup may be a problem, but the large signal makes this unlikely. The varactor diode circuit is quite sensitive because of its high impedance, but no problems were encountered when connected to the sawtooth oscillator output.

Connect the detector probe (**fig. 6**) to the receiver IF amplifier output at a point of maximum available signal. Some experimentation is necessary with a signal generator and oscilloscope to obtain the maximum detected output without saturating the IF amplifier.

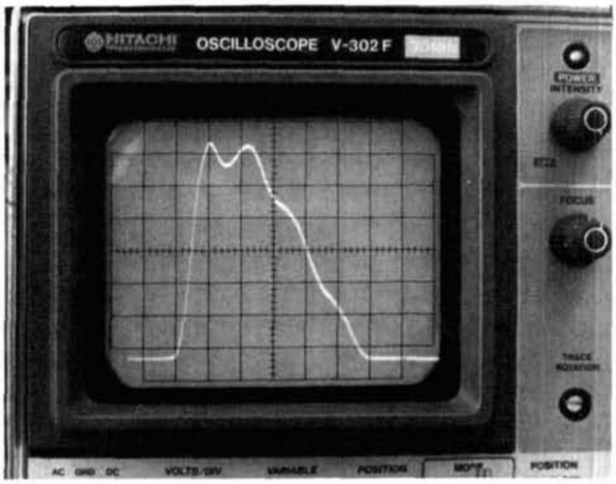

fig. 7E. Crystal filter load impedance is too low, resulting in a noticeable loss of audio frequency response and confusion in setting the BFO frequency.

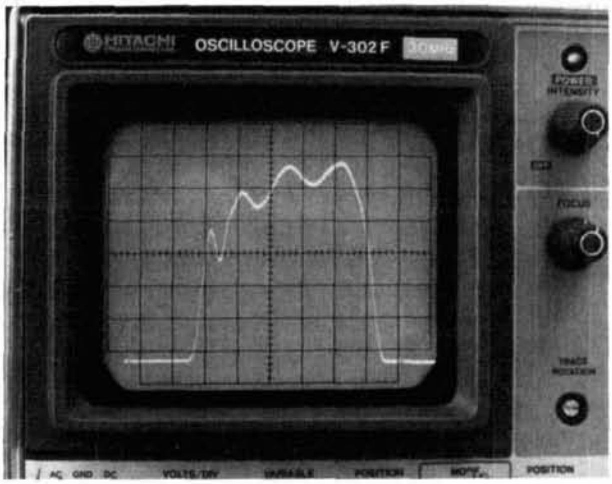

fig. 7F. The crystal filter load impedance is high, about 6000 ohms.

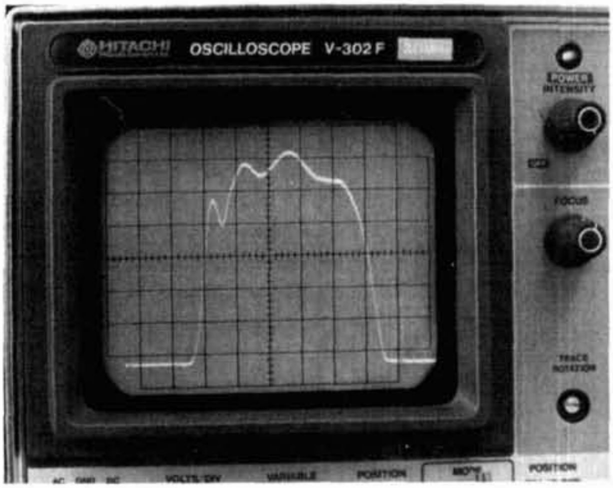

fig. 7G. The best response was obtained with approximately 3000 ohms load impedance. This is an unusually high load impedance for a crystal filter, and may indicate some problem in the filter.

The oscilloscope must be DC coupled with at least 0.1 volt per cm sensitivity.

Connect the sawtooth oscillator output to the varactor circuit. It's easier to start with a relatively low sweep amplitude when finding the frequency, so reduce the sweep signal amplitude in the beginning. Connect the diode detector output to the oscilloscope vertical input and synchronize the oscilloscope sweep from the sawtooth oscillator signal. The sawtooth signal could also be connected to the oscilloscope horizontal amplifier input if access is available. Tune the receiver to the signal generator frequency and make adjustments as required. Slowly increase the sawtooth signal amplitude until a sweep display indication is obtained. **Fig. 7** shows the results obtained from an SSB crystal filter sweep. In this case, the crystal filter output impedance was varied with a potentiometer in series with the filter output. The photographs show IF sweeps, but similar results were obtained by connecting the detector to the outputs of the product detector and audio amplifier. This provides analysis of other points in the receiving system that would be useful for troubleshooting or design evaluation.

conclusions

This article presented the concept of using expendable circuits as built-in test equipment for use during project construction. A handy sawtooth oscillator was presented for those who collect simple circuits for afternoon projects. This oscillator will find many applications in oscilloscope or spectrum analysis projects. No construction details were presented, since the concept was to show that circuits can be assembled without circuit boards for prototype or temporary use and discarded later.

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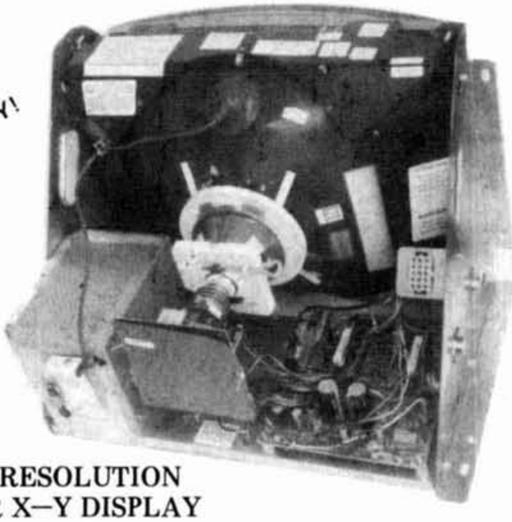

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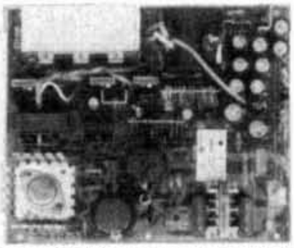

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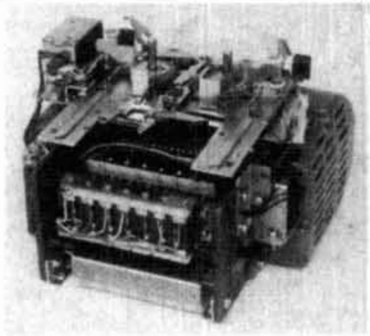

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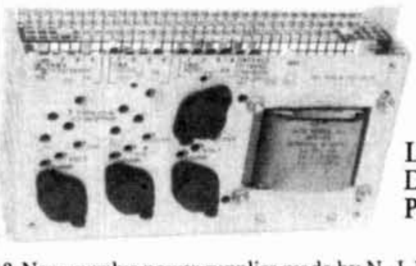

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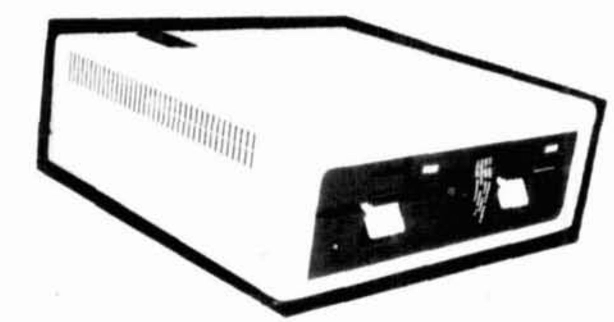

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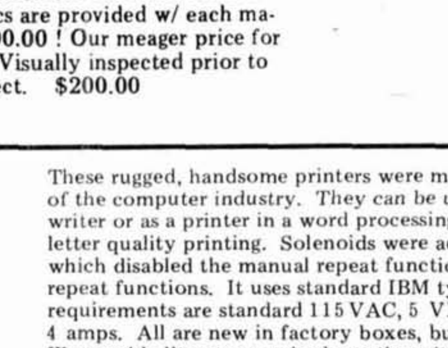

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MM8000	1.15	MRF905	2.55	PT6709	25.00	SD1080-7	7.50
MM8006	2.30	MRF911	2.50	PT6720	25.00	SD1080-8	6.00
MM8011	25.00	MRF965	2.55	PT8510	15.00	SD1080-9	3.00
MPSU31	1.01	MRF966	3.55	PT8524	25.00	SD1084	8.00
MRA2023-1.5	42.50	MRF1000MA	32.77	PT8609	25.00	SD1087	15.00
MRF134	10.50	MRF1004M	31.05	PT8633	25.00	SD1088	22.00
MRF136	16.00	MRF2001	41.74	PT8639	25.00	SD1088-8	22.00
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MRF221	10.00	MSC1720-12	225.00	PT8709	20.00	SD1095	15.00
MRF223	13.00	MSC1821-3	125.00	PT8727	29.00	SD1098-1	30.00
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MRF227	3.45	MSC2001	30.00	PT8742	19.10	SD1109	18.00
MRF230	2.00	MSC2010	93.00	PT8787	25.00	SD1115-2	7.50
MRF231	10.00	MSC2223-10	245.00	PT8828	25.00	SD1115-3	7.50
MRF232	12.07	MSC2302	POR	PT9700	25.00	SD1115-7	2.10
MRF237	3.15	MSC3000	35.00	PT9702	25.00	SD1116	5.00
MRF238	13.80	MSC3001	38.00	PT9783	16.50	SD1118	22.00
MRF239	17.25	MSC72002	POR	PT9784	32.70	SD1119	5.00
MRF245	35.65	MSC73001	POR	PT9790	56.00	SD1124	50.00
MRF247	31.00	MSC80064	35.00	PT31083	20.00	SD1132-1	15.00
MRF304	36.00	MSC80091	10.00	PT31962	20.00	SD1132-4	12.00
MRF306	50.00	MSC80099	3.00	PTX6680	20.00	SD1133	9.50
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MRF314	29.21	MSC80758	POR	RE3789	25.00	SD1134-1	2.50
MRF315	28.86	MSC82001	33.00	RF35	16.00	SD1134-4	12.00
MRF316	55.43	MSC82014	33.00	RF85	17.50	SD1134-17	12.00
MRF317	63.94	MSC82020M	130.00	RF110	21.00	SD1135	10.25
MRF412	18.00	MSC82030	33.00	S50-12	23.80	SD1135-3	12.00
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MRF422	38.00	MSC83005	70.00	S3031	22.00	SD1143-1	10.00
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SD1214-7	5.00	SD1311	1.00	SD1454-1	48.00	SRF2092	50.00	SD1272	10.95	SD1428	24.00	SD1545	33.00	SS4006	25.00
SD1214-11	5.00	SD1317	8.00	SD1477	35.00	SRF2147	22.00	SD1272-1	10.95	SD1428- 6084	12.00	SD1546H1	55.00	SS4152	15.00
SD1216	12.00	SD1319	2.50	SD1478	21.00	SRF2225	15.00	SD1272-2	10.95	SD1429-2	15.00	SD1561	79.00	TA7686	15.00
SD1219-4	15.00	SD1345-6	5.00	SD1480	53.00	SRF2264	25.00	SD1272-4	10.95	SD1429-3	14.90	SD1574-1	6.95	TA8559	15.00
SD1219-5	15.00	SD1347-1	1.00	SD1484	1.50	SRF2265	100.00	SD1278	13.75	SD1429-5	15.00	SD1575	6.95	TA8561	15.00
SD1219-8	15.00	SD1365-1	2.50	SD1484-5	1.50	SRF2281	5.00	SD1278-1	13.75	SD1430	12.00	SF4557	25.00	TA8562	15.00
SD1220	8.00	SD1365-5	2.50	SD1484-6	1.50	SRF2371	15.00	SD1278-5	13.75	SD1430-2	18.00	SK3048	5.00	TA8563	15.00
SD1220-1	9.50	SD1375	7.50	SD1484-7	1.50	SRF2347	50.00	SD1279-1	18.00	SD1434	28.00	SL501-59	15.00	TA8564	15.00
SD1220-9	8.00	SD1375-6	7.50	SD1488	22.85	SRF2356	38.00	SD1279-3	18.00	SD1434-5	28.00	SL501-173	15.00	TA8894	15.00
SD1222-8	16.00	SD1379	15.00	SD1488-1	28.00	SRF2378	16.00	SD1281-2	8.00	SD1434-9	28.00	SM7714	5.00	TIS189	3.55
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SD1232	4.00	SD1410	18.00	SD1528-1	24.00	SRF2821	25.00	SD1290-7	15.00	SD1446	4.03	SRF1005	50.00	V4101E	20.00
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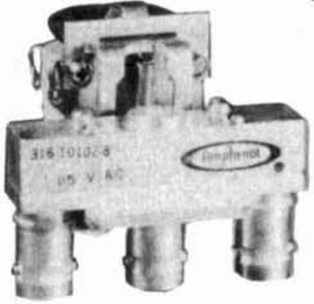

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DELIVERY: Orders are usually shipped the same day they are placed or the next business day, unless we are out of stock on an item. The customer will be notified by post card if we are going to backorder the item. Our normal shipping method is UPS or U.S. Mail depending on size or the weight of the package. Test Equipment is shipped only by air and is freight collect, unless prior arrangements have been made and approved.

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SALES TAX: ARIZONA residents must add 6% sales tax, unless a signed ARIZONA resale tax card is currently on file with us. All orders placed by persons outside of ARIZONA, but delivered to persons in ARIZONA are subject to the 6% sales tax.

SHORTAGE OR DAMAGE: All claims for shortages or damages must be made within 5 DAYS of receipt of parcel. Claims must include a copy of our invoice, along with a return authorization number which can be obtained by contacting us at (602) 242-8916 or sending a post card. Authorizations cannot be on our 800 number. All items must be properly packed. If items are not properly packed make sure to contact the carrier so that they can come out and inspect the package before it is returned to us. Customers which do not notify us within this time period will be held responsible for the entire order as we will consider the order complete.

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More Details? CHECK—OFF Page 158

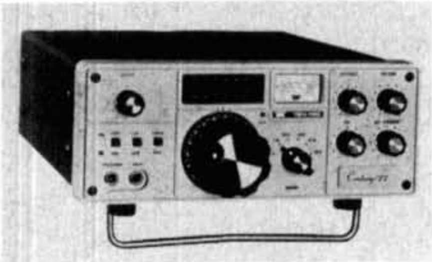
✓ 209

NEW products

the Century 22

Ten-Tec has announced the return of the Ten-Tec Century transceiver. The Century 22 is a 50-watt, 6-band CW transceiver that features a variable audio filter, automatic gain control, an SWR bridge, automatic level control, and an electronically switched "S" meter.

The Century 22 measures 4 × 10 × 10.5 inches (25 × 101 × 29 cm), weighs 6 pounds (2.7 kg), and is priced at $389.

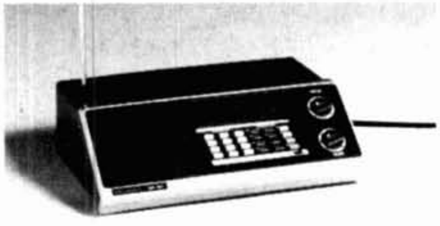


For information, contact Ten-Tec, Inc., Sevierville, Tennessee 37862.

7-band scanning radio

Heath Company has introduced the only kit-built scanner to cover aircraft, marine and public service bands, *all in one unit*. The GR-740 40-Channel Scanning Radio covers all seven UHF/VHF radio bands, scans 40 user-selected frequencies and provides direct access to any frequency in the seven bands.

The 24-key keyboard is divided into program and operate sections for simplified operation. Forty different channels (frequencies) are easily programmed into the two 20-channel memory banks. Either bank can be scanned at five or 15 channels per second; the GR-740's search can be programmed or changed at the touch of a button. A priority channel can be sampled every two seconds, with interruption when a signal is detected.



Patented track tuning permits receiving frequencies across the full band without adjustments; circuitry is automatically aligned to each monitored frequency. A large digital, front-panel display shows the channels and features selected. All circuit boards are factory-assembled and pre-aligned to ensure that even the first-time kit builder can build and operate one of the world's best scanning radios, with a minimum of time and at a substantial savings.

For more information about the GR-740 40-Channel Scanning Radio, contact Heath Company, Dept. 150-315, Benton Harbor, Michigan 49022.

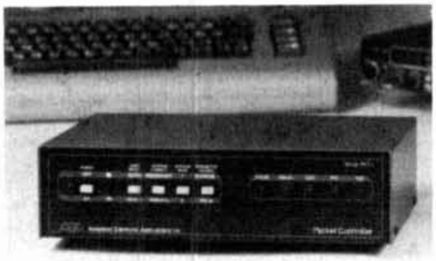
Circle #301 on Reader Service Card.

TAPR packet radio controller

Advanced Electronic Applications, Inc. has announced the introduction of the Model PKT-1, packet radio controller, through an arrangement with Tucson Amateur Packet Radio, Inc. (TAPR), Tucson, Arizona. While the end user price is $589.95, Amateur Radio operators can take advantage of a discounted price of $499.95 through participating AEA dealers.

The PKT-1 is a packaged and warranted version of the well-known do-it-yourself TAPR kit board with version 3.1 software. The purchase price includes application assistance and a year's conditional warranty.

Packet Radio is a burst mode of data or text transmission utilizing AFSK, FSK, or PSK modulation. On VHF it runs at 1200 Baud typically and uses CRC error checking, ensuring an extremely low error rate. Multiple users may share a simplex or duplex channel simultaneously on a timeshare multiplexed basis.



Any packet station using the PKT-1 may operate as a store-and-forward repeater (Digipeater) for someone else's transmission while concurrently functioning as a regular packet station. Up to 8 Digipeating stations may be used between two terminal stations. Digipeating allows routing the transmission path around physical obstacles blocking a line-of-sight radio path and allows extending the link beyond line-of-sight distances.

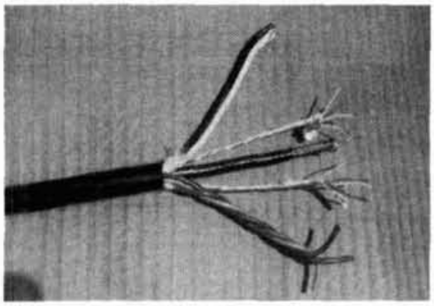
For detailed information, contact Advanced Electronic Applications, Inc., P.O. Box C2160, Lynnwood, Washington 98036-0918

Circle #304 on Reader Service Card.

NEW products

TVRO cable

Nemal Electronics International of North Miami, Florida, has introduced a new addition to its line of direct burial combination cable for use in TVRO installations. Nemal type-4 satellite control cable is the first combination cable available to the satellite industry containing an RG-6/U, 18 gauge, 95 percent copper shielded signal cable. SCC-4 also contains two conductors of 12 gauge, three conductors of 18 gauge, three conductors of 20 gauge shielded plus drain wire, and three conductors of 22 gauge shielded plus drain wire.



All Nemal satellite control cables utilize a patented direct-burial polyethylene jacket as well as tinned copper drain wires. Nemal also offers a complete line of over 500 types of cable, connectors, and SMATV products.

For additional information, contact Nemal Electronics International, Inc., 12240 N.E. 14th Avenue, North Miami, Florida 33161.

Circle #303 on Reader Service Card.

outdoor scanner antenna

Hamtronics, Inc. has announced a new antenna for scanner and monitor buffs. The compact ACT-1 Power Antenna, which may be installed easily on the side of a house, outside a window, in an attic, etc., without any special masts or brackets, is a broadband whip antenna with a low-noise preamplifier in its base. Although smaller than a full-size outdoor antenna (only 25 inches tall), the ACT-1 provides good coverage of distant signals and often outperforms larger antennas because of its active booster amplifier. A low-noise microwave transistor in the preamp provides excellent results from 30 MHz right up through the new 800-MHz band, and covers low-band, high-band, and UHF.

The ACT-1 Power Antenna is mounted to any flat vertical surface with four wood screws. The 50-foot cable plugs directly into the "antenna" and "12V" jacks on the rear of most scanner radios. If your particular scanner doesn't have

NEW products

a 12V terminal, a simple 12 VDC plug-in adapter is available.

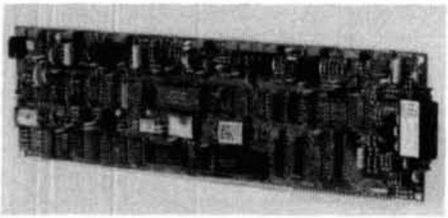
The price of the ACT-1 Power Antenna is only $79 plus $3 for shipping and handling.

For more information, contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535.

Circle #305 on Reader Service Card.

RTTY interface

HAL Communications RTTY Personal Computer Interface, PCI-2000, is a real RTTY modulator/demodulator, not a "computer compromise." All three shifts (170-425-850) transmit and receive. The full "103 type" modem for up to 300 baud may be set for either FDX (answer or originate) or HDX (either set on tones). The "202



type" modem may be used HDX up to 1200 baud, and may be jumper-selected for either "COM1" or "COM2" operation. Compatible with existing PC communications software (external DAA required for phone line connections, the PCI-2000 offer the multimode features of the CT2100 and CT2200 plus companion software that includes features such as split screen, TX and RX buffers, HERE IS storage, and disk message storage and retrieval. Of course, all PCI-2000 modes are set with the Personal Computer's FN keys.

For further information, contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801.

Circle #302 on Reader Service Card.

150-MHz mini-catalog

Sinclair Radio Laboratories has issued a new mini-catalog describing its line of 150-MHz products, which includes base station antennas, transmitter combiners, duplexers, receiver multicouplers and ferrite isolators.

Featured in this line-up is the Q-Circuit Base Station Duplexer, Model Q-201G, a six-cavity unit that provides high attenuation at close frequency separations in the 132-174 MHz band. Its Q-Circuit design provides 100 dB isolation at 300 kHz spacing with 50 dB mid-band isolation.

For a copy, contact Sinclair Radio Laboratories, 122 Rayette Road, Concord, Ontario, Canada L4K 2G3.

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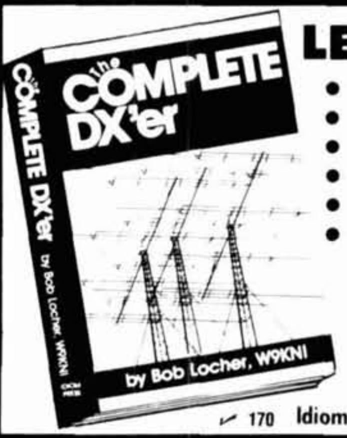

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A4....289.95
402CD....289.95
R3....279.95
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G7-144....119.95
Butternut HF6V...80 thru 10 vertical............125.00
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70CMCV-7..70cm vertical collinear.............39.00
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All B & W Dipoles....less 10%
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HG52SS....999.00
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2M22C....119.95
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VALOR mobile antennas 75-10M, ea............20.00
AVANTI ASP151 3G thru the glass 2M............33.00
ANTECO 2M...5/8 MAG MOUNT, compl. 25.00
METZ SW-1 SWL ANT..50khz to 54mhz............59.95

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ROHN 50' tower consisting of 4 sections of 25G, 1 section of 25AG-2 or 25AG-3....269.00
ROHN FK2548....prepaid....799.00

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45G....per section....107.00
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PL259 831SP silverplate....1.25
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4400 N Male-SO239....6.00
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8261 N Male....3.00
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CLOSEOUT CORNER-SOME GOOD DEALS IN HERE
We plan to feature things we "found" in our warehouse. If you ever saw the warehouse you would understand! This month's "FINDS" are:
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AEA MT-1P as above with nicad battery......50.00
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DRAKE TR7/R7 RX Cable....20.00
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✓ 179

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This package allows full MORSE, BAUDOT RTTY, ASCII RTTY and AMTOR operations.
AEA CP-1....retail....239.95
AEA TI-1....retail....119.95
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TOTAL $489.80
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AEA TI-1 tuning indicator....retail....119.95
AEA AC-1 12 volt ps....retail....4.95
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HAL HAL HAL HAL HAL HAL HAL
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MICROLOG - We have a few ACT-1 units available and need to move them. NEW and USED....CALL

NEEDLESS TO SAY, BUT THESE ARE SPECIAL PRICES. THE PRICES WILL BE GOOD UNTIL DECEMBER 15TH. ALL PRICES ON OVER-STOCK ITEMS GOOD ONLY UNTIL SUPPLIES ARE GONE. TAKE ADVANTAGE OF THESE DEALS NOW!!!!

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RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

COLLINS KWM-380 with all optional interfaces. Perfect $2495. WA6NWP, Sid, 209-642-3363. 52632 RD 426, Oakhurst, CA 93644.

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***ATLAS 350XL** owners group. Send QSL card with s/n your* rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

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ENJOY SATELLITE TELEVISION. Save money with easy, guaranteed, do-it-yourself antenna plans/kits. Electronic knowledge not necessary. Send $1.00 for catalog or $8.95 for "Consumer Guide to Satellite Television". GFI-41, Box 9108, Missoula, MT 59807.

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DIGITAL AUTOMATIC DISPLAYS for FT 101, and TS 520 (functional DG-5 replacement) Collins, Drake, Swan, Heath, and most others. 6½" digits. Write for information. Grand Systems, P.O. Box 2171, Blaine, WA 98230 (604) 530-4551.

WANTED: *Cash paid for used speed radar equipment.* Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

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WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

R L DRAKE COLLECTION 2A receiver $100, 2B receiver & 2BQ $125, 2C receiver & 2CQ $135, TR-3 Xcvr and AC-3 $175, 2NT transmitter x/10 Novice crystals $85. All completely reconditioned, re-tubed and guaranteed. Plus UPS Ev Taylor, W7BYF, 2921 Loyola Dr., Davis, CA 95616 (916) 756-7372 eves.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

FOX-TANGO Newsletters — Since 1972, the prime source of modifications, improvements, and repair of Yaesu gear, free to Club members. Calendar year dues still only $8 U.S., $9 Canada, $12 elsewhere. Includes five year cumulative index by model numbers, or send $1 for index and sample Newsletter. Fox Tango Club, Box 15944, W. Palm Beach, FL 33416.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

WANTED 220 MHZ FM TRANSCEIVER Prefer Xtal controlled WØBJ, 2801 Wright, North Platte, NE 69101.

IMRA International Mission Radio Assn. helps missioners — equipment loaned; weekday net, 14.280 MHz, 2-3 PM Eastern. Br. Frey, 1 Pryer Manor Rd., Larchmont, NY 10538.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details, W8URR, 1025 Wildwood Road, Quincy, MI 49082.

VIC-20 PHONE PATCH. Build your own simplex autopatch for less than $50 using your own transceiver and VIC-20 or Commodore 64. For full documentation and program cassette tape, send $20 to: KIE Enterprises, P.O. Box 72, Running *Springs*, CA 92382 (714) 867-7120.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

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PRINTED CIRCUIT DRILLS: sizes 53, 60, 68 and 70 thru 76, $1.25 each, 10 assorted $9.95. Western Electic transformers $1.50 each. Two Way Radio, 437 Payne Drive, Endicott, NY 13760.

Coming Events

ACTIVITIES

"Places to go..."

MASSACHUSETTS: The 35th annual New England DXCC Dinner, November 10, Concord Lodge of Elks, Baker Avenue, West Concord. Starts 2 PM with a variety of DX talk and slide *programs. Admission $2.00. Cocktail hour 6 PM followed by* family-style dinner starting at 7:30 PM. Banquet $14.95. For information: Steve Tolf, K1ST, 12 Phylmor Drive, Westboro, MA 01581.

PENNSYLVANIA: The Foothills ARC's 16th annual Hamfest, St. Bruno's Church, South Greensburg, Saturday, November 3. Tickets $2.00 or 3/$5.00. Indoor flea market tables $5.00. Food, refreshments. Mobile check-in 147.78/18. For information, tickets or tables contact WA3HOL or write F.A.R.C., PO Box 236, Greensburg, PA 15601.

PENNSYLVANIA: The R.F. *Hill Amateur Radio Club's annual* indoor Winterfest, Sunday, November 4, Sellersville National Guard Armory. Doors open 8 AM. Entry $2.00. Non-ham spouse and kids admitted free. Food on premises and nearby restaurants. Vendors indoor space $6.00 each, outdoor space $4.00 each. Admits one. Bring own tables. For reservations: PO Box 29, Colmar, *PA 18915 (215) 721-0278. Talk* in on 145.19 (R), 146.88 (R) and 146.52 simplex.

MASSACHUSETTS: The Honeywell 1200 Radio Club, sponsor of 147.72/12 repeater and the Waltham Amateur Radio Association, sponsor of 146.04/64 repeater, will hold their annual Amateur Radio and electronics auction, *Saturday*, November 17, Honeywell Plant, 300 Concord Road, Billerica. Doors open 10 AM. Free admission and parking. Snack bar and bargain parts store. Talk in on both repeaters. For information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

NEW YORK: Radio Central ARC "Ham-Central" Sunday, November 25, 1984, 9 to 3 PM. Social Hall, Temple of Isaiah, 1404 Stony Brook Road, Stony Brook, NY. Seminars will be presented. For information contact Bob Yarmus, K2RGZ (516) 981-2709 or write 3 Haven Ct., Lake Grove, NY 11755.

INDIANA: The 12th Fort Wayne Hamfest sponsored by the Allen County Amateur Radio Society, Sunday, November 11, Allen County Memorial Coliseum, Coliseum Blvd. Advance tickets $3.00; $3.50 at door. Tables $8.00. Premium tables

$20.00. No table sales at door. Ticket and table deadline October 20. All classes of exams given. Send Form 610 and SASE to: V.E. Coordinator, FWRC, P.O. Box 15127, Fort Wayne, IN 46885 by October 26. Large indoor flea market and commercial vendors. The infamous Ham Band directed by Luke Matthew, WB9DWJ. Vendor setup 5 AM to 7 AM. Public 8 to 4. Talk in .88. For information, tickets, tables: Hamfest Chairman AC-ARTS, PO Box 10342, Fort Wayne, IN 46851 or call Dave Smith, KA9FFET (219) 493-2439.

MICHIGAN: The Oak Park High School Electronics Club presents a Swap & Shop, Thanksgiving Sunday, November 25, Oak Park HS, Oak Park. Donations $2.00. Tables $6.00. Refreshments. For information: SASE to Herman Gardner, Oak Park HS, 13701 Oak Park Blvd., Oak Park, MI 48237. (313) 968-2675.

OHIO: The Massillon ARC will sponsor Auctionfest 84 on November 11, Massillon K of C Hall off Rt. 21. 8 AM to 5 PM. Sellers set up 7 AM. Admission $2.50 advance, $3.50 door. Tables available $7.00/8'. Refreshments. Dinner. Auction 11 AM. Talk in on W8NP, 147.78/18. For information/registration: MARC, 920 Tremont Avenue S.W., Massillon, OH 44646. SASE please.

OPERATING EVENTS

"Things to do..."

NOVEMBER 25 AND 26: The BOMB Squad (Best of Mt. Baldy) will operate W6HCP (Hollywood Christmas Parade) from 1600Z, November 25 to 0400Z, November 26. Frequencies: 7.284, 14.284, and 21.284 MHz SSB. SASE to W6GVR for special commemorative QSL.

NOVEMBER 22, THANKSGIVING DAY: A special events station (WA1NPO) will be operating from Plimoth Plantation in the museum's 1627 Pilgrim Village from 1300 GMT to 2000 + GMT with participation of the UK Club Station GB0UST, GB2UST, GB4UST. To receive a certificate, send proof of contact and $1.00 domestic or 4 IRC's to Whitman ARC, PO Box 48, Whitman, MA 02382. For information: KA1CZS (617) 826-4772; WB1CNM (617) 586-7524. Rosemary Carroll, Plimoth Plantation, PO Box 1620, Plymouth, MA 02360. (617) 746-1622 or Peter Jackson, G3ADV, 32 Brown Avenue, Parkfield, Nantwich, Chesshire, UK Phone 0270-627149.

NOVEMBER 10 AND 11: The Armored Force Amateur Radio Nationwide Emergency Team (A FAR NET) will sponsor a Veteran's Day special event station event station from 1200 GMT Saturday to 2400 GMT Sunday. Primary frequencies: 7.285, 14.325, 21.375 and 28.640 ± QRM. For a certificate send QSL and large SASE to Alfred G. Beutler, 36 Manchester Road, East Aurora, New York 14052.

NOVEMBER 17 AND 18: VK versus the World. Sponsored by the CW Operators QRP Club. Contestants may work DX or own country for scoring. QRP stations must sign QRP for identification. 0000Z Nov. 17 to 2400Z Nov. 18. Exchange: All stations 6 digits comprise RST followed by serial number, commencing with 001 to 999 then commence again. For information SASE to Contest Manager, PO Box 109, Mt. Druitt, N.S.W. 2770 Australia.

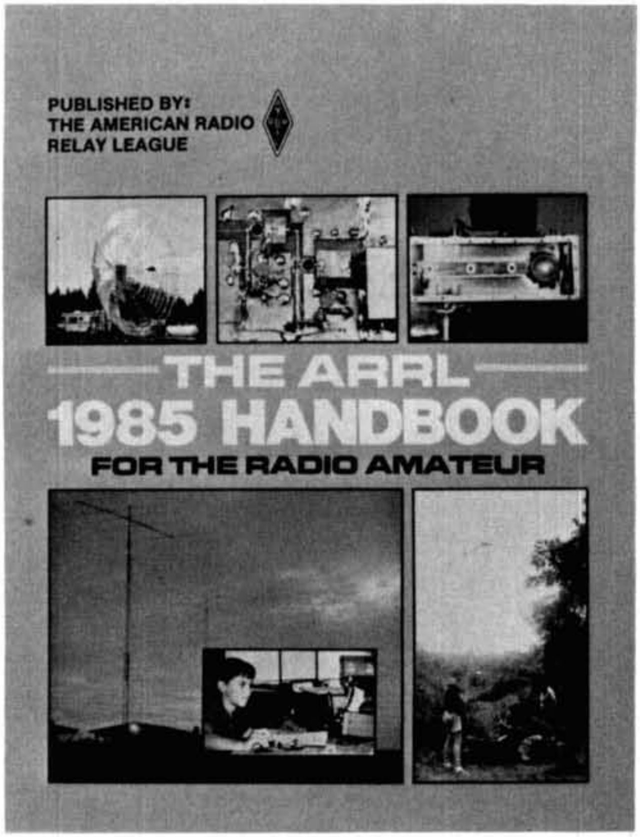
DECEMBER 1 AND 2: The 20th annual Telephone Pioneer QSO Party starts 1900 UTC Saturday to 0500 UTC Monday. 1.8-420 MHz +. Exchange: Contact number and chapter number. ITPA Club or chapter name. Send logs showing date, time station worked, chapter name and number, contact number and claimed score prior to January 15, 1985 to: Ted Phelps, W8TP, c/o John D. Burlie Chapter No. 89, TPA, 6200 East Broad St., Columbus, OH 43213.

DECEMBER 2: "Packet Radio Overview and Prospective" will be the subject of the 2nd North American Teleconference Radio Net (TRN). Learn about packet radio from two of its leading developers by tuning into TRN at 6 PM CST (0000Z). For a complete list of gateway station locations and frequencies write TRN Manager, c/o Midway Amateur Radio Club, PO Box 1231, Kearney, NE 68847-1231 SASE please.

THE AMATEUR RADIO MOTORCYCLE CLUB NET has moved to 3.888 MHz each Thursday night at 0300Z. All brands of bikers and riders are welcome. For more info send large SASE to Gary McDuffie, Rte. 1 Box 464, Bayard, NE 69334.

THE DELAWARE-LEHIGH ARC (W3OK) will operate Dec. 21, 22, 23, 1984 on 3.990, 7.299, 14.225, 21.325 and 28.525 MHz speading Holiday best wishes from Bethlehem, PA, The Christmas City. Large SASE to colorful certificate c/o DLARC Greystone Bldg. Gracedale, Nazareth, PA 18064.

VIRGINIA FONE NET 50th ANNIVERSERY CERTIFICATE commemorating 50 years of continous operation on the 75 meter band passing traffic in Virginia is being offered by the VFN. Work 25 VFN members and send log to K4IEC with #10 SASE for certificate. Contacts must be made between 9/30/84 and 6/30/85.



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CONTENTS

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ADM ____ 101
ACC ____ 102
ARR ____ 103
AEA ____ 104, 304
Alden ____ 105
ALL ____ 106
Alpha Delta ____ 107, 108
Aluma ____ 109
AES ____ 110
Amateur-Wholesale ____ 111
ARRL ____ 112
Amidon ____ 113
Applied Inv. ____ 114
Astron ____ 115
Atlantic Surplus ____ 116
Audio Amateur ____ 117
Auto. Techniques ____ 119
ATV Mag. ____ 120
Azotic ____ 121
B & W *
Barry *
B.G. Micro ____ 122, 123, 124
Break ____ 128
Buckmaster ____ 125
Burghardt ____ 126
Burtons ____ 127
Butternut *
C & A Roberts ____ 129
Caddell *
CeCo ____ 131
CES ____ 132
Centurion ____ 133
Coaxial Dynamics ____ 134
Comm. Con. ____ 135
Comm. Spec. ____ 136
Computer Trader ____ 137
Connect ____ 138
Daytapro ____ 139
Delaware ____ 140
DEO ____ 141, 142
Digital Ins. ____ 143
Discount Discs ____ 144
Dixie Elect. ____ 145
Doppler ____ 146
Dynetic Systems ____ 147
Electro-Com ____ 148
EEB ____ 149
Encomm ____ 150
Eng. Cons. ____ 151, 152
Epsilon ____ 153
Fabtron ____ 154
Falcon *
Fair ____ 155
Fluke *
Fox Tango ____ 156, 157
GLB Elec. ____ 158
Hal Comm. ____ 159, 302
Hal-Tronix ____ 160
HamMaster Tapes ____ 161
HRB ____ 162
HR Mag. ____ 163
HRO ____ 166, 167
Ham Shack *
Hamtronics, N.Y. ____ 305
Harrison ____ 164
Hatry ____ 165
Heath ____ 301
ICOM ____ 168, 169
Idiom Press ____ 170
Inter. Med. ____ 171
Jensen ____ 172
JRS Dis. ____ 173
Kantronics ____ 174
KCS Elect. ____ 175
Kenwood *
Lewis Elect. ____ 176
MFJ ____ 177
MHz ____ 178
Madison ____ 179
Martin Eng. ____ 180
Merrimac Sat. ____ 181
Meshna ____ 182
Microbyte ____ 183
Micro Control Spec. ____ 184
Microwave Filter ____ 185
Midwest ____ 186
Missouri Radio ____ 187
Morning Dist. ____ 188
Mosley ____ 189
NCG ____ 190
Nampa ____ 191, 192
Nemal ____ 193, 303
N. A. Soar ____ 194
N&V ____ 195
Orbit ____ 196
Paullin ____ 197
P.C. Elec. ____ 198, 199
Phillips-Tech ____ 200
Precision Sat. ____ 201
Processor Concepts ____ 202
Pro-Search ____ 203
Public Domain ____ 204
RF Enterprises ____ 205
RF Products ____ 206
Callbook ____ 207
Radiokit ____ 208
Radio Warehouse *
Ramsey ____ 209
SAROC *
Sartori ____ 210
Sat. TV Mag. ____ 211
Scamp Systems ____ 212
Schurmann ____ 213
Shure ____ 214
Sinclair ____ 306
S.M.C. ____ 215
Skylane ____ 216
Spectrum Inter. ____ 217
Spectrum West ____ 218
Spi-Ro Dist. ____ 219
Ten-Tec *
Terra *
TE Systems ____ 220
Texas Towers ____ 221
TNT Radio ____ 222
Transleteronic ____ 223
UNR-Rohn *
Universal Elec. ____ 224
University Microfilms *
Vanguard ____ 225
Varian ____ 226
Vector Radio ____ 227
Vid-Com ____ 228
VHF Comm. ____ 229
Walter Test Equip. ____ 230
Westcom ____ 231
Western Elec. ____ 232
Western Sat. ____ 233
Wheeler *
Williams ____ 234
Yaesu ____ 235

*Please contact this advertiser directly.

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November 1984

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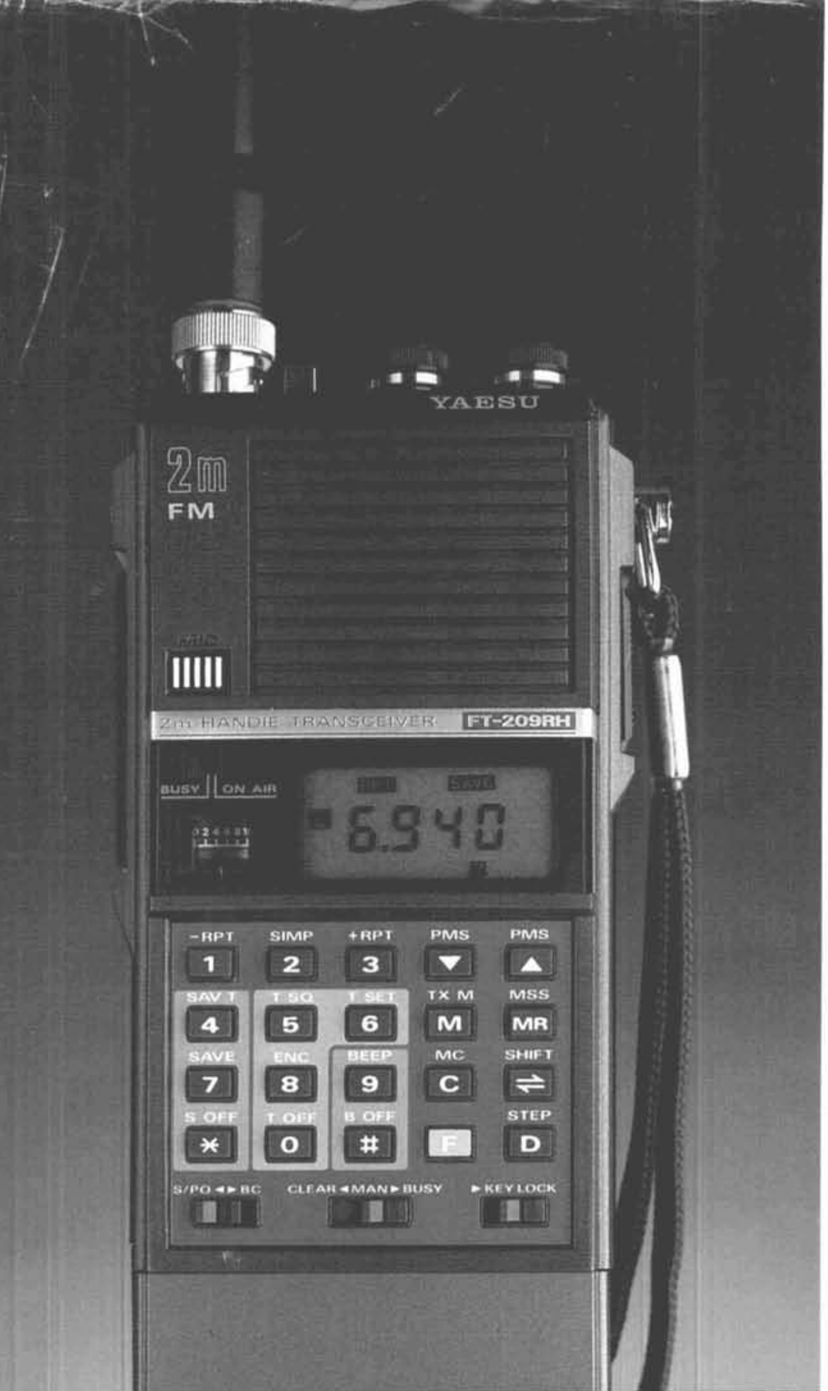
Advertisers index



...9RH shown actual size.

FT-203R transceiver.

The Yaesu FT-209RH. 5 watts that your batteries can live with.

Have the power you need when you need it with Yaesu's new 5-watt, 2-meter handheld. Power to get out in situations where ordinary HTs just won't make it.

We designed our HT with a unique user-programmable Power Saver that puts the rig to "sleep" while you're monitoring and "wakes it up" when the squelch breaks. So you can listen for hours and still have plenty of power to hit those hard-to-reach repeaters when you need to.

With the FT-209RH there's no need to fiddle with knobs when you change from one memory channel to another. That's because you can independently store everything you need in each of the ten memories: receive frequency, standard or non-standard offset, even tone encode/decode with an optional module. And then recall any channel at the touch of a button.

It's easy to hear what's happening on your favorite repeaters or simplex frequencies. Just touch a button and scan all memory channels, or selected ones. Or all frequencies between any two adjacent memories. Use the priority feature to return automatically to your special frequency when it becomes active.

Bring up controlled-access machines with the optional plug-in subaudible tone encoder/decoder, independently programmed from the keyboard for each channel. Listen for tone-encoded signals on selected channels—without having to hear a bunch of chatter—by enabling the decode function.

The FT-209RH, which covers 10 MHz for CAP and MARS use, comes complete with a 500-mAh battery, charger and soft case.

For those who want a basic radio without the bells and whistles, consider the compact, lightweight FT-203R. This economical HT features 2.5 watts of power and an optional DTMF keypad. Most all the accessories for the 209 work with the 203, including an optional VOX headset that gives you hands-free operation that's perfect for public service events.

So when you visit your dealer, let him know you won't settle for anything but the best. A radio built by Yaesu.

YAESU

Yaesu Electronics Corporation
6851 Walthall Way, Paramount, CA 90723
(213) 633-4007.

Yaesu Cincinnati Service Center
9070 Gold Park Drive, Hamilton, OH 45011
(513) 874-3100.

Prices and specifications subject to change without notice.

✓ 235

KENWOOD

...pacesetter in amateur radio

Digital Code Squelch...

TR-2600A

Kenwood's TR-2600A introduces DCS (Digital Code Squelch) circuitry, a signaling concept developed by Kenwood. DCS allows each station to have its own "private call" code or to respond to a "group call" or "common call" code. There are 100,000 different 5-digit ASCII code combinations possible. You can program in call signs up to 6 digits in the ASCII code. When operating in the DCS mode, this information can then be automatically transmitted each time the transmit key is depressed. This revolutionary feature is only the beginning! The TR-2600A also sports a high impact plastic case, that is extra rugged and scuff-resistant. The molded-in color adds to the attractive appearance. The large L.C.D. display is easy to read in direct sunlight or in the dark with a convenient lamp switch. It displays transmit/receive frequencies, memory channels, and five arrow indicators for "F LOCK" frequency lock, "REV" repeater reverse, "PROG.S" programmed scan, "MS" memory scan, "ALERT.S" alert scan. A star indicates "MEMORY LOCK-OUT" is activated, and repeater offset indicated by "+, −, S and M." The TR-2600A has 10 memories, nine for simplex or transmit with frequency offset ±600 kHz and one (memory 0) for non-standard split frequencies. Memory scan and programmable band scan have the added convenience of "Time operated Resume" that stops on busy channel and holds for approximately 5 seconds, then resumes scanning, or "Carrier Operated Resume" that stops on busy channel and resumes when signal ceases.

Memory scan, scans only those memories in which data is stored, and memory lock-out allows you to skip selected memory channels without loss of data previously stored! Manual Scanning UP/DOWN in 5-kHz steps and programmable automatic band scan are also useful features. The TR-2600A has a built-in "S" meter on the top panel which also indicates battery level when in transmit mode. Extended frequency coverage, 142.000-148.995 MHz allows transmit capability in 5-kHz steps for simplex or repeater operation on most MARS and CAP frequencies. Receive frequency coverage includes 140.000-159.995 MHz.

These features only tell part of the story. The TR-2600A also has keyboard frequency selection, built-in 16-key autopatch encoder, "TX STOP" switch, HI (2.5)/LOW (300 mw) power switch, REV switch, "SLIDE-LOC" battery pack, high efficiency speaker, BNC antenna terminal, and all of this in an extremely compact and lightweight package!

Kenwood's TR-2600A, with D.C.S., leads the way in high technology handheld transceivers!

Optional accessories:

- TU-35B built-in programmable sub-tone encoder
- ST-2 Base Stand
- MS-1 Mobile Stand
- PB-26 Ni-Cd Battery
- DC-26 DC-DC Converter
- HMC-1 Headset with VOX
- SMC-30 Speaker Microphone
- LH-3 Deluxe Leather Case
- SC-9 Soft Case
- BT-3 AA Manganese/Alkaline Battery Case
- EB-3 External C Manganese/Alkaline Battery Case
- RA-3, 5. Telescoping Antenna
- CD-10 Call Sign Display

More information on the TR-2600A is available from authorized dealers of Trio-Kenwood Communications, 1111 West Walnut Street, Compton, CA 90220.

Specifications and prices are subject to change without notice or obligation.