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ham radio magazine

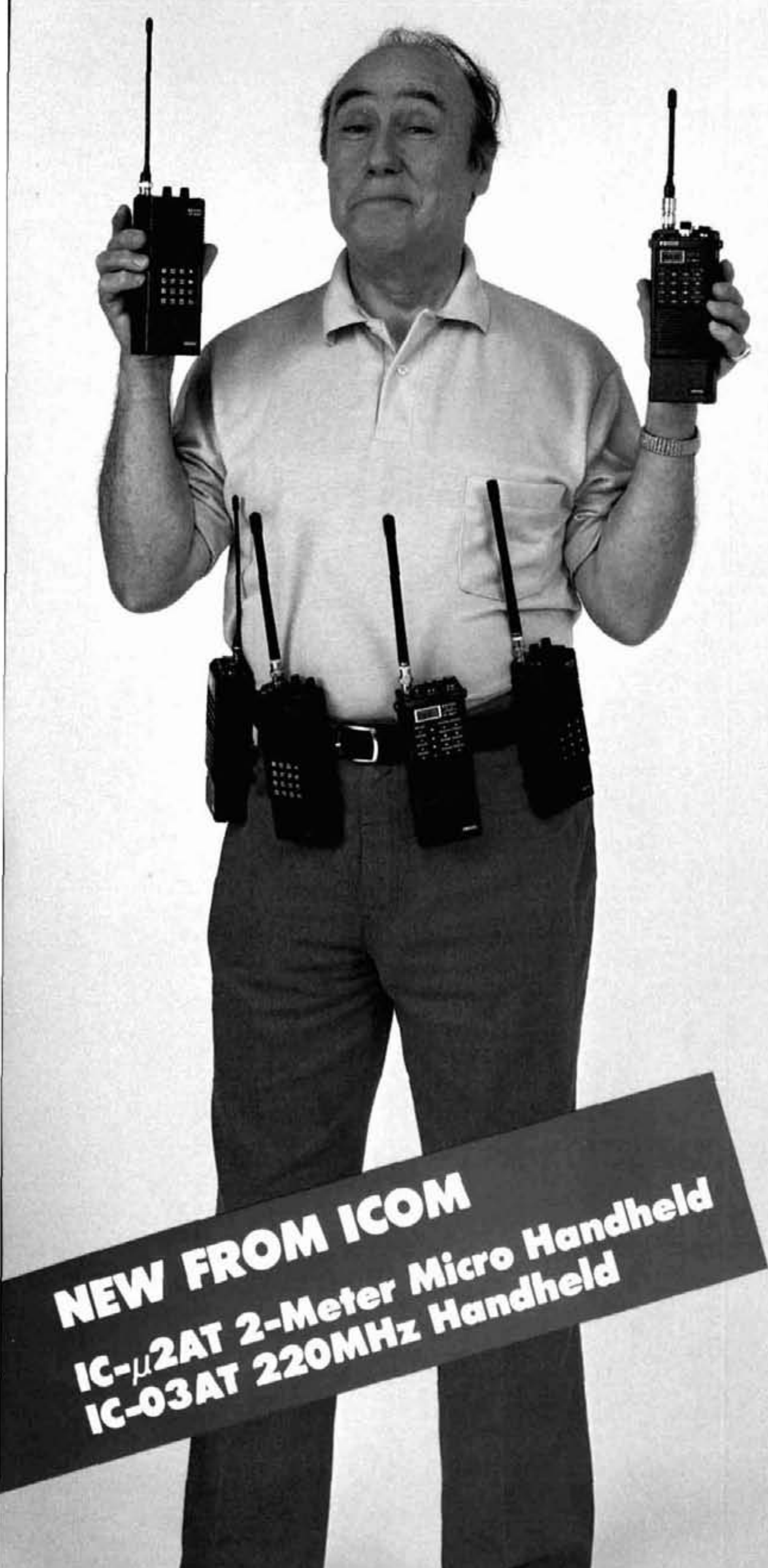
hr

focus on communications technology

annual RECEIVER issue

ICOM HAND HELDS

SURROUND YOURSELF WITH THE BEST!



Reliable. ICOM's extensive line of reliable, field-proven handhelds and interchangeable accessories give you the most options for handheld communications. 2-meter, 220MHz, 440MHz or 1.2GHz...ICOM has your frequency covered.

2-Meters. For 2-meter coverage, ICOM offers the IC-02AT and IC-2AT handhelds. The versatile IC-02AT covers 140.000–151.995MHz, the IC-2AT 141.500–149.995MHz...both include frequencies for MARS and CAP operation. The IC-02AT features an LCD readout, 32 PL tones standard, DTMF, direct keyboard entry, three watts output, (optional 5 watts output with IC-BP7 battery pack), 10 memories and three scanning functions. The IC-2AT, the most rugged handheld on the market, has a DTMF pad, 1.5 watts output and thumbwheel frequency selection. The IC-2A is also available and has the same features as the IC-2AT except DTMF.

220MHz. To get away from the crowd, ICOM has the IC-3AT 220.000–224.990MHz handheld with 1.5 watts output, thumbwheel selection and a DTMF pad.

440MHz. For 440MHz operation, ICOM has two handhelds available, the versatile IC-04AT and the IC-4AT. The IC-04AT and IC-4AT offer full coverage from 440.000–449.995MHz. The IC-04AT includes an LCD readout, 32 PL tones standard, DTMF direct keyboard entry, three watts output, (optional 5 watts output with IC-BP7 battery pack), 10 memories and three scanning systems. The IC-4AT has a DTMF pad, thumbwheel selection and 1.5 watts output.

1.2GHz. ICOM announces the IC-12AT 1260.000–1299.990MHz handheld, the first 1.2GHz handheld available. The IC-12AT features 10 memories, an LCD readout, DTMF direct keyboard entry, two scanning systems and one watt output.

Accessories. A variety of interchangeable accessories are available, including the IC-BP8 800mAH long-life battery pack, HS-10 boom headset, CP1 cigarette lighter plug and cord, HM9 speaker mic (for IC-02AT, IC-04AT and IC-12AT), leather cases, and an assortment of battery pack chargers.

ICOM
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ICOM America, Inc., 2380–116th Ave NE, Bellevue, WA 98004 **Customer Service Hotline (206) 454-7619**
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The new TS-940S is a serious radio for the serious operator. Superb interference reduction circuits and high dynamic range receiver combine with superior transmitter design to give you no-nonsense, no compromise performance that gets your signals through! The exclusive multi-function LCD sub display graphically illustrates VBT, SSB slope, and other features.

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- **Low distortion transmitter.** Kenwood's unique transmitter design delivers top "quality Kenwood" sound.
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- **Semi or full break-in (QSK) CW.**
- **40 memory channels.** Mode and frequency may be stored in 4 groups of 10 channels each.
- **Programmable scanning.**
- **General coverage receiver.** Tunes from 150 kHz to 30 MHz.
- **1 yr. limited warranty.** Another Kenwood First!

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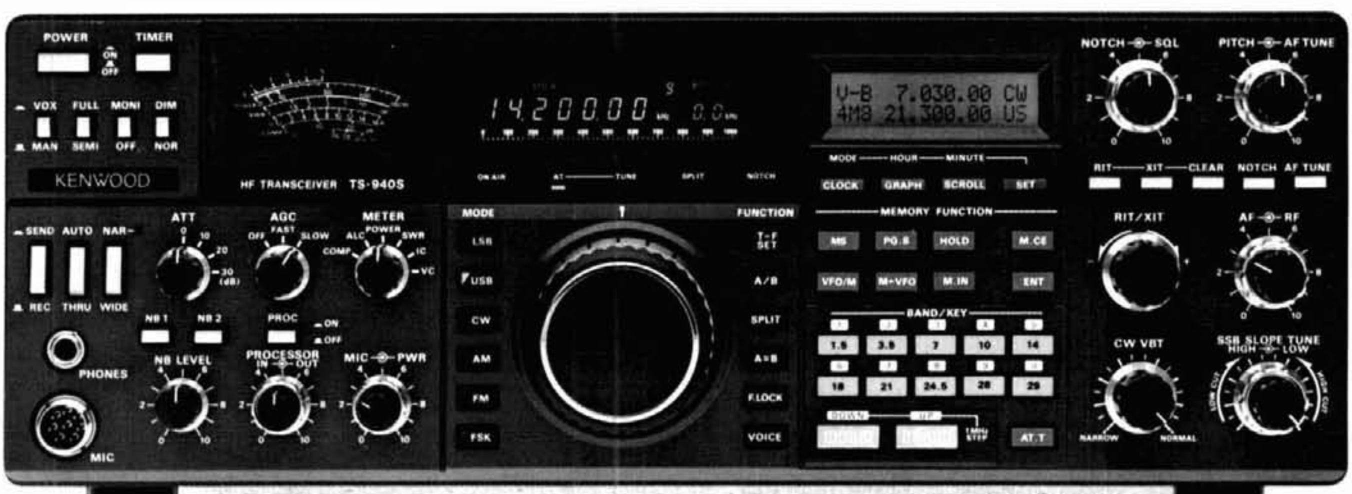

Complete service manuals are available for all Kenwood transceivers and most accessories.
Specifications and prices are subject to change without notice or obligation.

More TS-940S information is available from authorized Kenwood dealers.

KENWOOD

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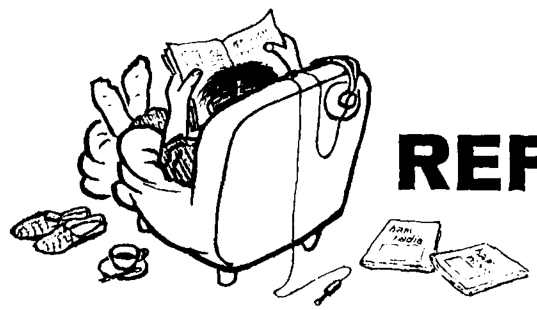

REFLECTIONS

specs in secs: an idea whose time has come

What's smaller than a bread box, *lighter than a Bic pen, faster than a speeding* bullet, and of interest to thousands of Radio Amateurs (or least it *will* be)?

You guessed it. It's Motorola's solution to our transistor specification problem, all on a single floppy disk. Appropriately labeled *Specs in Secs*, that's exactly what you get when you load it into your IBM PC-compatible computer.

In the program, Motorola has provided device information for over 1600 bipolar power transistors and TMOS™ Power MOSFETs, and included a user-friendly method of retrieving the data as well. It's not only extremely useful, but also fun to use. And because of its flexibility, you feel like you're making real design decisions — and you are!

Did I just hear someone say "Now just hold on one dang minute! We're Radio Amateurs, not engineers. Why do we need this information, anyway?"

I'm really glad you asked. First of all, some of us are engineers or technicians, in addition to being licensed Radio Amateurs. And even if you're not an engineer or a technician, I'm willing to bet that sometime in your Ham career you've designed some circuit, or at least wished that you could have. *Specs in Secs* won't do the circuit design for you (although there are software programs out there that will). What it will do is, through a few keystrokes, provide parts choices in *seconds*.

"Great!" you say, with a little bit of sarcasm. Now you have 1600 or so *power* device specifications. But like most other Radio Amateurs, you're interested in rf circuits — oscillators, preamplifiers, etc. Well, don't sell a power device short before you look at its specifications — f_T, for example, which is related to maximum frequency of operation. Some of these power devices provide real gain at frequencies that are of interest to us.

Without dwelling on this point, let me mention that Motorola indicated in the brochure that accompanied this diskette that they're working on entering their entire semiconductor product line on a single 360K floppy disk, and that takes time. I should know; I've done something similar with my compendium of Amateur Radio article references, *From Beverages Thru OSCAR — A Bibliography* (November 1980), and that only took me six years and eight diskettes, and I'm *almost* finished!

"But is *Specs in Secs* easy to use?" you ask.

It couldn't be simpler. You place the disk in any drive, type the letter M (Enter), and away you go. A menu provides a number of choices, including, in essence, an on-screen manual. Actually, the first thing you see is a "Start-up Screen" with a carefully chosen set of choices (defaults) that you'll probably want anyway.

Here's where it really becomes useful. Selection "D" is called "Parametric Search." With this choice, you're given the opportunity not only to choose the important parameters, but their order and value as well. The more specific you are, the more quickly you'll arrive at the appropriate component. Motorola does, however, recommend that you also have the hard copy (selection guide) available for the final decision-making process.

If you're reasonably sure of the component you want to use, just enter its part number after pressing Selection "C," appropriately labeled "Part Number Search." If it can't be found, you won't be hit over the head or knocked out of your chair by a loud noise. The program will just quietly tell you that it can't find that particular part number. Believe me, you still have many other choices.

I could easily go on about its other features, but for $2 you can get your own copy and see for yourself. By the way, Parlez-vous Francais? Or German, Spanish, or Italian? The on-screen manual is written in these other languages as well, and you can print all of them out and have a copy at your side.

For your copy of *Specs in Secs* (DK101/D), send a check or money order for $2 to Motorola Semiconductor Products, Literature Distribution Center, P.O. Box 20924, Phoenix, Arizona 85063.

Rich Rosen, K2RR
Editor-in-Chief

comments

Dick Smith kits

Dear HR:

Perhaps you've heard this, but in case not: I just found out that Dick Smith Electronics is no longer doing business in the United States. The Dick Smith kits are now being sold by American Electronics, P.O. Box 301, Greenwood, Indiana 46142.

Joe Moell, K0OV
Fullerton, California 92633

say what you think

Dear HR:

What a delightful guest editorial by Robert Zavrel, Jr., W7SX (September, 1987)! Nearly everything listed was something I could identify with. However, I had to move some of the "don't likes" into the "do likes" column, and vice versa. Not many, but some.

That started me thinking. Wouldn't it be interesting to see how other readers felt? As Bob said in the article, "there comes a time when a guy's got to say what he really thinks!" My mind started wandering. It would be really fun to see what men and women hams think about things! What about a checklist with an item per line, and a box for "Y" or "N" after each one?

Bob listed 28 "really don'ts" and 49 "really dos." That makes a total of 77 things to feel strongly about. I would like readers to send me their opinions on any or all of these 77 "reallys," either by mail or by packet BBS (KZ1O @ WB1DSW). The message could be kept short by just sending a 77-character long "opinion string" of Y's (like) and N's (don't like).

I'll tabulate the answers and send the results to *ham radio* magazine, because, as the last line of the editorial said I also like "magazines crazy enough to publish this."

That's what the last "Y" in my opinion string stands for:
YYNNNNYYNNNNNNNNNNNNNNN
NNNNNNYYYYYYYYYNYYYYYYY
YYYYYYNYYYYYYYYYYYYYYYY
YYYYYYYY.
I'm looking forward to hearing from a lot of readers!

Dave Bushong, KZ1O
2 South Spring Street
Concord, NH 03301-2424

a word of caution

Dear HR:

Thank you for the article, "Solar Activity and the Earth's Magnetosphere," (August, 1987). It's very well written, and Bradley Wells is to be congratulated. For the layman, it answered many questions regarding the sun's relation to radio propagation. It also provided a good summary of the mechanics of the lines of magnetism.

There is one point I would like to bring to your attention, however. There are at least two instances where the phrase "may be visible to the naked eye" occur: on page 11, under the subhead "sunspots," and on page 13 under "flares."

At the risk of appearing overly cautious, it seemed to me that this implies that all one has to do is to gaze at the sun and these flares or sunspots would be apparent. However, readers who value their eyesight should be warned that "gazing" (my word) at the sun can be devastating. I'd therefore suggest a word of caution to the casual reader.

I hope there will be many more pieces about the sun; with the Solar year apparently coming out of its "dip," we could all benefit from a greater understanding of solar phenomena.

Rupert A. Wood, WB4ZOF
Bethesda, Maryland 20816-1760

tomorrow's receivers:
what will the next 20 years bring?

New techniques, technologies promise lower power, smaller size, higher performance

It seems that with each new issue of *ham radio* and other journals, we witness new technological developments affecting nearly every aspect of Amateur Radio. From time to time many of us ponder the questions, "Where is it all going?" and "What will rigs look like 10, 20, or more years from now?"

Predicting the effects of today's research on tomorrow's reality is always tricky business. But new techniques now in use in commercial and military radio systems are likely to find their way into Amateur applications sooner or later. I've discussed some of these techniques here, emphasizing what I see as their implications for Amateur Radio design.

Recent breakthroughs in technology may change some of the fundamental ground rules not only of radio engineering, but of electrical engineering in general. Consequently, two levels of development will be addressed here: first, those techniques in current use and second, those which may be possible in the near future.

Figure 1 shows a block diagram of a typical radio receiver. This is the familiar superheterodyne design, which has been predominant for nearly 60 years. Most commercial receivers on the market today use this same architecture — and in truth, there's been little improvement in radio performance for the past 30 years![1]

Other characteristics have certainly changed for the better, however: size, power consumption, 12-volt operation, frequency stability and accuracy, and ease of operation, for example. Yet the basic receiver functions remain the same; an rf signal down to the submicrovolt level must be amplified, converted, filtered, amplified again, demodulated, amplified again, and then converted into acoustic audio. There's nothing to suggest that the principle of rf input and audio output will change as a basic function for radio engineers. Rather, it's how you get from point A to point B that's undergoing a quiet revolution.

Perhaps the most striking developments have been in miniaturization. There's a continuous trend towards developing components with excellent specifications but with smaller sizes and lower power consumption.

rf amplifier

The first stage of a receiver largely determines the noise figure. Up to about 15 or 20 MHz, there's little advantage in using low-noise amplifiers because atmospheric and galactic noise are more significant than the noise figure of the typical first mixer stage. For this reason the preamp has actually disappeared from many hf receivers. Diode ring mixers typically have about a 7-dB noise figure, which is quite adequate for most hf receivers. The preamp also lowers the dynamic range of a receiver as it increases the rf level to the first mixer. At VHF and higher, the atmospheric noise drops and the receiver noise figure becomes one of the most important system specifications. At VHF frequencies and higher, the GaAsFET has dramatically reduced receiver noise figure specifications. This trend will continue as GaAs technology improves and prices decline.

first mixer

The most common rf mixer is the passive quad diode ring. It has a relatively low noise figure and the limitation to dynamic range is mainly a function of LO power used. A generalization can be made about the LO power necessary to handle a given rf power level: the ratio of maximum rf input power to minimum LO power is about 1:10. That is, for a diode mixer to handle 100 mW of rf, the LO power must be at least 1 watt.

By Robert J. Zavrel, Jr., W7SX, P.O. Box 23447, Tucson, Arizona 85734

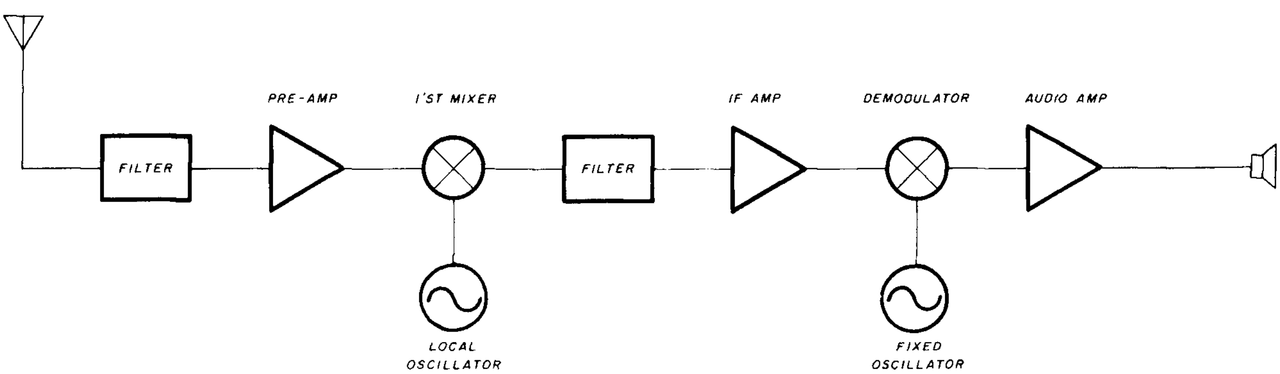

fig. 1. Typical present-day superheterodyne receiver.

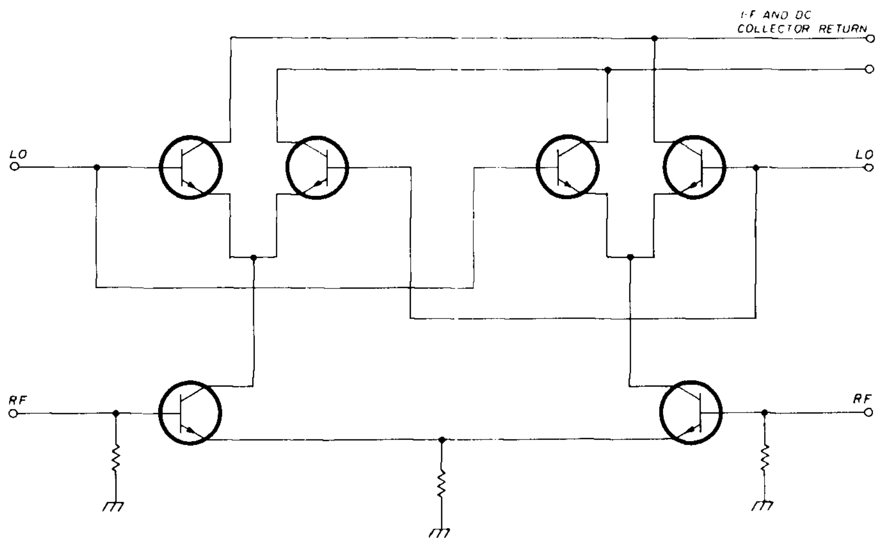

fig. 2. Gilbert Cell multiplier-mixer.

There have been two major developments in mixer technology during the past few years. The first is the passive FET ring, exemplified by the Siliconix Si8901. With this device, gate voltage rather than a forward-biasing current turns the switches "on" and "off." Since the gates represent high impedances, voltage/power ratios can be increased, thus lowering the LO power requirements dramatically. Indeed, to handle the same 100-mW rf power in our diode ring example, the Si8901 requires about 25 mW of LO power instead of the 1 watt mandated by the diode rings. The other critical specification is the third-order intercept point, which is necessary for defining the useful dynamic range. Again, the Si8901 greatly surpasses the old diode ring mixer.[2]

Working from the same empirical thinking that led to the development of the Si8901, a passive GaAsFET ring should surpass the performance of the silicon Si8901 by perhaps a 7-dB increase in third-order intercept point specification. Since the GaAs devices would be used as switching elements rather than active amplifiers, the 1/f noise limits of these devices wouldn't be an issue. They could be used at sub-audio frequencies with comparable noise performance at hf.

Although purported as a passive mixer, the Si8901 should also make an excellent active mixer. Using the same concepts as the old U350, the DMOS Si8901 should outperform its JFET cousin. The smaller geometry SD201 DMOS family made excellent VHF low-noise amplifiers before Signetics discontinued its DMOS line several years ago. Siliconix and the other DMOS manufacturers have chosen not to build these smaller devices, although both mixer and amplifier performance could be enhanced by such a modification.

As the world of digital integrated circuits has shifted its attention to faster CMOS technologies, advances in analog bipolar IC techniques have quietly proceeded. A fundamental bipolar mixer circuit is the

Gilbert Cell (**fig. 2**). Though the most familiar Gilbert Cell ICs are the Motorola MC1496 and MC1596, there are many manufacturers of these devices today. Over the past few years there have been several variations on this original commercial design. Good noise performance has been achieved with the Signetics NE602, but it can't handle the higher input levels necessary for good hf first mixer design. The NE602, however, is perhaps the finest mixer available among low power consumption mixers. Advances in bipolar processing are pushing noise figures down and power handling capabilities up in simulation models as well as in newly available devices. If this trend continues, the diode ring may become an endangered species.

local oscillator

Perhaps the most dramatic advances in receiver technology have been in the design of local oscillator circuitry. Very exotic mechanical assemblies evolved for LO tuning in the 1950s. (Remember the NC300, HQs, and Collins receivers?) Permeability tuned oscillators (PTOs) simplified things in the 1960s with the SB300, R4, and S-lines. Today, PTO LO performance remains difficult, if not impossible to duplicate with PLL synthesis, given the constraints of typical Amateur budgets. Synthesizers offer distinct advantages in that they can be directly controlled by microprocessors and don't require special mechanical rigidity or moving parts. A single crystal oscillator frequency is divided down to some low value and then "phase-locked" up to the desired LO frequency. At the output of the VCO, a multiple of the reference frequency equal to the reference frequency times the divider's "N" value appears (see **fig. 3**). This represents a simple phase-locked loop synthesizer, but it contains all the necessary building blocks of a more sophisticated system. If the reference is 1 kHz and the divider is set to N = 7005, the VCO output will be 7005 × 1 kHz, or 7.005 MHz. Discrete frequency steps of 1 kHz will be possible because only division by whole numbers is possible in this system. N determines the output frequency because the dc control voltage feedback will "lock" the VCO to the output frequency that provides equal frequency inputs to the phase detector.

Another type of synthesizer — the "direct digital synthesizer" or DDS — holds great promise. To understand how DDS works, two concepts must be understood: first, the concept of how digital-to-analog converters function and second, the Nyquist Theorem. I'll discuss each of these very briefly.

A digital-to-analog converter (DAC) takes a binary number and "converts" it into a discrete voltage or current value. An eight-bit DAC, for example, can have a maximum of 256 different voltage outputs. The digital eight-bit "word" can be generated by a micro-

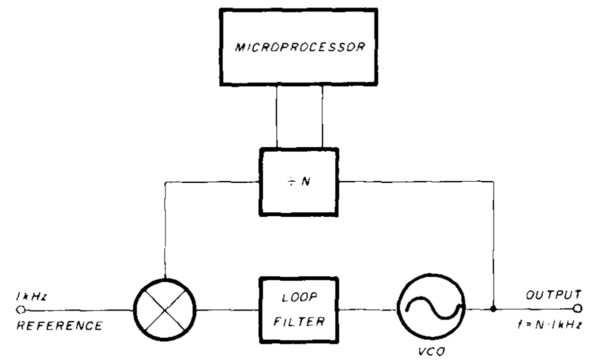

fig. 3. Simple digital PLL frequency synthesizer.

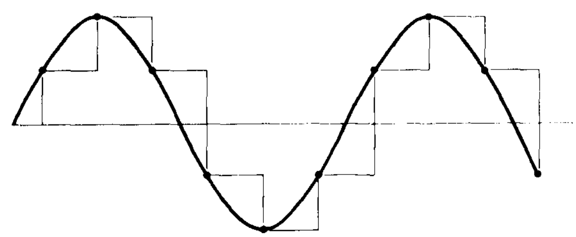
fig. 4. Sine wave digital approximation.

processor or memory circuit. A sine wave or any other waveform can be approximated by discrete steps as shown in **fig. 4**. For LO applications, sine wave approximation is preferred because sine waves have the lowest harmonic content. (A perfect sine wave will have no harmonic content.)

The Nyquist Theorem states that a sine wave can be derived if at least two discrete amplitude samples per period are obtained. Thus, if we want to synthesize a 5-MHz sine wave with a DDS, we'll need an updating clock rate of at least 10 MHz. With DDS, a constant sample rate can be used to synthesize any frequency up to half the sample rate with excellent frequency resolution (0.1 Hz typical). **Figure 5** shows how 1- and 2-MHz sine waves can be generated by a 10-MHz clock and a DAC. The microprocessor computes the values to be sequenced by the DAC for a given frequency output and a given clock rate. This special processing function is called a phase accumulator. Higher clock rates allow for more samples per period. More samples, in turn, allow for better approximation of the sine wave shape. Also, the greater number of bits allows for more discrete amplitude steps, which also enhance the accuracy of the sine wave approximation. DAC integral linearity affects waveform accuracy and harmonic levels. Indeed, the major efforts in DAC development are directed towards

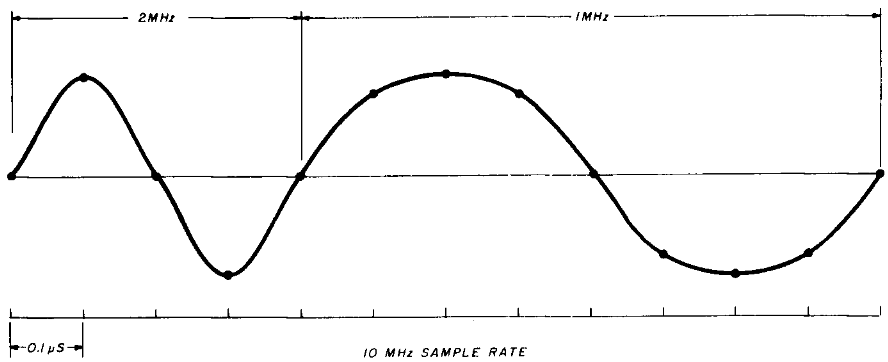

fig. 5. DDS sinewave synthesizer. Any frequency can be synthesized up to 1/2 the sample rate by changing the digital numerical value at each sample point.

better resolution (i.e., more bits) at higher update speeds. Off-the-shelf video-speed DACs now allow excellent performance for DDS circuits well into the hf spectrum. The Burr-Brown DAC63 is such a device. As DACs become faster and more accurate, the phase noise and harmonic performance improve.

Compared with PLL synthesizers, DDS offers other advantages. Since all the parameters of the waveform (frequency, amplitude, and phase) are digitally controlled, frequency hopping, or QSY, is almost *instantaneous*. PLL systems, on the other hand, have some finite settling time. In addition, nearly any type of modulation is possible if the applicable parameter is changed in accordance with a modulating waveform. Outstanding linearity is possible even at very wide bandwidths — in fm, for example.[3] To achieve the necessary low spectral noise densities demanded by hf LOs, more work is needed; however DDS holds great promise as a replacement for PLL synthesizers.

Figure 6 shows a direct conversion phasing SSB receiver. Traditional analog phase and amplitude nulling techniques employing all-pass active filters could yield 40-dB image rejection at best. The problem of this nulling is compounded by the need for broadband 90-degree phase shifters.[4] However, if we digitize the two signals the phase and amplitude nulling can be performed using a digital signal processor (DSP). Small errors in phase and amplitude can be removed

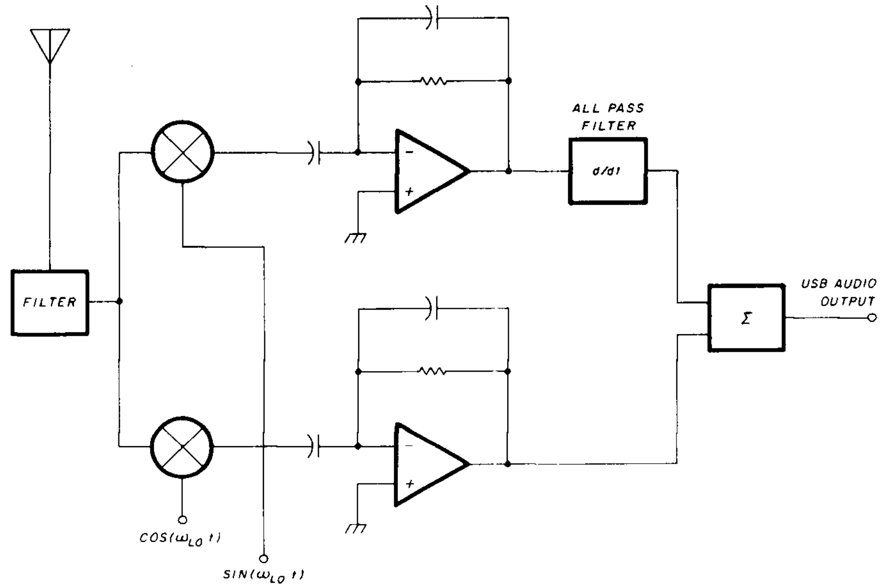

fig. 6. Direct conversion SSB receiver.

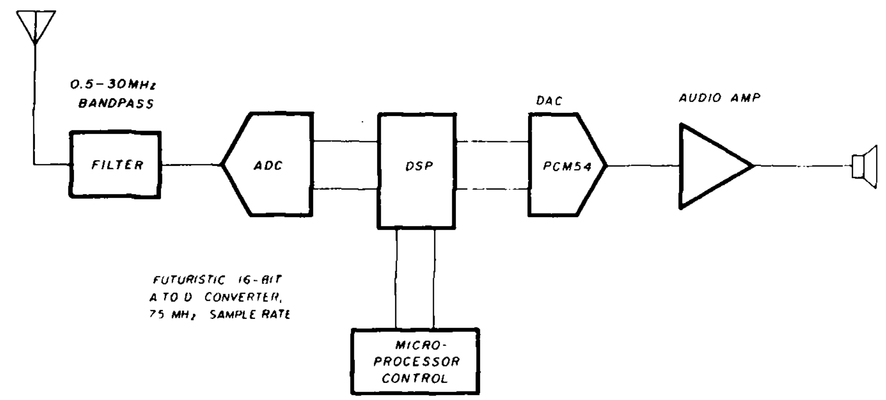

fig. 7. Digitized radio of tomorrow.

and the ultimate image rejection will correspond to the bit resolution of the analog-to-digital converter ADC. A 16-bit ADC, for example, could provide 96- dB image rejection using an ideal DSP. This would make an excellent receiver. Very sharp analog audio filters would be required before the ADCs. Philips has produced such a dual-channel filter in monolithic form, but it's not yet commercially available. More traditional upconversion schemes using DSP are discussed in an outstanding new textbook.[5]

Once the demodulated audio baseband signal is digitized, digital filtering techniques can be used. Very steep skirt notch and bandpass filters can be arranged. The amplitude coefficients can be manipulated to allow idealized audio AGC. Alternately, an AGC voltage can be provided through a DAC which controls an rf stage gain. Finally, the digitized audio can be converted back into analog form via a DAC, in much the same way as compact disc players function. The Burr-Brown PCM54 is an excellent 16-bit DAC for this application.

Figure 7 represents a daydream of where radio engineering state of the art might be within 20 or 30 years. A bandpass filter from 0.5 to 30 MHz tunes the entire hf spectrum. A futuristic 16-bit ADC with a 75-MHz sampling rate provides a 96-dB dynamic range over the bandwidth. The entire spectrum is digitized. All filtering, demodulation, AGC, and such, is performed in the rather sophisticated DSP. A PCM54 is then used to output the audio to an audio amplifier.

superconductors

In April the news media reported that a group working at an IBM facility in Switzerland had developed a new material that remains a superconductor up to 85 degrees K. This material, by itself, makes superconductors possible at liquid nitrogen temperatures, thus dramatically lowering the costs of using this class of material. Perhaps of far greater importance is that this work represents a crucial breakthrough for creating better superconducting materials. There is even talk of superconductors at room temperatures. What can this mean?

Superconductors are created when a material is cooled down to a critical temperature. Below the critical temperature the material exhibits *zero* electrical resistance. When superconductor offers literally no resistance at room temperature, electronic device technology could advance dramatically. Josephson junction or SQUID digital circuits could render even the fastest computers now available obsolete. For analog and rf circuits, zero resistance could have great implications for speed and noise specifications. Thermal noise disappears in superconductors as $V^2 = 4KTRB$. Never mind Boltzman; R is now zero. With zero resistance, charge mobility — a limiting factor for the speed of any semiconductor material — becomes quite high.

With superconductors, storage "battery" technology would be revolutionized. Imagine an electromagnetic car battery the size and weight of a donut! Charging efficiency would approach 100 percent.

Superconducting antennas and transmission lines would improve efficiency and lower system noise figures. Can you imagine 160-meter loading coils with zero resistence?

This is the most exciting development in solid-state physics since the invention of the transistor. The implications may by far outweigh the transistor's effects on the world. Electronic and electrical power engineers will have to rewrite all the books — again.

what to watch for

The radio art is in constant evolution. Here are some of the trends to watch for between now and the advent of the twenty-first century:

- Continued miniaturization of all components, thanks to higher levels of circuit integration and advances in wafer processing techniques.
- Lower supply voltage and current requirements for

✓ 250

comparable performance and improved performance, generally among all building-block components.

- Increased use of digital techniques in LO circuitry, and use of DSP for filtering, demodulation, and other functions.
- Appearance of data conversion devices, first at baseband, then at the i-f, and then moving closer to the antenna circuitry.
- Dick Tracy's two-way wrist radio will become a reality by 1995.

Putting predictions into print preserves the prophet's prognostications for posterity. It might be amusing, in the year 2007, to dust off your yellowed, musty copy of this issue to see just how far off the mark we were. Happy dreams!

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ham radio

designing a state-of-the-art receiver

Readily understood — though not greatly utilized — concepts mean better performance

The state of the art in hf receiver design using semiconductors has improved greatly. The use of either CATV-type transistors and double-balanced mixers using hot carrier diodes or double-balanced mixers with switch-type FETs has eased the large-signal handling problem of just a decade ago.

One weak link in the chain, however, remains; this is the synthesizer, with its inherent noise contributions. To a large extent the overall architecture of the receiver and the synthesizer determines its performance, and even the best high-performance components — placed in the wrong sequence — can cause a good design to fail.

systems approach

Because military and commercial users depend on high performance receivers for surveillance and/or point-to-point communication, it's inevitable that these same technological advances will filter down to the Amateur community. In fact, in a cursory examination, the spec sheets of both commercial and Amateur communications receivers look quite similar.

Besides providing the "essentials," modern communication receivers offer additional features, sometimes referred to as "bells and whistles"; these features include improved user interfaces or computer interfaces for remote control. Since the commercial and Amateur markets are price-sensitive and also very sensitive to proof of performance, any claims of lower capabilities are noticed. Consequently, when on-the-air tests of some late-model receivers suggested poorer performance than previous models, this raised the question of why, despite the knowledge acquired in recent years, such an inconsistency should occur.

Figure 1 is a functional block diagram of a modern microprocessor-controlled communication receiver. This diagram is representative of most modern design approaches and can be used to evaluate possible advantages and weaknesses, and to point out areas of potential difficulty.

operation

The rf signal is introduced into the receiver in one of two ways:

- at the input of a 30-MHz low-pass filter via a variable attenuator, which is controlled by an overload detection circuit activated at the first and second i-f level;
- or for receiving frequencies below 400 kHz, via a variable attenuator and low-pass filter combination.

The first mixer, which is responsible for the third-order intercept point, is driven by an extremely pure synthesized local oscillator. To terminate the double-balanced mixer properly, a diplexer (high-pass/low-pass filter) is used to absorb energy outside the crystal filter passband. The input impedance of the crystal filter rises significantly outside the passband of the crystal filter.

By *Ulrich L. Rohde, KA2WEU/DJ2LR*, *52 Hillcrest Drive, Upper Saddle River, New Jersey 07458*

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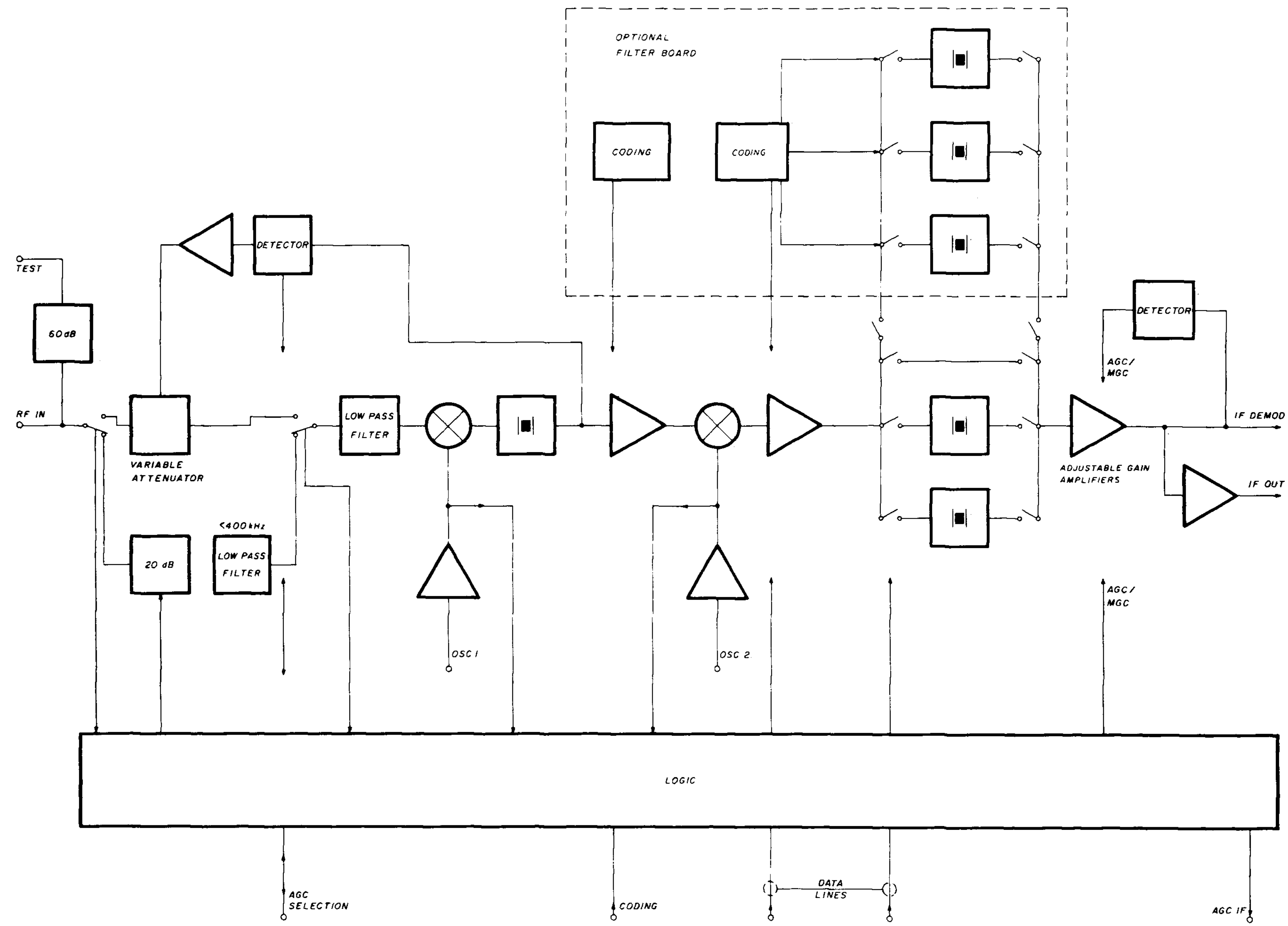


fig. 1. Functional block diagram of a modern microprocessor-controlled communication receiver.

While most commercial or military high-performance receivers employ the input stage combination, most Amateur equipment uses a double-balanced mixer that incorporates adjustable-gain JFETs. As a result of its sensitivity to output impedance changes, the mixer suffers reduction in large- signal performance. The recent trend in mixer design involves the use of termination-insensitive mixers whose cleverly designed bridge circuits ignore the effects of reactive terminations. Passive mixers use hot carrier diodes; switching-type mixers use FETS — for example, a pair of matched SD 100 transistors — and achieve intercept points between +35 and +45 dBm.

Following a six- to eight-pole crystal filter is an amplifier stage which has medium gain and high dynamic range. This is typically achieved through the use of rf feedback. In addition, it is worthwhile to incorporate an AGC circuit between this amplifier and the mixer.

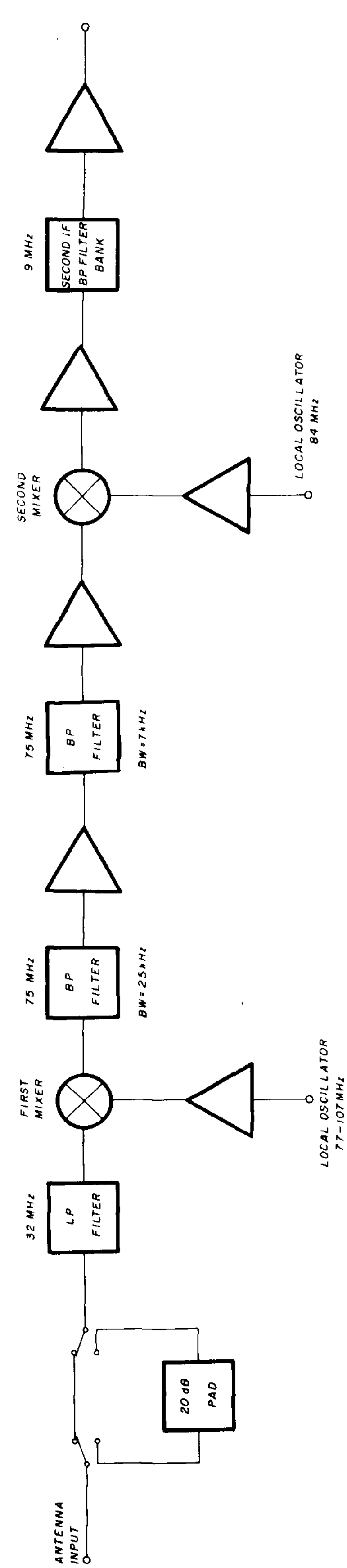
The signal, which is now in the 75- to 100-MHz frequency range, is converted to a low i-f frequency — for example, from 200 kHz to 2 MHz — by the second mixer. The second mixer must also have high dynamic range, but it can be a passive double-balanced mixer. (The latest advances in receiver design have included the use of careful filtering of the local oscillator synthesizer outputs, thereby reducing spurious responses.) The amplifier that follows this mixer compensates for the second i-f mixer losses. This, in turn, is followed by a popular frequency range crystal filter that is readily available from a number of manufacturers.

gain distribution is important

Each of these stages has very little gain — typically less than 12 dB. The main amplification of the signal takes place in the i-f sections. (This is different from what happens in Amateur receiver designs.) The problem with designing most of the gain into the i-f stages has to do with the ability to build the i-f amplifier circuits stable and free of unwanted oscillations. To minimize in-band intermodulation distortion, differential-type amplifiers with AGC stages are used. In many cases this requires a great deal of shielding and careful selection of grounds, since up to 100-dB open loop gain in the i-f section may be required.

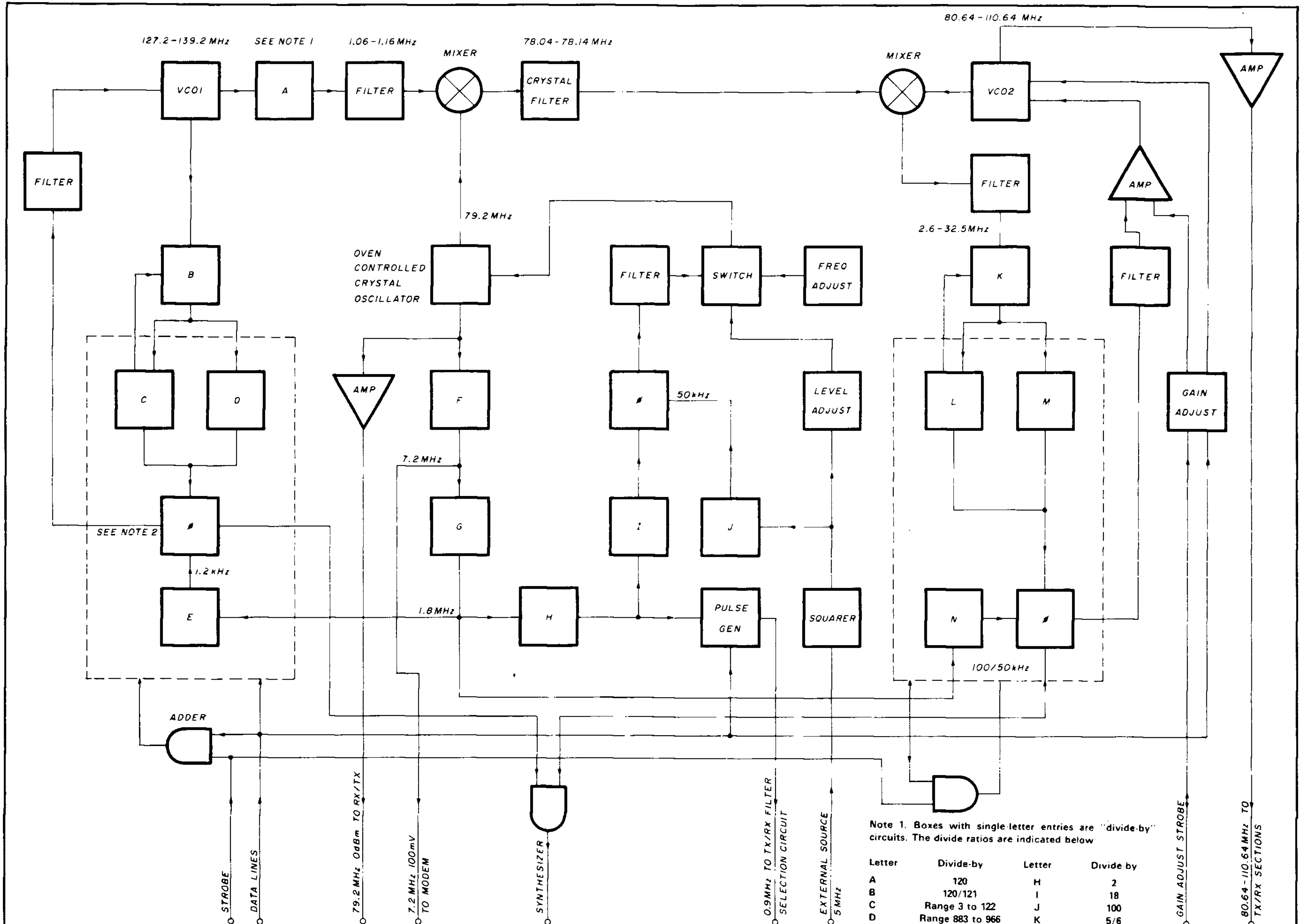
One sign of good receiver design is evident when the noise of the first mixer, with no antenna connected, already shows slight AGC action, which can be monitored on the S-meter of the receiver. If signals of 1-μV or better are required before any S-meter action occurs, then the above design guidelines have *not* been followed.

Although I've noted these things thoroughly and repeatedly in previous articles, very few companies have followed through with this concept because it's much less expensive to move the gain towards the antenna than to build high-gain i-f stages.



Power gain	dB	-0.5	6	1	10	3.5	12	6	8	4	>90
Noise figure	dB	–	6	–	2	–	3	6	5	–	4
Cumulative noise figure	dB	10.81	10.31	4.31	3.31	8.18	4.68	12.02	6.02	8	4
IP_2	dBm	–	80		•	–	•	•	•	–	*(out)
Cumulative IP_2	dBm	86.5	86	–	–	–		–	–	–	*(in)
IP_3	dBm	–	32	–	30	–	30	27	30	–	*(out)
7-kHz in-band IP_3	dBm	13.9	13.4	7.4	6.4	16.6	13.1	26.8	22	–	*(in)
25-kHz in-band IP_3	dBm	27.2	26.7	21	20	–	•	•	•	–	*(in)
Out of band IP_3	dBm	38.5	38		•		•	•	•		*(in)

fig. 2. Noise, gain, and intermod products are indicated on this block diagram of a modern communications receiver.



Note 1. Boxes with single-letter entries are "divide-by" circuits. The divide ratios are indicated below

Letter	Divide-by	Letter	Divide-by
A	120	H	2
B	120/121	I	18
C	Range 3 to 122	J	100
D	Range 883 to 966	K	5/6
E	1200	L	Range 3 to 5
F	11	M	Range 9 to 64
G	4	N	18 or 36

Note 2. All boxes labeled with Greek letter "phi" are phase detectors.

fig. 3. Multi-loop frequency synthesizer utilizes carefully separated analog and digital circuits.

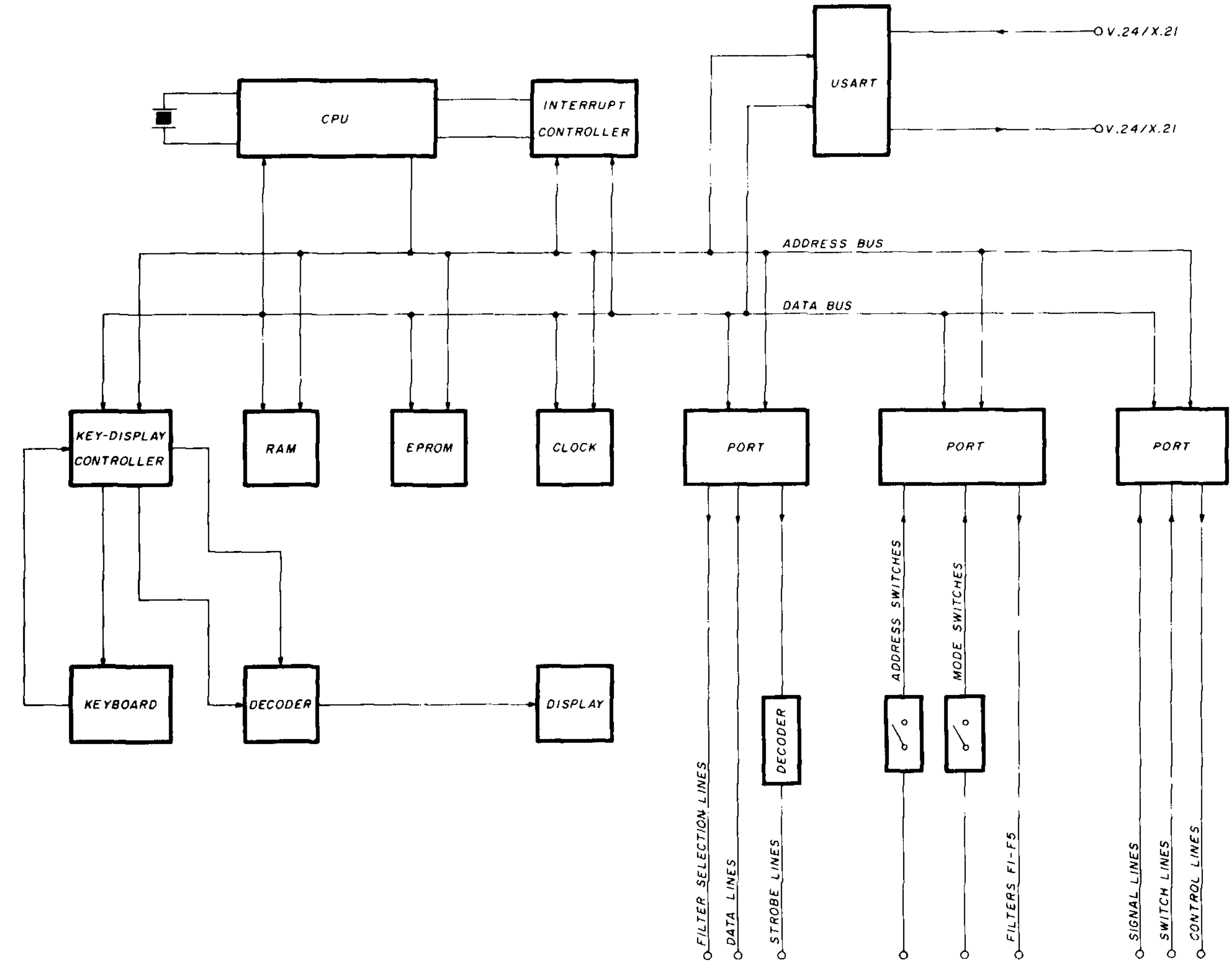


fig. 4. Architecture of the internal computer system found in a modern fully synthesized receiver.

Interestingly enough, the NRD 525 receiver, which follows these recommendations but is still fairly inexpensive, is one of the better designs. **Figure 2** shows a typical block diagram with the various noise, gain and intermodulation distortion products specified. Such an analysis must be carried out and should be published together with the receiver specifications.

In **fig. 1**, the internal bus for the receiver that controls a variety of functions transfers digital data streams that consist of short-duration pulses with fast rise times. Consequently, significant shielding is required in this section in order to isolate the digital circuits and their concommitant switching spikes from the analog portion of the receiver. Many modern receivers suffer from this effect, in which background switching noise masks lower level signals. To make things worse, the synthesizer can also pick up some of the switching signals.

synthesizer design

Modern receivers should incorporate fully synthesized local oscillators and provide between 1- and 10-Hz resolution. All of the auxiliary frequencies in the design must be derived from the master (oscillator) standard.

The frequency synthesizer in this example uses two main loops in its multi-loop design. VCO1, which oscillates between 127 MHz and 139 MHz, is phase locked in steps of 1.2 kHz; its output is then divided by 120 to an output frequency of approximately 1 MHz, which is then mixed with the 79.2-MHz standard. The difference frequency components are introduced into a 78.1-MHz crystal filter which removes all unwanted signals. The other portions of the synthesizer provide auxiliary frequencies. The areas where the phase detectors are located are heavily shielded; fairly high frequencies for the reference detectors are used for best noise performance.

Figure 3 shows a synthesizer that utilizes this design approach. In this design, analog and digital circuits are carefully separated. The sections of the synthesizer most vulnerable to picking up extraneous signals are the lines going into the output VCO (VCO2).

Depending on the type of phase detector used, the lines that feed the tuning diodes can be either very high impedance or low impedance. Inexpensive solutions frequently lead to such high-impedance feeding points, which then become "antennas" collecting all the switching noise. The use of circuits incorporating microwave transistors allows the design of discrete low-impedance amplifiers for this purpose. If the driving point impedances for the tuning diodes can be held at 100 ohms as opposed to the 100-k line impedance typically found, *the sensitivity for pickup of stray signals is reduced by* a factor of 100,000.

Another reason for noisy synthesizers is the use, in the synthesizer loops, of operational amplifiers that are too noisy. Wherever possible, either discrete low-noise amplifiers or Darlington stages must be used.

use of microprocessors

Today's microprocessor-controlled receivers feature built-in clocks, frequency scanning with variable scan rates, availability of at least 100 channels and channel scanning, plus a combination of receiver control functions such as the serial RS-232 or IEEE-488 bus remote control capabilities. Because the BFO and the main oscillator are both synthesized, the combination of the two allows either passband tuning or variable bandwidth.

Another area of interest in the use of microprocessors *is the linearization of the transfer characteristic of the* tuning range of the oscillator and the linearization of the S-meter. The microprocessor can also switch the tuning rates to correspond to the operating mode and select the appropriate bandwidth receiving crystal filters required for that same mode. Digital implementation of signal analysis allows demodulation of RTTY and Morse code. Many other novel approaches are possible.

Figure 4 shows the architecture of such an internal computer system. One of the frequent mistakes made in this context is the use of only one microprocessor, which gets overloaded, or the use of four-bit microprocessors. In better radios, eight-bit microprocessors, which can handle all these functions efficiently, are used. The best approach is parallel processing.

summary

By following these simple guidelines and using architecture similar to that illustrated in **figs. 1** and **2**, it is possible to build receivers that come close to the limits of physics, yet still remain cost- effective.

bibliography

Ulrich L. Rohde, KA2WEU/DJ2LR, *Digital PLL Frequency Synthesizers: Theory and Design*, Prentice-Hall, Inc., Englewood Cliffs, New Jersey 07632, 1983. (Currently out of print, this book is available in limited quantities through the author.)

Ulrich L. Rohde, *T. T. N Bucher Communications Receivers: Principles and designs*, McGraw Hill, New York, N. Y., 1988.

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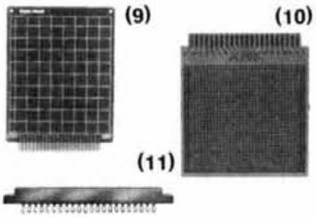

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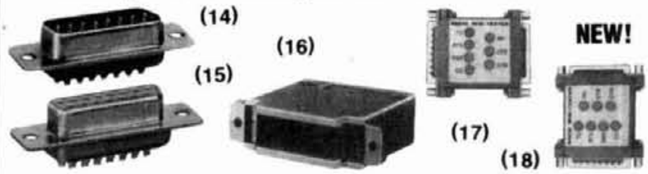

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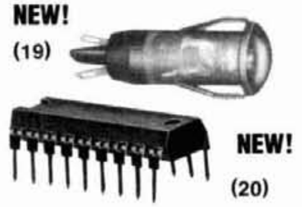

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a CAT control system
for the Yaesu FT-757GX

C-64 BASIC routine tunes popular transceiver

An earlier article on the CAT system, based on the Tandy TRS-80 Model 100, attracted the attention of a considerable number of users of Commodore 64® microcomputers.[1,2] But while it was a relatively simple task to convert the TRS-80 program for other micros with a standard RS-232C port, it wasn't as easy to convert the program for the popular C-64.

Converting the unique Commodore "inverted TTL level" format is no problem because the CAT system also works with TTL levels. But the standard baud rates of the C-64 stop at 2400, and CAT works at 4800 baud. (In this case, baud and bps give the same number, so we'll stick to "baud," since that's what's used in the Commodore and Yaesu literature.)

It's possible to obtain the proper parameters for 4800 baud using a two-step formula for calculating "user-specified baud rates."[2] Checking the output with an oscilloscope looked promising, but in practice the data transfer wasn't reliable, and bytes sometimes were lost on their way from BASIC to the radio. Commodore specialists have several explanations for this. Some say that it depends on the physical layout of the printed circuit board; others contend that Commodore BASIC has problems handling the NUL character. Whatever the reason, the data didn't always reach the RS-232C output buffer in good order.

In this program these problems are avoided by replacing the BASIC RS-232C statements with routines written in machine language. The parameters are POKEd into temporary storage in memory locations 52592 to 52596. The command byte (see **table 1**) goes into location 52592. The frequency number is sliced into four bytes (i.e., 12,345.67 kHz converts to 01, 23, 45 and 67), which are stored in the next four locations.

Another problem has also been solved in machine language. The frequency bytes just mentioned are in the CAT Binary Coded Decimal format, and ideally would be passed as such to the RS-232C port. But the Commodore BASIC interpreter expects all numbers to be in decimal form and converts them into hexadecimal when the program is executed. A look-up table is used to reconvert the hex into the proper BCD format. The five bytes are then sent out one by one, starting from the highest address, as required by the CAT system.

BASIC program

The BASIC part of the program (**fig. 1**) is straightforward and can be segmented as follows:

The program is initialized in **lines 100** to **220**. A title screen at **lines 3000** to **3160** (subroutine) is displayed while the machine language part in **lines 4000** to **4114** (subroutine) is being loaded. **SYS 52480** is similar to **OPEN File No. 1** for 4800 baud, eight bits per byte, two stop bits, and no parity. **Line 210** determines the starting frequency.

The main screen is set up in **lines 1000** to **1150** (see **fig. 2**). The keys available for commands are indicated within brackets. The upper row on the keyboard is used for general commands; the lower row — plus "A" and ";" — for tuning in steps of 10 Hz to 100 kHz. "F" is for steps of any size and "Q" for quitting the program.

Note that Band Down and Band Up have different functions, depending on whether the FT-757GX is in Amateur Band or General Coverage mode, and whether or not MR/VFO has been activated (see the FT-757GX Operating Manual). This short program doesn't take into account frequency changes made with Band Down or Band Up, so in order to get the correct screen display, you'll need to reinitialize the program by pressing "F" and entering the actual frequency before using the Fine Tuning keys. This isn't a problem, since you can still go directly to any frequency with the "F" command.

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Table 1. Command byte codes.

function	hexadecimal	decimal
Band Down (multifunction)	08	8
Band Up (multifunction)	07	7
Dial Lock (on/off)	04	4
Clarifier (on/off)	09	9
Split Frequency (on/off)	01	1
Switch between VFO A and B	05	5
Transfer VFO to Memory	03	3
Transfer Memory to VFO	06	6
Exchange between VFO and Memory	0B	11
Temporary check of Memory	02	2
Frequency Set	0A	11

fig. 1. BASIC language program for CAT control of Yaesu FT-757GX.

```
10 REM =============================
20 REM CAT CONTROL FOR YAESU FT-757GX
30 REM        AND COMMODORE 64
40 REM   BY KJELL W. STROM, SM6CPI
50 REM                  JUNE 5, 1987
60 REM =============================
70 REM
100 REM *** LOAD ML AND OPEN FILE FOR 4800 BAUD ***
110 GOSUB 3000:GOSUB 4000:SYS 52480
200 REM *** SET INITIAL FREQ (10000 KHZ) ***
210 A=10000
220 GOSUB 2100
1000 REM *** MAIN SCREEN ***
1010 PRINTCHR$(147)"[RVS]  YAESU FT-757GX CAT PROGRAM BY SM6CPI  "
1015 PRINT:PRINT:PRINT
1020 PRINT"NEW FREQUENCY [F]                   QUIT [Q]"
1030 PRINT:PRINT:PRINT:PRINT"                   FINE TUNING"
1040 PRINT"           << - <KHZ> +> > "
1050 PRINT"[A]100                               100[;]"
1060 PRINT
1070 PRINT"10   5   1  .1  .01 .01 .1   1   5   10"
1080 PRINT"[Z] [X] [C] [V] [B] [N] [M] [,] [.] [/]"
1090 PRINT"[DOWN]------------------------------------"
1100 PRINT"    [1] BAND DOWN      [2] BAND UP"
1110 PRINT"    [3] DIAL LOCK      [4] CLARIFIER"
1120 PRINT"    [5] SPLIT FQ       [6] VFO A/B"
1130 PRINT"    [7] V=>M           [8] M=>V"
1140 PRINT"    [9] V=>/<=M        [0] MR/VFO"
1150 PRINT "[HOME][DOWN][DOWN]          KHZ":PRINT"[UP]"A
1160 GET C$:IF C$="" GOTO 1160
1170 IF C$="B" THEN A=A-.01:GOTO 1600
1180 IF C$="N" THEN A=A+.01:GOTO 1600
1190 IF C$="V" THEN A=A-.1:GOTO 1600
1200 IF C$="M" THEN A=A+.1:GOTO 1600
1210 IF C$="C" THEN A=A-1:GOTO 1600
1220 IF C$="," THEN A=A+1:GOTO 1600
1230 IF C$="X" THEN A=A-5:GOTO 1600
1240 IF C$="." THEN A=A+5:GOTO 1600
1250 IF C$="Z" THEN A=A-10:GOTO 1600
1260 IF C$="/" THEN A=A+10:GOTO 1600
1270 IF C$="A" THEN A=A-100:GOTO 1600
1280 IF C$=";" THEN A=A+100:GOTO 1600
1290 IF C$="F" THEN GOSUB 2800:GOTO 1000
1300 IF C$="1" THEN POKE 52592,8:SYS 52526:GOTO 1160
1310 IF C$="2" THEN POKE 52592,7:SYS 52526:GOTO 1160
1320 IF C$="3" THEN POKE 52592,4:SYS 52526:GOTO 1160
1330 IF C$="4" THEN POKE 52592,9:SYS 52526:GOTO 1160
1340 IF C$="5" THEN POKE 52592,1:SYS 52526:GOTO 1160
1350 IF C$="6" THEN POKE 52592,5:SYS 52526:GOTO 1160
1360 IF C$="7" THEN POKE 52592,3:SYS 52526:GOTO 1160
1370 IF C$="8" THEN POKE 52592,6:SYS 52526:GOTO 1160
1380 IF C$="9" THEN POKE 52592,11:SYS 52526:GOTO 1160
1390 IF C$="0" THEN POKE 52592,2:SYS 52526:GOTO 1160
1580 IF C$="Q" GOTO 2900
1590 GOTO 1160
1600 REM *** ROUND OFF AND CHECK RANGE ***
1610 A=INT(A*100+.5)*.01
1620 IF A<500 THEN A=500
1630 IF A>29999.99 THEN A=29999.99
1640 GOSUB 2100
1650 GOTO 1150
2100 REM *** SLICE FREQ AND OUTPUT ***
2110 A$=MID$(STR$(A),2)
2120 IF A=INT(A) THEN A$=A$+"."
2130 A$="000"+A$+"00"
2140 FOR I=1 TO LEN(A$)
2150 DP$=MID$(A$,I,1)
2160 IF DP$="." THEN DP=I:GOTO 2400
2170 NEXT
2400 A$=MID$(A$,DP-6,6)+MID$(A$,DP+1,2)
2410 F1=VAL(MID$(A$,1,2))
2420 F2=VAL(MID$(A$,3,2))
2430 F3=VAL(MID$(A$,5,2))
2440 F4=VAL(MID$(A$,7,2))
2450 POKE 52593,F1
2460 POKE 52594,F2
2470 POKE 52595,F3
2480 POKE 52596,F4
2490 POKE 52592,10
2500 SYS 52512
2510 RETURN
2800 REM *** NEW FREQUENCY ***
2810 INPUT"[DOWN][DOWN][DOWN]FREQUENCY KHZ";A
2820 A=INT(A*100+.5)*.01
2830 IF A<500 THEN A=500
2840 IF A>29999.99 THEN A=29999.99
2850 GOSUB 2100
2860 RETURN
2900 REM *** REM CLOSE FILE AND QUIT ***
2905 PRINT:PRINT:PRINT
2910 PRINT"                      ARE YOU SURE?"
2920 GET C$:IF C$="" GOTO 2920
2930 IF C$="Y" THEN SYS 52578:PRINT CHR$(147):END
2940 GOTO 1000
3000 REM *** TITLE SCREEN WHILE ML IS LOADING ***
3010 H1$="     C A T  P R O G R A M  F O R"
3020 H2$="     Y A E S U  F T - 7 5 7 G X"
3030 H3$="           BY KJELL W. STROM, SM6CPI"
3040 PRINTCHR$(147):PRINT:PRINT:PRINT:PRINT
3050 FOR I=1 TO LEN(H1$)
3060 PRINT MID$(H1$,I,1);
3070 NEXT
3080 PRINT:PRINT
3090 FOR I=1 TO LEN(H2$)
3100 PRINT MID$(H2$,I,1);
3110 NEXT
3120 PRINT:PRINT:PRINT
3130 FOR I=1 TO LEN(H3$)
3140 PRINT MID$(H3$,I,1);
3150 NEXT
3160 RETURN
4000 REM *** LOAD ML ROUTINES ***
4010 FOR I=52480 TO 52696:READX:POKEI,X:NEXT
4020 RETURN
4100 DATA 169,1,162,2,160,3,32,186,255,169,4,162,104,160,205
4101 DATA 32,189,255,32,192,255,162,3,189,108,205,149,247,202
4102 DATA 16,248,96,162,3,188,113,205,185,117,205,157,113,205
4103 DATA 202,16,244,162,1,32,201,255,173,14,220,41,254,141
4104 DATA 14,220,162,4,189,112,205,168,173,161,2,41,1,208,249
4105 DATA 152,32,210,255,202,16,238,173,161,2,41,1,208,249
4106 DATA 32,204,255,173,14,220,9,1,141,14,220,96,169,1,32
4107 DATA 195,255,96,0,0,2,0,0,206,0,207,75,87,83,56,55,0,1,2
4108 DATA 3,4,5,6,7,8,9,16,17,18,19,20,21,22,23,24,25,32,33
4109 DATA 34,35,36,37,38,39,40,41,48,49,50,51,52,53,54,55,56
4110 DATA 57,64,65,66,67,68,69,70,71,72,73,80,81,82,83,84,85
4111 DATA 86,87,88,89,96,97,98,99,100,101,102,103,104,105,112
4112 DATA 113,114,115,116,117,118,119,120,121,128,129,130,131
4113 DATA 132,133,134,135,136,137,144,145,146,147,148,149,150
4114 DATA 151,152,153
10000 REM *** 'RUN 10000' FIRST TO CHECK ML DATA LINES ***
10010 S=0:FORI=52480T052696:READX:S=S+X:NEXT
10020 IF S<>22360 THEN PRINT"CHECK ML DATA!":END
10030 PRINT"ML DATA OK!"

READY.
```

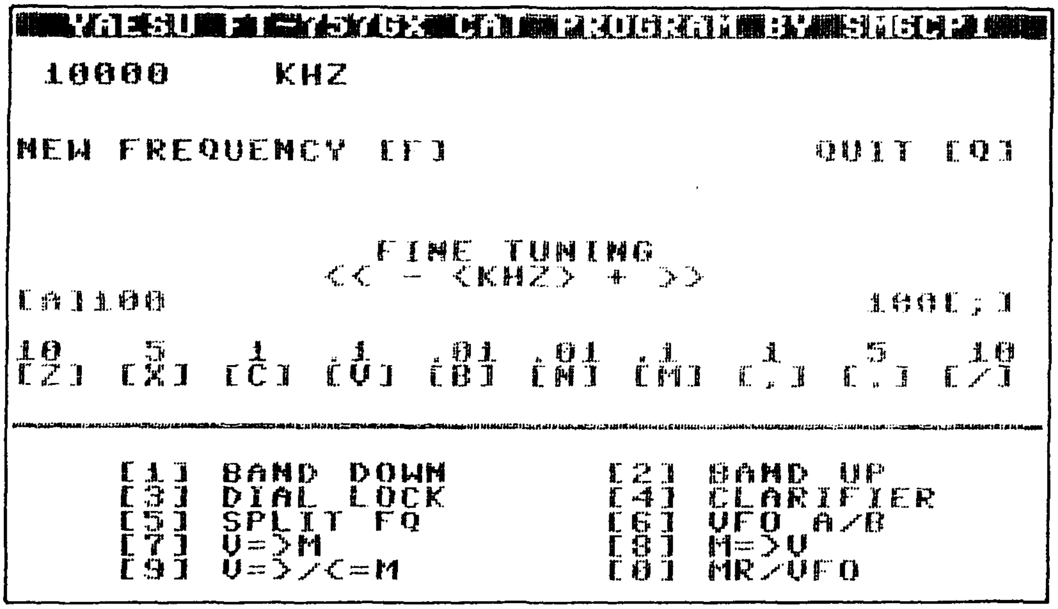


fig. 2. Screen display.

Commands entered from the keyboard are decoded by **lines 1160** to **1590. SYS 52526** sends the bytes to the radio without converting from hex to BCD, since this isn't required for the single-byte commands.

Verification that the frequency is inside the range of the FT-757GX is accomplished with **lines 1600** to **1650.**

In **lines 2100** to **2510** the frequency parameter is sliced into four bytes, as described above, after having been converted into a string and having found the position of the decimal point. **SYS 52512** is the call for sending out the frequency byte after a hex-to-BCD conversion.

The frequency input subroutine used with the "F" command is contained in **lines 2800** to **2860.**

The subroutine for the QUIT command is in **lines 2900** to **2940.**

The last part of the program contains subroutines for the title screen and for loading of the machine language part, as mentioned before.

The numbering of the program lines may seem haphazard; this is because some numbers were intentionally omitted in order to reduce the time needed for some of the GOSUBs and GOTOs. Renumbering to a tighter sequence may slow down the program a little.

After typing in the program, SAVE it and type **RUN 10000 / RETURN /** to confirm that the DATA lines have been entered correctly. Otherwise, if there's a mistake in the DATA and you try a RUN, you may lose the program!

interface

An interface circuit serves two purposes: it translates the input level to a suitable output level and stops potentially harmful interference from one connected unit from reaching the other, and vice versa.

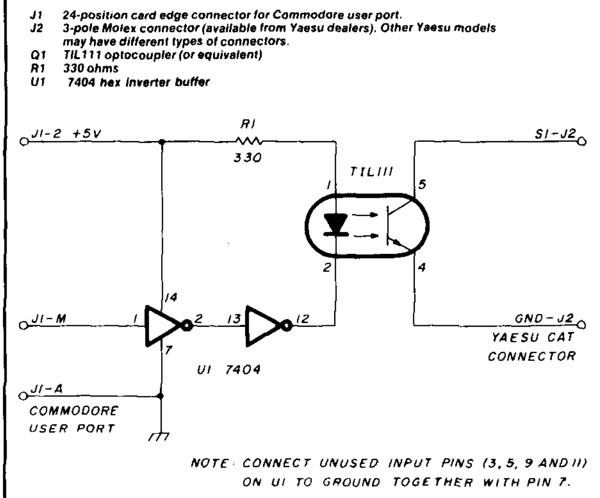


fig. 3. Schematic diagram.

Both the C-64 and the FT-757GX CAT system work with the same TTL levels, but the current needed to pull the CAT SI line to low, through a 680-ohm resistor, is probably higher than can be considered safe for the delicate C-64 CIA chip. We also want to minimize the possibility of computer noise reaching the receiver and of transmitter rf reaching the computer.

The Yaesu FIF-232C interface does this job. It accepts the inverted TTL level format from the C-64 if the internal switch S01 is set to the position opposite from normal. Since it also works with standard RS-232C format and with the two-way CAT systems of the FT-980, FT-757GX II, and FT-767GX, it can be regarded as a very flexible solution, especially since it has its own built-in power supply.

For our purpose we can also use the simple interface circuit shown in **fig. 3**. It receives 5 volts from the C-64 and can be assembled on a small IC-spacing perforated board as close as possible to the 24-position user port card edge connector. The cable from the TIL111 optocoupler output side should be shielded, with the screen connected to ground at the transceiver side only, not to computer ground. Connect all unused input pins on the 7404 IC to ground, together with pin 7.

summary

This article has been intended to show how easily you can interface and use the CAT system with the Commodore 64 and the Commodore 128 in C-64 mode. It also offers four machine language routines you can use in much more powerful programs, thereby adding your own special features to one of today's most appreciated transceivers.

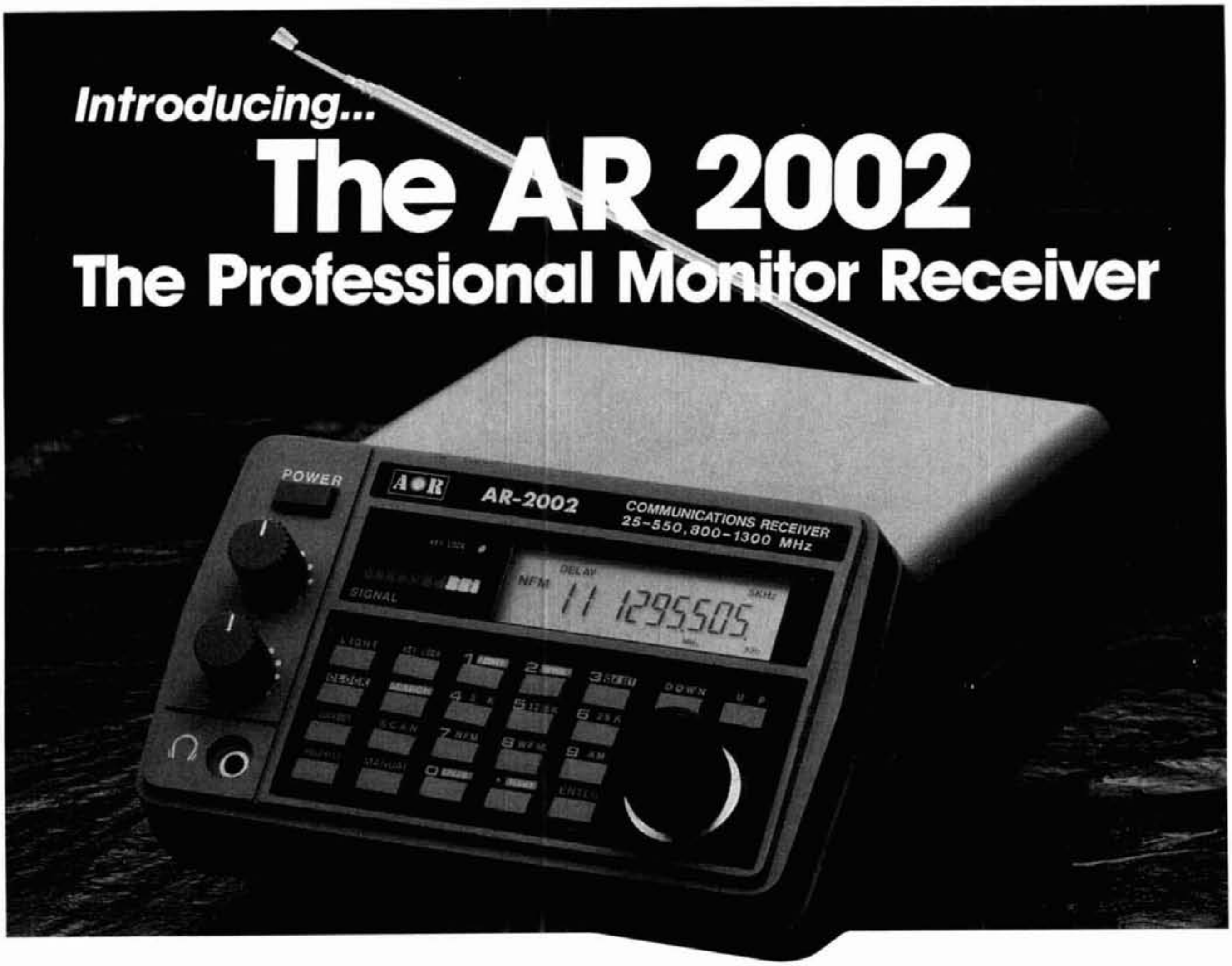
references

1. K. W. Strom, SM6CPI, "A CAT Control System," *QST*, October, 1985, page 38.
2. K. W. Strom, SM6CPI, "Feedback", *QST*, April, 1986, page 41.

ham radio

✓ 124

✓ 125



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the **weekender**

a thumbwheel frequency selector for the Yaesu FT-757GX

If you happen to own a Yaesu FT-757GX transceiver, this one-evening project will make its operation more convenient and enjoyable. If you don't own one, you may find that this article will show the ease and simplicity with which features of modern computer-controllable radios can be accessed.

Because the 757's minimum incremental tuning is in 500-kHz steps and the tuning dial moves only 10 kHz per rotation, any frequency change in other than 500-kHz steps takes several presses of the button and an average of 12-1/2 knob rotations! This project allows thumbwheel entry of any frequency, in or out of the current band. The design makes *no* modifications to the radio and connects to it with only two plugs on the back panel. A circuit board layout is given for those who wish to etch a board; the design can also be wire-wrapped, with no degradation in performance. All normal radio functions — including tuning with normal front panel controls — are still available.

design considerations

It's best to think of the thumbwheel frequency selector (TWFS) as a unidirectional data source. Yaesu provided an interface to accept commands from an external source on the back panel of the 757, which, among other things, accepts serially encoded binary data as requests for frequency selection options; these are documented on page 10 of the 757 Technical Supplement Manual. In addition to duplicating front panel operations, a command is provided for loading any specific frequency. Using this command, discrete frequencies may be entered without all the front panel dial spinning. The 757 never sends data, so no provision needs to be made for data flow *from* the transceiver.

This circuit sends a prepackaged stream of 50 bits to the radio at 4800 bits per second every time you push a "load" button. The outputs of six BCD thumbwheel switches are part of the 50-bit stream and indicate what frequency is requested. The thumbwheels provide frequency selection with a resolution of 100 Hz. A two-wire data cable goes from the TWFS to the radio back panel and connects to pins 1 and 2 (the left and middle pins as viewed from the back) of a three-pin connector formally designated as J12 or the EXT CNTL jack. (See pages 9, 10, and 40 of the 757 Technical Supplement Manual for additional details.)

This pair of data wires, carrying TTL voltage levels, is the only connection to the 757 radio that's necessary. But in order to operate the TWFS, a power source providing 5 volts is also required. One additional connection to the 757 back panel provides 13.8 vdc at 800 mA through an RCA phono plug. Using a three-terminal fixed regulator to convert the 13.8 volts provided by the transceiver to a regulated 5 volts, the prototypes were measured to draw 165 mA from the radio. If no other accessories use this connection, it may be used to power the TWFS.

design details

The schematic for the TWFS is shown in **fig. 1**. The 555 provides the clock pulses to the 74LS393 at a frequency of 19.2 kHz. The internal Q_a and Q_b stages of the 393 divide the signal frequency by 4 and the remaining six count-stages provide a 6-bit binary number updated at the rate of 4800 Hz. The frequency adjustment of the 555 is done with a ten-turn rheostat. Frequency stability of the 555 is sufficient in this application because of the short bursts of data that are sent. For instance, assume that the Yaesu can tolerate a 1/8 bit error in order to read the data correctly. A 1/8 bit error on the 50th bit implies 0.25-percent relative error, allowing a frequency range of the 555 from 19.152 to 19.248 kHz.

The 393 is configured as a 6-bit counter that automatically shuts itself off when it gets to a count of 50. Every time the load button is pressed, the counter is reset to zero and counts to 50 one more time. The 6-bit binary number generated by the 393 is interpreted by the three 74150 and one 153 multiplexers so that one of 50 TTL levels are sequentially provided at the output of the 153.

When no data is being sent (i.e., you haven't pressed the load button recently), both address lines to the 153 are high and the HALT signal on pin 13 of

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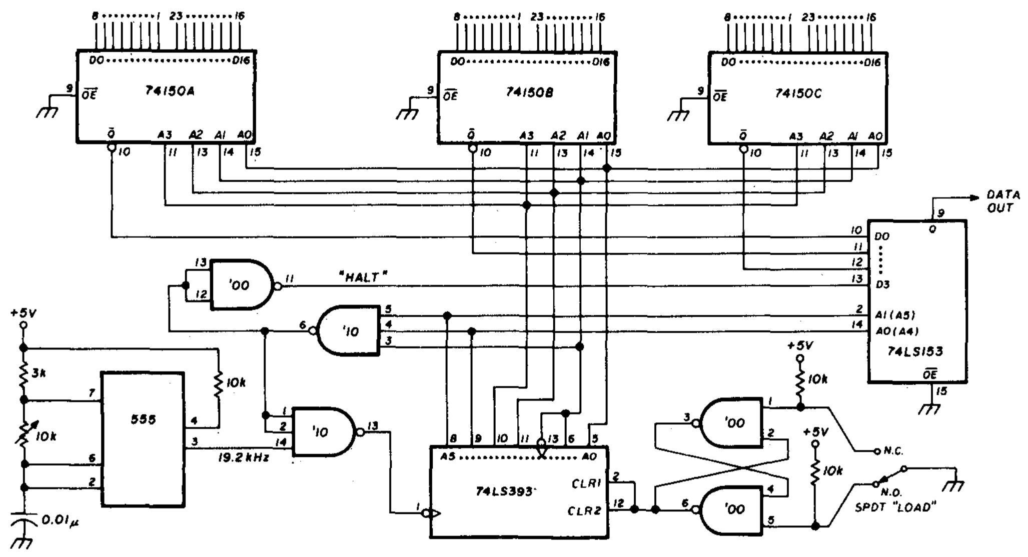


fig. 1. Schematic for the TWFS. Connections for these 48 pins are given in tables 2 and 3.

Table 1. Wiring and data description.

Bit no.	Data source	Wire to	Description	Bit no.	Data source	Wire to	Description
0	74150A – 8	0	partial bit				
1	– 7	1	start	26	74150B – 21	p20	1 MHz – 1
2	– 6	1	10 Hz – 1	27	– 20	p19	1 MHz – 2
3	– 5	1	10 Hz – 2	28	– 19	p22	1 MHz – 4
4	– 4	1	10 HZ – 4	29	– 18	p21	1 MHz – 8
5	– 3	1	10 Hz – 8	30	– 17	0	stop
6	– 2	p4	100 Hz – 1	31	– 16	1	start
7	– 1	p3	100 Hz – 2	32	74150C – 8	p24	10 MHz – 1
8	– 23	p6	100 Hz – 4	33	– 7	p23	10 MHz – 2
9	– 22	p5	100 Hz – 8	34	– 6	p26	10 MHz – 4
10	– 21	0	stop	35	– 5	p25	10 MHz – 8
11	– 20	1	start	36	– 4	1	100 MHz – 1
12	– 19	p8	1 kHz – 1	37	– 3	1	100 MHz – 2
13	– 18	p7	1 kHz – 2	38	– 2	1	100 MHz – 4
14	– 17	p10	1 kHz – 4	39	– 1	1	100 MHz – 8
15	– 16	p9	1 kHz – 8	40	– 23	0	stop
16	74150B – 8	p12	10 kHz – 1	41	– 22	1	start
17	– 7	p11	10 kHz – 2	42	– 21	1	
18	– 6	p14	10 kHz – 4	43	– 20	0	
19	– 5	p13	10 kHz – 8	44	– 19	1	
20	– 4	0	stop	45	– 18	0	
21	– 3	1	start	46	– 17	1	
22	– 2	p16	100 kHz – 1	47	– 16	1	
23	– 1	p15	100 kHz – 2	48	– HALT –	command	
24	– 23	p18	100 kHz – 4	49	– HALT –	to load	
25	– 22	p17	100 kHz – 8	50	– HALT –	frequency	

NOTE: The column headed "Wire to" lists either a header pin number, a "0," or a "1." Header pin numbers 1 and 2 are grounded. Each of the remaining 24 header pins must have a 10k pullup resistor. "0" bits are wired to ground and "1" bits are wired to the 580 – ohm pullup resistor. The column headed "Description" indicates how the bit will be interpreted by the 757 transceiver. Remember that the 150s invert the data.

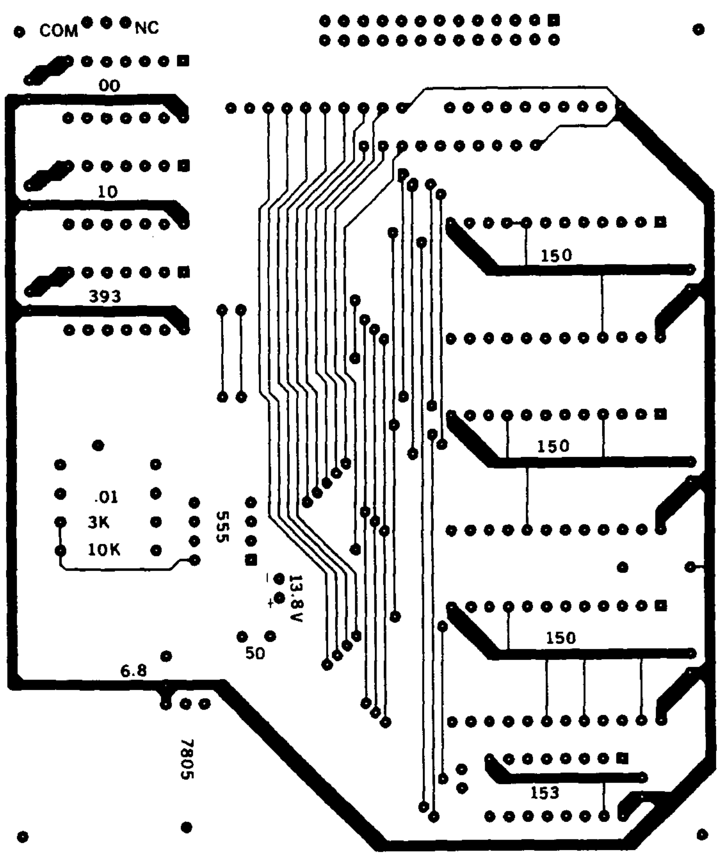


fig. 2. Upper (component) side artwork of frequency selector circuit. Be sure to use IC sockets that allow soldering on the top side of the board.

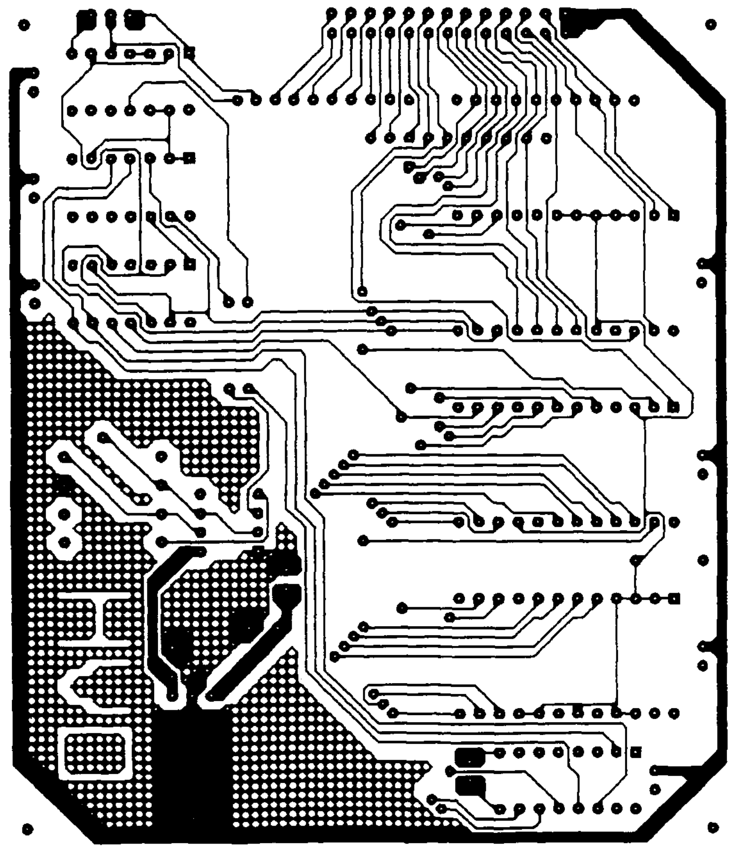


fig. 3. Lower (major foil) side artwork of frequency selector circuit.

Components for thumbwheel frequency selector.

Integrated circuits
1 7805
1 555
1 74LS00
1 74LS10
1 74LS393
1 74LS153
3 74150

Resistors (all 1/4 watt)
1 580-ohm (pullup for 17 TTL lines)
26 10k (thumbwheel and two switch pullups, 3x9 SIPS)
1 10k miniature ten-turn potentiometer
1 3k (555 timing)
1 10k (555 timing)

Capacitors
7 0.01 μF (IC decoupling)
1 0.01 μF (555 timing, mylar)
1 50 μF (decoupling of 13.8-volt supply, electrolytic)
1 6.8 μF (decoupling of 5-volt supply, tantalum)

Hardware
1 7805 heatsink
1 Pushbutton switch, 1-pole, 2-throw
6 BCD thumbwheel switches
1 RCA phono plug (power, 757 connector)
1 3-pin SIP socket (0.1-inch spacing, 757 connector)
1 26-pin, 0.1-inch DIP header and socket
1 3-pin 0.1-inch SIP header and socket
2 2-pin 0.1-inch SIP header and socket
1 Cabinet

the 153 is available as data — a TTL high level. As soon as you press the load button, the 393 counter clears to zero and pin 8 of 74150A is sent out as data. Because your button presses aren't synchronized with the clock, this first bit of data may or may not be held for a full 208.3 μsec (1/4800). That's OK because it's a TTL high signal — the same as the quiescent voltage level. In effect, the bit sequence has started but the 757 doesn't care that the first bit is too short because it doesn't recognize the TTL high level as the first bit.

The next 40 bits (bits 1-40) encode the thumbwheel-selected frequency digits. Data is sent in four 8-bit chunks. Each chunk is prefixed with a start bit (TTL low) and followed by a stop bit (TTL high), yielding a total of 40 bits. Within each 8 bits, two BCD numbers are sent least significant bit first. These numbers are read directly from the contacts of the thumbwheels, necessitating selection of thumbwheel switches with BCD outputs. The frequency is sent least significant digit first. The 100-MHz digit and the 10-Hz digit are permanently wired as zero. The 10-MHz through 100-Hz digits are read from the thumbwheels.

The last 8-bit value sent is decimal 10, which indicates to the transceiver that all the previous data should be interpreted as a frequency. The start bit and the first 6 bits of this data are retrieved through the 150/153 multiplex chain as all other bits were. The last 3 bits (0,0,1, sent in that order) actually reflect the status of the counter HALT signal. During the last 3 bits (48-50) the HALT line, which goes high to disable the counter when a count of 50 is reached, is fed through the 153. During counts 48 and 49, it is false (TTL low); on the 50th count, it goes true (TTL high). Because 555 pulses are locked out with this same signal, the data line stays indefinitely high.

Note that the 150 chip inverts data before passing it to the 153. This is the way most BCD switches work — grounding the logical true pins. Bits 0-47 are sent through the 150s and so are wired inverted. Bits 48-50 (the HALT line) are not inverted by the 153 alone.

construction

The artwork for a double-sided circuit board is shown in **figs. 2** and **3**. A parts list is provided. The entire prototype was built using available components, though substitutions may be made. The only critical parts values are in the 555 timer circuit, and even these values could be recalculated to allow use of what you have in stock, following the formulas provided in manufacturers' data books. Although low-power Schottky integrated circuits are called for wherever possible, standard TTL chips would suffice. The power supply drain will increase, but no difference in operation will be noticed.

The circuit is simple enough to be wire-wrapped by hand; in fact, the prototype was wired on two small prototype boards with all the little holes. Provision is made on the pc board for a 7805 voltage regulator and its two bypass capacitors, assuming the 13.8 volt supply from the radio will be used. This portion of the circuit is not shown on the schematic. Refer to a manufacturer's data book or the ARRL's *Radio Handbook* for details about the 7800 family of regulators. The connections to the 48 data lines of the 150s are listed in **table 1**.

Power connections and unused pins for all ICs are tabulated in **table 2**. I advise the following general order of construction and check-out. Using IC sockets, wire the entire circuit, including all non-IC components. Component placement — following the markings on the circuit board — should be straightforward. Each of the IC positions is marked with a square pad to indicate pin 1. Three columns of ten holes provide a mount for the SIP resistor packages. If SIPs aren't available, mount nine 10k resistors on end, then connect all the tops of the resistors together and into the tenth hole (5 volts). Three holes are provided for a horizontal-style ten-turn potentiometer. Two holes relatively close together provide a mount for a radial lead, 13.8-volt decoupling capacitor, but a larger axial lead capacitor can be mounted by doubling one lead back next to its casing.

Four off-board connections need to be made. The power and data connections are each made with two wire cables. The load switch connects to the board

Table 2. Power connections and unconnected pins for ICs.

IC	+5 volts	Ground	Not Connected
555	8	1	5
74LS393	14	7	3,4
74LS10	14	7	8-11
74LS00	14	7	8-10
74LS153	16	8	1,3-7
74150	24	12	None

Table 3. Bit sequences for two sample frequencies: (A) 10.000 MHz, and (B) 29.999 MHz. The data sent by the TWFS will be interpreted by a normal ASCII computer terminal as the characters listed. Ten^{H} bit groups must be sent with a maximum of 100 msec betwwen them in order to be interpreted correctly by the transceiver.

(A) 10.000 MHz

bits 1-10:	0 0000 0000 1	cntrl-@
11-20:	0 0000 0000 1	cntrl-@
21-30:	0 0000 0000 1	cntrl-@
31-40:	0 1000 0000 1	cntrl-A
41-50:	0 0101 0000 1	line feed

(B) 29.999 MHz

bits 1-10:	0 0000 1001 1	cntrl-P (8 bit set)
11-20:	0 1001 1001 1	cntrl-Y (8 bit set)
21-30:	0 1001 1001 1	cntrl-Y (8 bit set)
31-40:	0 0100 0000 1	cntrl-B
41-50:	0 0101 0000 1	line feed

with three wires. Each of these connections can use 0.1-inch space header strips for neatness. The connection to the thumbwheels is made with a 26-pin 0.1-inch space dip header. Twenty-five pins are needed (24 BCD data lines and one common ground line); they're listed in **table 1.**

Install the 7805 regulator. A mounting hole is provided for laying the regulator flat on its back, but it can be left upright if you wish. In either case, a heatsink is suggested. Apply power and verify that the correct voltages are applied to the proper IC socket pins. With the power off, insert only the 555. Then — with the power back on — check to see that pin 14 of the 74LS10 socket is receiving an oscillating TTL signal. Adjust the ten-turn potentiometer until a frequency of 19.20 kHz is obtained. Turn the power off.

Install the 393, the 10, and the 74LS00. Apply power again and confirm that pins 2 and 12 of the 393 are stable at a TTL low level. Pressing and holding the load switch should make these pins go to a high level. As you're holding in the switch, pin 1 of the 393 should be receiving an oscillating signal. Release the load switch. Pin 1 should stop oscillating in approximately 10.4 msec (the time needed to send 50 bits). If a digital event counter is available, measure pin 5 of the 393. It should be low while the load button is depressed and count 25 low-to-high transitions when the load button is released. Many frequency counters can be used as event counters by locking their count gate open.

With the power off, install the remaining four ICs. Turn on the power one last time, using the plug on the back of the radio if that's your final intention. At this point, the data output (pin 9 of the 153) should be stable in a TTL high state and a valid 5-byte sequence of data should appear whenever the load switch is pressed and released. Connect the ground side of the data connection to the left side and the data line itself to the center pin of the three-pin connector on the back of the 757 (J12). Dial a valid Amateur band frequency on the thumbwheels and press the load button. The radio should switch directly to that frequency!

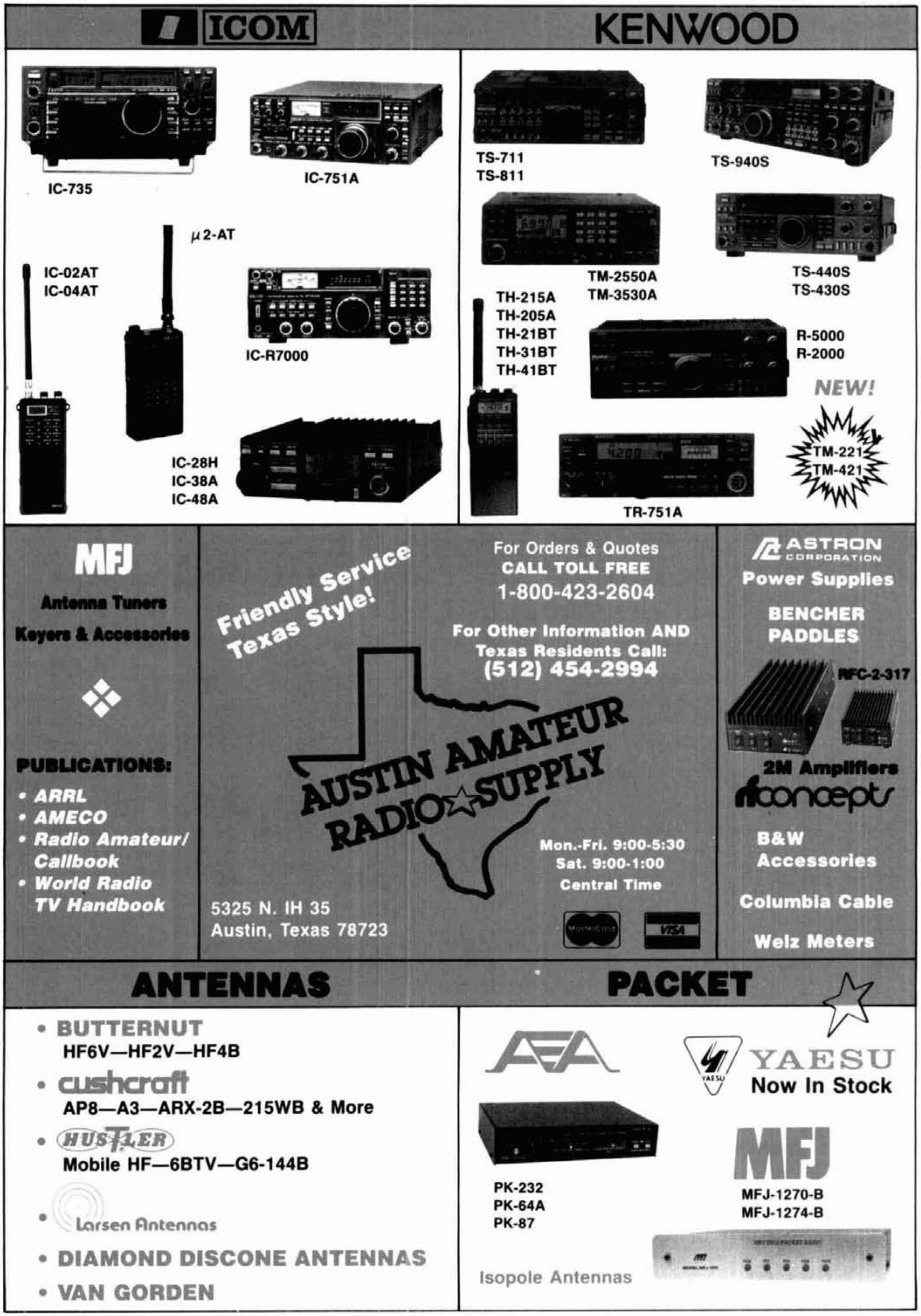
If the transceiver doesn't switch to the requested frequency after you release the load button, most likely one of the data lines into the 150 multiplexers has been wired incorrectly. If, after careful checking, everything looks correct, the following procedure may reveal the problem. Disconnect pin 3 of the 555 from its IC socket (bend it out horizontally) and plug a slowly oscillating TTL signal (about 4 Hz) or a low-pulse generator into pin 3 of the socket. Press and release the load switch. A TTL high level should be available as output data. Every four oscillations of the signal generator should cause pin 5 of the 393 to change and the data line to the radio should update. Set the frequency to 100000 and confirm the bit sequence shown in **table 3(A).** A frequency of 299999 should provide the bit sequence shown in **table 3(B).**

conclusion

This should be a fairly simple project for anyone who has worked — even just a little — with digital logic circuits. The idea of simply multiplexing hard-wired data sequentially out a data line is about as simple a communication scheme as possible. I hope others will expand this concept to enable a computer to provide data to the radio. I know useful commercial programs are available, but designs that add greatly to operating convenience don't need to be difficult to do yourself!

Ironically, although I've had fun designing the circuit and building several copies of it, I no longer have a 757 with which to use it. Therefore, I wish to express thanks to Gary LaPook, KA9UHH, for the use of his radio during the design of the project and for proofing the first prototype under actual use conditions. *There's a hamfest coming up. I wonder what a used 757 might cost?*

ham radio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ham radio TECHNIQUES

Bill Orr W6SAI

a new country for you?

We were sitting on the beach in front of the luxury hotel. As cool tradewinds caressed us, we enjoyed a second tall, iced glass of Long Island Iced Tea (1/3 each vodka, gin, and white rum, with a splash of cola for color).

"How about going on a DXpedition to a new country?" I asked The Big DXer.

"No way," he replied. "Look at the cost of the Bear Island trip. Activating a new country is too expensive these days."

I gestured toward the horizon. "There's your new country — only a few miles away. You can spend your nights right here in this dandy resort, living it up, and helicopter over and back each day. A 15-minute trip to a new country! Think how many hams you'd make *very* happy!"

The Big DXer stared at the island. "It looks easy," he said. "What's the trick? Why hasn't it been activated before?"

"Well," I replied, "there's a little problem . . . ."

The barren island of Kahoolawe lies a scant dozen miles south of the island of Maui, Hawaii. Uninhabited, it has no water or electric power. Jurisdiction rests with the United States Navy, much as in the case of Midway Island; the same reasons that Midway is accepted as a separate DX country, then, should be applicable to Kahoolawe.

The "little problem" is that this island is known as the "bombing island" because it's used by the military for bombing and assault practice. The whole surface of the island, it is said, is covered with unexploded ordinance. There's a boat dock, however, with an area around it that's considered safe for occupancy. Armed with a gasoline generator, a transceiver, and a portable antenna, DXers could conceivably operate from this area — *if* they could obtain permission to land.

Squinting at the island, The Big DXer admitted, "It would be nice to know who in the Navy to approach. Maybe a DXpedition to Kahoolawe is in the cards after all. I'll have another Long Island Iced Tea."

the 10, 18, and 24-MHz bands

As a result of an ITU conference some years ago, Radio Amateurs were granted operating privileges in narrow bands in the 10, 18 and 24-MHz region. Let's look at these bands as they appear in the fall of 1987, together with some simple antennas that will get you on these bands quickly.

Although the 50 kHz-wide *10-MHz band (10.100 to 10.150 MHz)* has enjoyed a modest amount of CW activity over the years, it's still largely ignored by most Amateurs. One or two loud commercial stations operating in this range must be avoided, but otherwise the band is good for DX — if activity can be found.

It seems to me that in order to awaken general interest in the band, a portion of it should be opened to sideband operation. I therefore propose that the top 25 kHz be opened to sideband. Before the band became available for general Amateur use, I operated sideband on 10.105 MHz with an experimental license for several months; I can attest to the fact that allocating even a small segment of this assignment to sideband would be of great benefit to Amateur Radio. How about a petition in this regard, ARRL?

Although available to Amateurs in over 40 countries, the *18-MHz band (18.068 to 18.168 MHz)* is barred to United States hams because of alleged use by the military. But a number of operators monitored this frequency region for several months during the early summer of 1987 — see **fig. 1** for a summary of the results of their work.

There's a small amount of overseas Amateur CW operation from 18.068 to 18.075 MHz. Sometimes there's a RTTY signal on 18.070. When it's ac-

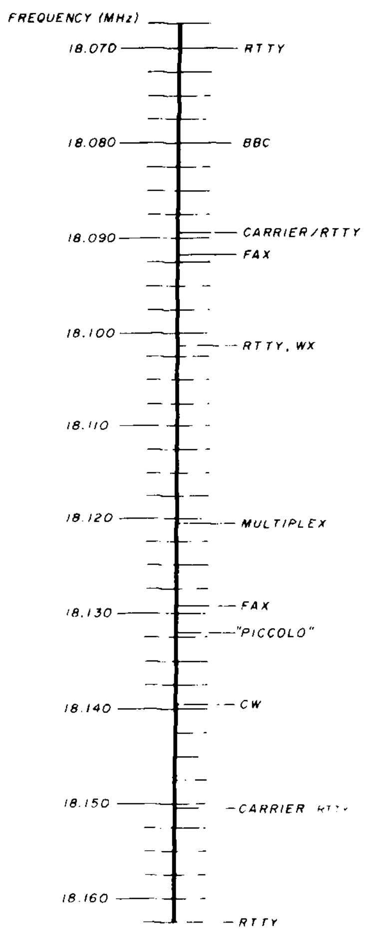

fig. 1. Spread of non-Amateur signals in the 18.068- to 18.168-MHz range.

tive, that frequency is avoided. This signal isn't on often, however, and when it's on, it's usually in "idle" mode. The BBC overseas service is quite active on 18.08, and there's a cluster of RTTY and FAX activity from 18.088 to 18.093 MHz.

Two RTTY weather signals are quite active near 18.104, with a multiplex signal slightly higher than 18.120 MHz. Near 18.130 there's a FAX transmission with a "piccolo" signal just above it. There are CW signals near 18.140 and 18.150. At the top of the range, there are several RTTY stations near 18.165 MHz.

It isn't clear that any or all of these stations are United States military. *In fact*, some of the signals sound as if they're arriving from overseas. None of them are on continuously except for the BBC relay transmission on 18.080 MHz, which is very active.

Amateur sideband operation takes place around 18.10 MHz. Stations in Europe, Africa, South America, and Australia have been logged from time to time.

After observing the 18-MHz band for over four years, it seems to me that the military's need for this span of frequencies makes for a pretty thin argument. I propose that the burden of proof of usage be placed on those opposing the presence of United States Amateurs in this band. Merely stating that the band is in use by the military, without proof of occupancy, *should not justify the FCC's withholding* the band until 1989.

Australia had several government and military assignments in this frequency range. When they opened the band to VK hams, they merely identified certain small spots as "off limits" to Amateur operation. Our government could do the same thing if it wanted to!

The 24-MHz band (24.89 to 24.99 MHz) is in good use and, as the sunspot cycle rises, will quickly become one of the better DX bands. Overseas DX signals pound in, even from stations running low power. Make sure you operate this band during the fall and winter DX season!

You can get on the 10, 18, and 24-MHz bands quickly with a dipole antenna cut to the dimensions shown in **fig. 2**. Even if you can't transmit on these bands, I urge you to listen, just to get the "feel" of propagation in these interesting portions of the radio spectrum.

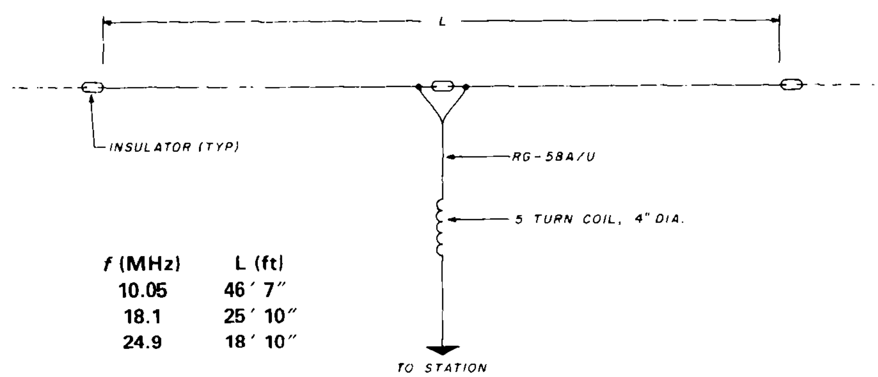

fig. 2. A "quickie" dipole for 10, 18, or 24 MHz. Feedline is wound into choke at point it attaches to antenna. Waterproof coax end at antenna.

W0SVM's mini-dipole

Amateurs have used miniature antennas on 40 meters for years. The smallest antenna that comes to mind is the loaded 40-meter mobile whip. Some hams who've placed two of them back-to-back to form a compact dipole (about 16 feet long) have found that the mini-antenna works quite well if it's mounted up in the air.

Even so, the 16-foot antenna is still too big for some hams who are unlucky enough to be handicapped by location. How about an indoor 40-meter antenna?

Jack Sobel wrestled with this idea for some time and finally came up with the antenna shown in **fig. 3**. He decided to use a long, thin radiating coil, or helix. Many hams frown on the helix antenna, but Jack and others have had good luck with the configuration.

Since material was at hand to make an antenna about 2 feet long, he decided this length would be a good place to start. The idea was to wind the antenna into a helix, or coil, and then place a matching coil at the center. Because the radiation resistance of the helix was bound to be very low, Jack chose a tapped matching coil technique to approximate a match to a 50-ohm line. And since antenna *Q* would be very high, he decided to use a Transmatch at the station end of the transmission line.

construction details

The antenna can be wound on a cardboard mailing tube. Coils L1 and L2 are 10 inches long and wound at six turns per inch (60 turns per coil of No. 12 AWG insulated wire). Coil L3 is 5 inches long and wound at six turns per inch (30 turns) of No. 12 tinned wire. Coil L3 is wound in the direction *opposite* to the direction in which coils L1 and L2 are wound. The braid of the coax line is attached to the center point of L3, and the center conductor is tapped out until a match is found to provide the lowest value of SWR at the resonant frequency of the antenna. Resonance is established with the aid of a dip meter before the antenna is attached. The tap point is about 2/3 the distance from the center to one end of L3. (The exact point depends upon the coupling between the antenna and nearby objects and the exact spacing of the windings.)

The antenna is fed with an electrical half-wavelength of RG-58/U coax. The line can be coiled up if it's too long to fit the available space.

The passband of the antenna is quite sharp, and if operation across the 40-meter band is desired, a Transmatch is necessary at the transmitter end of the line.

The antenna should be sprayed with a transparent polyurethane material or "corona dope" (a cellulosic resin such as General Cement's No. 10-4702). So far, the antenna has been run only at low power levels; since there may be danger of corona discharge at the outer ends of the windings, it's best to use the spray coating to help prevent this.

The antenna should be mounted as high in the air as possible and placed in a position where it can't be touched. There's *very* high rf potential at the ends of the little antenna, and an unsuspecting person could get a nasty rf burn during transmitter operation.

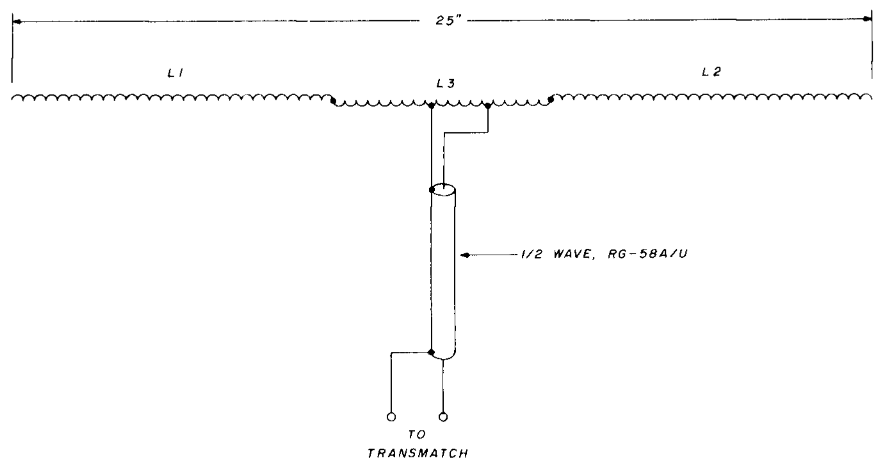

fig. 3. The mini-antenna of W0SVM. Adjust coax tap for best SWR at resonance.

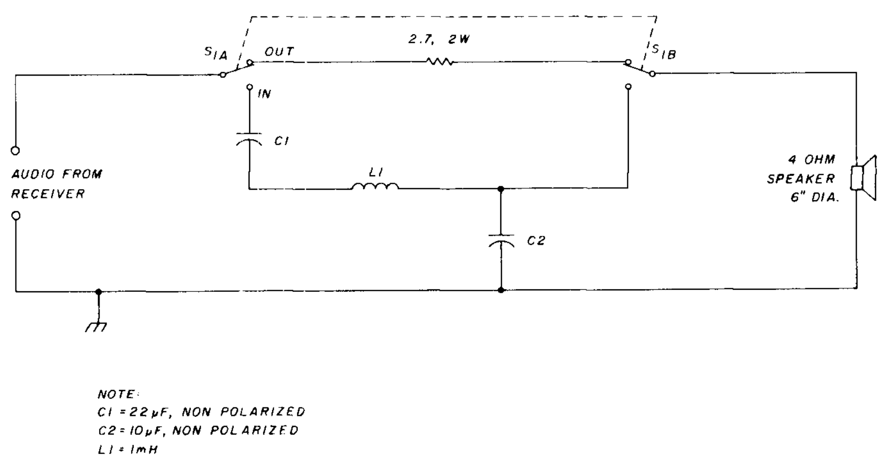

fig. 4. Communications network speaker system.

If more space is available, two mailing tubes can be epoxied together and the turns-per-inch increased so that the antenna covers more length along the longer tube. The winding will have to be readjusted to frequency with a dip meter if this change is made.

So far, Jack's contacts have been limited to a few hundred miles on 40 meters. He says it isn't as good as a full-size dipole, but it's small and can be used in an apartment.

a communications speaker system

Tired of listening to that annoying high-frequency "monkey chatter" that seems to sneak through the receiver filter system? It can be extremely tiring, especially in a DX contest. The VK boys seem to have a handle on that problem. Rodney Champness, VK3UG, writing in the March, 1987, issue of *Amateur Radio*, the publication of the Wireless Institute of Australia, shows a simple audio filter that can be placed in the speaker line (**fig. 4**) to substantially reduce frequencies above 2500 Hz. A switch on the filter allows normal wideband reception to be retained.

The 1-mH inductor should be able to carry the audio current in the speaker circuit (something less than 1 ampere).

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PRACTICALLY SPEAKING ... Joe Carr K4IPV

using voltage comparators

A voltage comparator is basically an operational amplifier (or derived from the op amp) that has no negative feedback network (see **fig. 1A**). The open-loop gain of the op amp is very high — on the order of 50,000 at the low end to more than 1,000,000 for many devices. Thus, with no negative feedback the operational amplifier functions as a very high-gain amplifier with an output that saturates with only a few millivolts input potential. For example, with a gain of 100,000, and a maximum output potential of 10 volts, the amplifier will saturate with only 10 volts/100,000, or 0.1-mV input.

So what use is an amplifier that saturates with only a few millivolts of input voltage? The comparator is used to compare two input voltages and generate an output that denotes their relationship. In **fig. 1A**, potential V_1 is applied to the inverting input, and V_2 is applied to the noninverting input. If $V_1 = V_2$, then $V_o = 0$. Otherwise, the output voltage obeys the relationships shown in **fig. 1B**. The transfer function of the comparator is shown in **fig. 1B**. According to the normal rules for operational amplifiers, when V_1 is larger than V_2 (see **fig. 1A**), it looks as if a positive input has been applied to the inverting input, so the output potential saturates at V−. Alternatively, when V_1 is smaller than V_2, it looks like a negative input potential, so the output is V+.

Typical Amateur examples include over- and under-deviation alarms on repeater receiver detector outputs, and over-temperature alarms for electronic equipment. For example, some form of alarm is needed at unattended repeater sites. The comparator can be used to let the control operator know by telephone or radio telemetry link that something is amiss. Other applications will occur to most readers in the light of their own special needs.

There is a small hysteresis band around zero, however, where no output changes occur. This is an unfortunate defect in practical op amps, and seems to fly in the face of the theory when that hysteresis band is larger than the potential needed to saturate the output terminal.

Several years ago, while working in a hospital electronics lab, I measured the hysteresis band on a number of operational amplifiers and IC comparators (LM-311). Not surprisingly, the 741-family devices had terrible hysteresis levels, on the order of 25 mV. The LM-311 devices had 8- to 10-mV hysteresis, which surprised me. Also surprising was the fact that then-premium devices such as the μA-725 had 10 to 20 mV of hysteresis (I haven't tested modern high-performance units). The overall best device was a non-premium device that is readily available to Amateurs and other hobbyists: the CA-3140 (a BiMOS operational amplifier), which uses the industry standard "741" pinouts, as shown in **fig. 1A**.

The LM-311 device (**fig. 2A**) is a low-cost voltage comparator in IC form. Although based on op amp cir-

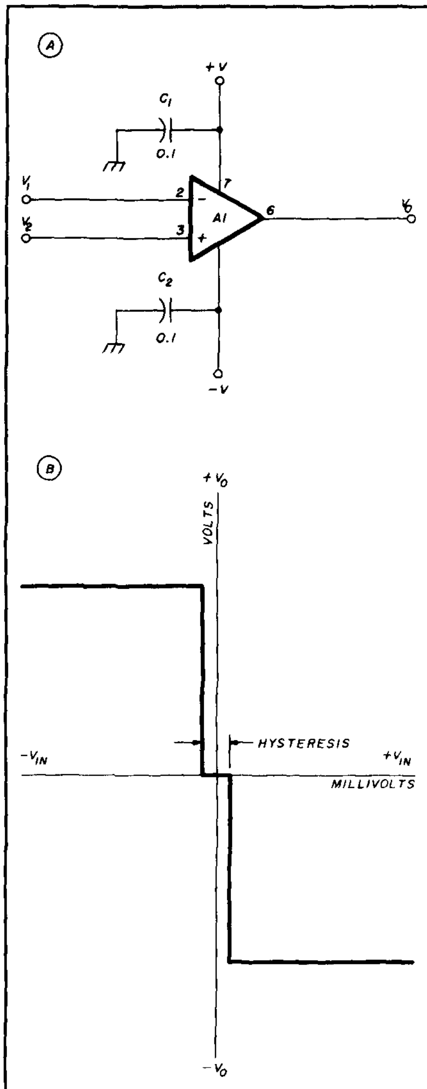

fig. 1. (A) Simplified diagram of voltage comparator; (B) comparator transfer characteristic.

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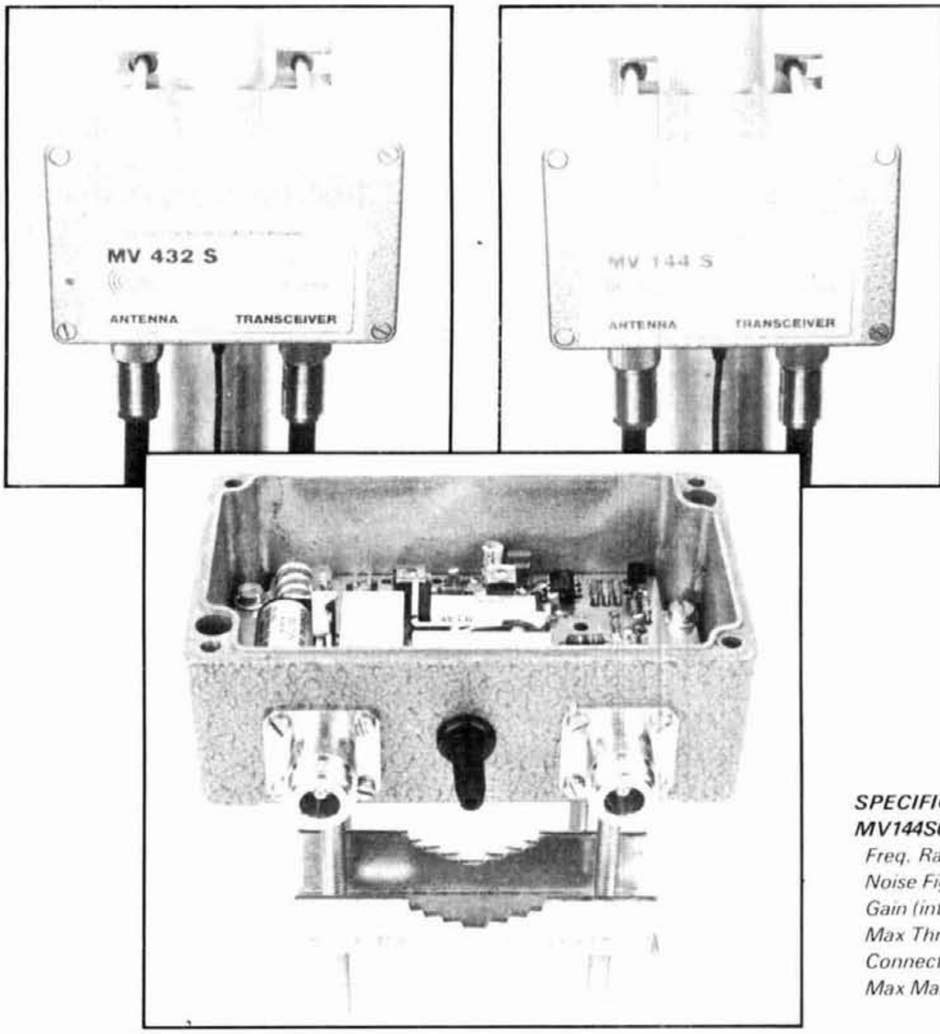

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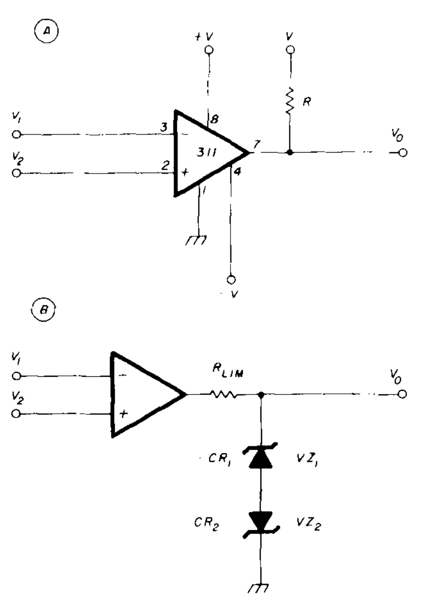

fig. 2A. LM-311 voltage comparator.

fig. 2B. Zeners improve transfer characteristic and speed, and limit output voltage of the comparator.

cuitry, this device is specifically designed as a comparator. Contrary to op amp practice, it has a ground terminal (pin 1) and requires an output pull-up resistor (R) to a positive voltage. The output terminal can drive loads such as relay coils, lamps, and LEDs to potentials of 40 to 50 volts (depending upon the type of device) and 50 mA. If the LM-311 is operated for compatibility with TTL digital logic, the pull-up resistor should be terminated at a +5 VDC potential, and R should be 1 to 3.3 k.

A means for limiting the output level, improving the sharpness of the transfer function corners (see **fig. 1B**), and improving speed by reducing latch-up problems, is shown in **fig. 2B**. In this circuit, two zener diodes are connected back-to-back across the output line. When the output voltage is HIGH, it is then limited to V_{Z1} + 0.7 volts; when LOW, it is V_{Z2} + 0.7 volts. These potentials represent the zener voltages of CR_1 and CR_2, plus the normal forward-bias voltage drop of the alternate diode.

Figure 3 shows a means for increasing the drive capacity of the comparator. In this circuit a switching transistor (a 2N3704, 2N2222, etc.) is used to control a larger load such as the relay coil shown here. The output voltage ($V_o$of the comparator)is used to set up the bias for the NPN transistor. When the comparator output is HIGH, the transistor is biased hard-on and the load is grounded. Alternatively, when the comparator output is LOW, the transistor is reverse biased and the load remains ungrounded.

The diode across the relay coil is essential for any inductive load. When the magnetic field surrounding a coil collapses the counter EMF generates a high-voltage spike that is capable of damaging components or interrupting circuit operation (especially digital circuits). Though the diode is normally reverse biased, it is forward biased for the counter EMF spike. The diode therefore clamps the spike to about 0.7 volts.

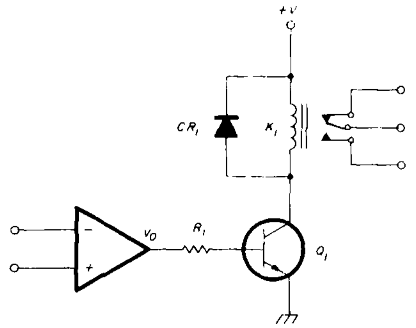

fig. 3. Addition of switching transistor allows higher current loads to be controlled.

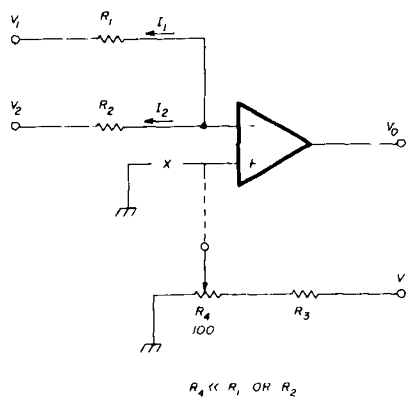

fig. 4. Two methods of reducing the hysteresis band effect: zero offset control and current mode operation.

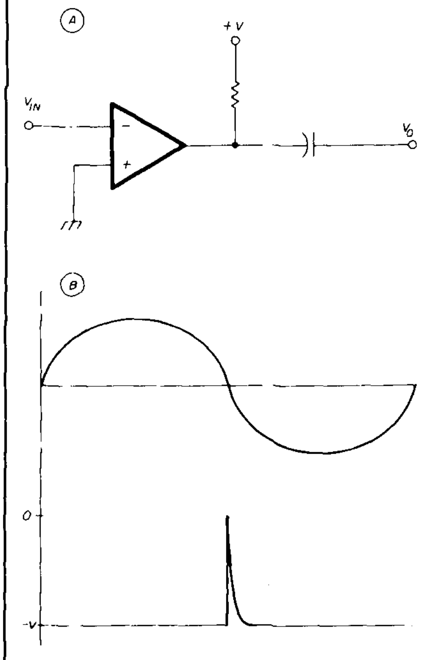

fig. 5. Zero crossing detector circuit and waveforms.

Figure 4 shows two methods. One is a zero-offset control used to reduce the effects of the hysteresis band, while the other is the so-called current mode. The offset control (R_4) slightly biases one input to a non-zero level so that it's ready to trip when the other input is non-zero. In this particular case the inverting input is grounded (V_2 = 0), but could as easily be a non-zero voltage.

Current mode operation is usually faster and less prone to latch-up than voltage mode. For this reason, current mode comparators are sometimes used in high-speed analog-to-digital converters (A/D). Assume that the noninverting input is grounded. In this case, the output potential V_o will reflect the relationship of the two currents. If $I_1 = I_2$, then $V_o = 0$. This circuit is, to the outside observer, a voltage comparator in that $I_1 = V_1/R_1$ and $I_2 = V_2/R_2$. Of course, it's also useful for current output devices such as the LM-334 temperature monitor IC.

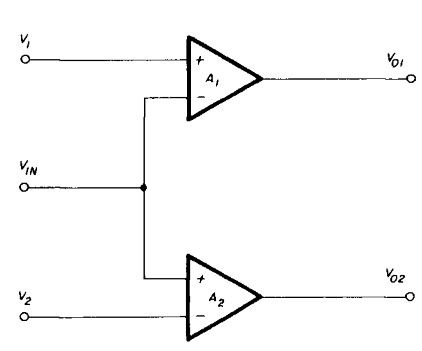

fig. 6. Window comparator triggers on two different reference levels being exceeded.

Figure 5 shows a zero-crossing detector circuit. In this case a comparator is connected with its noninverting input grounded. When V_{in} is non-zero, then the output will also be non-zero. But when the input voltage crosses zero, the output briefly goes to zero, producing the differential output pulse shown.

A window comparator is shown in **fig. 6**. This circuit consists of two comparators connected so that one or the other input is activated when the input voltage (V_{in}) exceeds either positive or negative limits. The limits are determined by setting V_1 or V_2 reference voltages. An over- and under-deviation alarm, as heard on some repeaters, is a possible application for this circuit. The output of the fm demodulator is a voltage that is proportional to deviation, so this circuit is a natural for that use (although the demodulator voltage will probably require some amplification). The circuit is also used for over- and under-temperature alarms, and other such applications where a range of permissible values exists between two forbidden regions.

Figure 7A shows a method for biasing either input to a specific reference voltage. Although in this case the noninverting input is biased and the inverting input is active, the roles can just as easily be reversed. Two methods of biasing are used: resistor voltage divider and zener diode. If R_2 is replaced with a zener diode, then the reference potential is the zener potential. In that case, R_1 is the normal current-limiting resistor needed to protect the zener from self-destruction. In the case where a resistor voltage divider is used, the bias voltage V_1 is set by the voltage divider equation:

$$V_1 = \frac{R_2\,(V+)}{R_1 + R_2} \qquad (1)$$

For example, suppose $R_1 = R_2 = 10k$, and $V+ = 12$ VDC:

$$V_1 = \frac{(10k)\ (+12\ VDC)}{(10k + 10k)}$$

$$V_1 = \frac{120\ volts}{20k} = 6\ volts$$

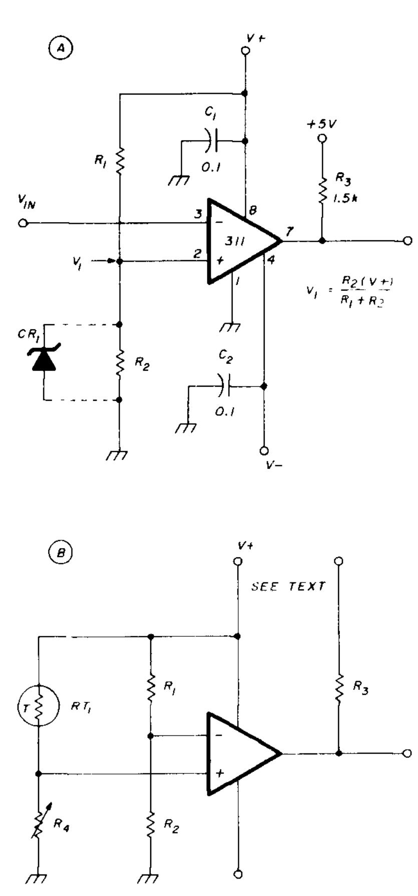

fig. 7. (A) Method of biasing comparator to a specific reference voltage; (B) over-temperature sensing circuit is biased by R_1/R_2 resistor combination.

Figure 7B shows an over-temperature circuit based on **fig. 7A**. In this circuit the inverting input is biased by R_1/R_2, while the noninverting input is set by another voltage divider, R_4/RT_1. Resistance RT_1 is a thermistor, which has a resistance proportional (or inversely proportional in some types) to the temperature. Potentiometer R_4 is used to set the trip point temperature. The values of the resistors depend upon the set trip point desired and the resistance of the thermistor over the range of temperatures being monitored.

an invitation

I'd like to hear what you think of this column. I also welcome your suggestions for future topics. You can reach me at P.O. Box 1099, Falls Church, Virginia 22041.

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winter DX

With the fall DX season (September and October) behind us, the winter DX season (November through February) is about to begin. Wintertime DX is characterized by:

- Better signal strengths on all bands most of the time — but especially during daytime on the lower frequencies, particularly in low sunspot years.
- Nighttime DX openings earlier in the evening.
- More frequent transequatorial paths, with higher MUFs and longer distances.
- Lower incidence of local thunderstorm-generated QRN conditions.
- More stable signal strengths and fewer geomagnetic field disturbances.

There are two main reasons for the first four characteristics: the tilting of the Earth's axis away from the sun, which results in shorter, colder days; and less ionization in the lower ionosphere, which results in less energy absorption as signals travel through the D region.

The amount of absorption per hop is related to the zenith angle to the sun at the location of each D region crossing. In working DX, it pays to use a higher frequency band to obtain more distance per hop (resulting in fewer transits) for less total signal loss. This is why we generally think of 6, 10, or 15 meters for DXing.

But in winter, particularly near sunspot minimum, we have the opportunity to work DX on the lower frequency bands with lower signal loss, day or night, than at any other time of the year. You can't always count on it, however; signals traveling a high latitude path may be poor for several days at a time. This is known as the *winter anomaly*.

Along with lower signal attenuation, QRN decreases as fewer local thunderstorms pass through. As the large thunderstorm areas near the equator move further south, their noise decreases by about 6 to 8 dB. This is particularly noticeable on the 160, 80, and 40-meter bands.

Even though ion production in the D, E, and lower F regions is lower, ions are better able to diffuse and drift upward along the geomagnetic field lines into the F region. This layer is the major factor in defining the maximum usable frequency and maximum on each side of the geomagnetic equator, as shown in my October, 1983, column. These maximums, which are reached most evenings at about 2200 local time, eliminate one whole earth bounce and its accompanying double-D region transits for one-long-hop propagation.

The fifth characteristic of winter DX conditions — the increased stability of signal strengths and the decrease in the number and intensity of geomagnetic field disturbances — is attributable to the eccentricity of the Earth's orbit. When the Earth is closer to the sun, the solar flux pressure on the magnetosphere surrounding the earth tends to hold the magnetosphere steadier. This means that the geomagnetic field is least disturbed during November and December. This manifests as least variation of the magnitude and direction of the geomagnetic field lines in an hour's time, translating into fewer periods of QSB during these months.

last-minute forecast

The higher-frequency bands are expected to be best during the first two weeks of the month because of the probability of higher solar flux at that time. The new 11-year solar cycle is also expected to be well underway, with higher solar flux from the more active regions. More potent geomagnetic storms may accompany this increased activity, but they're not really expected during this quiet time of the year and at this point in the sunspot cycle. The most probable times of the month are November 2, 11, and 20. The lower bands are expected to be best the third week of the month. Thanksgiving weekend should be a good one for the whole frequency range.

The Taurids meteor showers will occur from October 26 to November 22, with a maximum count of ten per hour from the 3rd through the 10th of November. Lunar perigee is on the 5th.

band-by-band summary

Ten and twelve meters, the highest day-only DX bands, are nearest the MUF for southern hemisphere paths. They will be open most days when the solar flux is above 75 during the 3- to 5-hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 4000 km (2400 miles) in single-hop length, and on occasion double that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variability than 10 and 12 meters. It will be best when the MUF is resting just above this band, until it drops below it — a transition period that occurs

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	*40*	*40*	*40*	*40*	40	40	40	40	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	40	40	40	30	30	40	N
EUROPE	40	40	40	40	40	30	20	20	20	20	30	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	40	40	40	NE
S. AFRICA	20	15	15	15	12	12	12	15	15	15	20*	*30*	*30*	*30*	*30*	*30*	*30*	30	30	30	30	20	20	20	E
S. AMERICA	12	12	10	10	12	12	12	12	15	15	20*	*30*	*30*	*30*	*30*	*30*	*30*	*20*	20	20	20	20	15	15	SE
ANTARCTICA	15*	15	15	15	15	15	15	15	15	20	*30*	*30*	*40*	*30*	*30*	*20*	20	20	20	20	20	15	15	15*	S
NEW ZEALAND	12	12	12	15	15	15	20	20	20	*30*	*30*	*30*	*30*	*30*	*30*	*20*	20	20	20	20	15	15	12	12	SW
OCEANIA AUSTRALIA	15*	15	15	15	20	20	*30**	20	20	*30*	*30*	*30*	*30*	*30*	30	30	20	20	20	20	15	15	12	12	W
JAPAN	20	20	20	20	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	30	30	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	*40*	*40*	*40*	*40*	40	40	40	40	30	30	*40*	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	40	40	40	N
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S. AFRICA	20	20	15	12	12	10	12	12	12	12	15	20*	*30*	*30*	*30*	*30*	*30*	*30*	30	30	30	30	30	20	E
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ANTARCTICA	15*	15*	15	15	15	15	15	15	15	20	20	*30*	*30*	*40*	*30*	*30*	*20*	20	20	20	20	20	20	15	S
NEW ZEALAND	12	12	12	15	15	15	15	20	20	20	*30*	*30*	*30*	*30*	*30*	*20*	20	20	20	20	15	15	12	15*	SW
OCEANIA AUSTRALIA	12	15	15	15	15	20	20	*30**	20	20	*30*	*30*	*30*	*30*	*30*	30	20	20	20	20	15	15	12	12	W
JAPAN	20	20	20	30	*40*	*40*	*40*	*40*	40	40	40	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	40	40	40	NW
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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	*40*	*40*	*40*	40	40	40	40	40	40	40	40	30	30	*40*	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	N
EUROPE	40	40	40	40	30	20	20	20	20	20	20	20	20	30	*40*	*40*	*40*	*40*	*40*	*40*	*40*	*40*	40	40	NE
S. AFRICA	20	20	15	15	12	12	10	10	10	12	12	12	15	20*	*30*	*30*	*30*	*30*	*30*	*30*	30	30	30	20	E
CARIBBEAN S. AMERICA	15	12	12	10	10	12	12	12	15	15	15	15	20	*30*	*30*	*30*	*30*	*30*	*20*	20	20	20	20	20	SE
ANTARCTICA	15	15	15*	15*	15*	15	15	15	15	15	20	20	*40**	*30*	*30*	*30*	*30*	*20*	20	20	20	20	20	15	S
NEW ZEALAND	12	12	12	15	15	15	20	20	20	20	20	20	*30*	*30*	*30*	*30*	*30*	*30*	30	30	20	20	20	15	SW
OCEANIA AUSTRALIA	15	15	15	15	20	20	*30**	20	20	20	20	20	*30*	*30*	*30*	*30*	*30*	*30*	*30*	30	20	20	20	15	W
JAPAN	20	20	20	30	*40*	*40*	*40*	*40*	*40*	40	40	40	*40*	*40*	*40*	*40*	*40*	40	40	40	40	40	40	40	NW

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

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

right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Twenty, thirty, and forty meters are both daytime and nighttime DX bands. Twenty is the maximum usable band for DX in the northern directions during the day. In combination with 30 meters, it provides nighttime paths for the day-only bands. Forty meters becomes the main over-the-pole DX daytime band, with some hours covered by 30 meters. This path may be affected by anomalous absorption during a few days of the month.

Eighty and one-sixty meters, the night-only DX bands, exhibit short-skip propagation during daylight hours, then lengthen at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the norh-south near midnight, and ending up in the Pacific areas for a few hours before dawn. Remember the DX windows of 3790 to 3800, 1825 to 1830, and 1850 to 1855 kHz.

ham radio

an rf voltmeter

Don't let its apparent simplicity fool you — this instrument has many uses

Many of us who experiment with circuits need to measure the level of signal sources such as oscillators, amplifiers and multipliers in transmitters, and local oscillator systems in receivers. The voltmeter design described in this article came about when I wanted to measure the voltage reflection coefficient of antenna systems using a return loss bridge at low levels so as not to interfere with other band users.

The common method of measuring signal levels is through the use of a simple diode detector. In its basic form, however, it has a number of shortcomings, some of which can be easily overcome.

voltmeter requirements

This voltmeter covers a range from less than 70 millivolts to greater than 3 volts rms (equivalent to -10 to $+23$ dBm in a 50-ohm system), covers a frequency range from 10 kHz to 150 MHz, and provides readings accurate to within ± 2 dB without calibration — i.e., as built and tested. Its input impedance is set by the input resistor; a value of 50 ohms was used in the models shown. Its output is linear; if an analog meter is used, no special marking of the meter scale is necessary. An external general-purpose meter can also be used. The meter draws less than 15 mA from a pair of 9-volt transistor batteries.

diode detectors

The characteristics of an ideal linear voltage detector are illustrated in **fig. 1**. This mythical device conducts current in one direction only, with a low and constant resistance when forward biased and an infinite resistance in the reverse direction. The constant forward resistance is maintained right down to 0 volts,

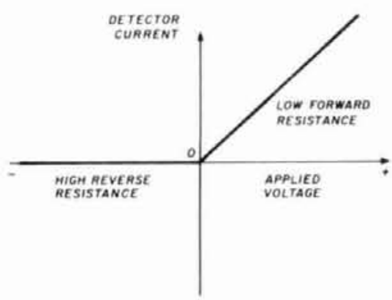

fig. 1. Ideal detector characteristics.

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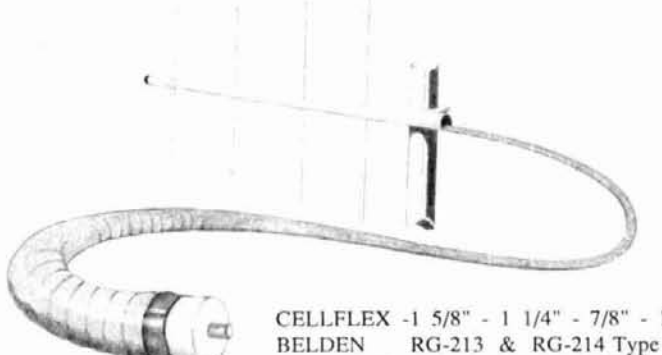

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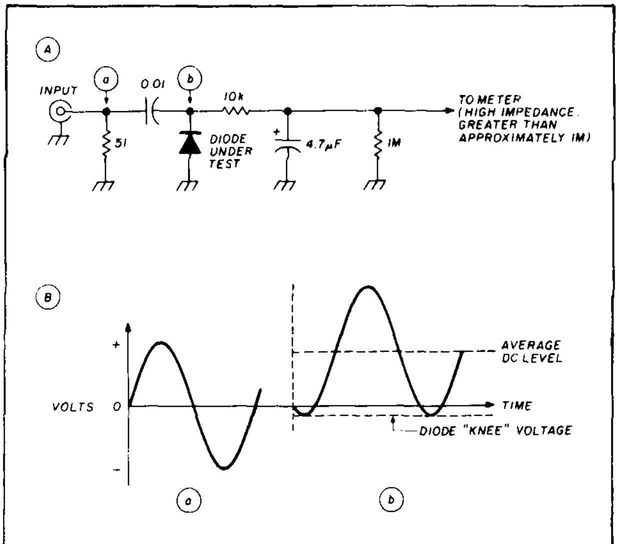

fig. 2. Unbiased single diode detector: (A) schematic (B) waveforms at "a" and "b."

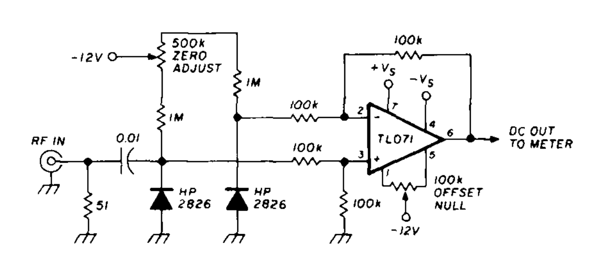

fig. 3. Two-diode biased detector.

with an abrupt transition to the reverse region. Such a device used as a rectifier would deliver a dc output proportional to the applied ac.

Real diodes, however, don't behave this way. Most do not conduct appreciably in the forward direction until the input voltage across them exceeds a threshold or "knee" voltage, which for an ordinary silicon junction diode is around 0.7 volts. The threshold voltage for germanium and Schottky diodes is lower — 0.2 to 0.4 volts. Real diodes also conduct slightly in the reverse direction (the so-called reverse leakage current).

The transition between the conducting and nonconducting states is not sharp, but occurs over a region where the diode is said to have "square law" behavior and the dc output is proportional to the applied *power* (voltage squared), rather than the signal voltage. This is used to advantage in low-level diode power meters.

To see how real detectors behave, I made some measurements on a few types using a crystal oscillator signal source at 10 MHz, and a power meter and attenuator to give a range of calibrated levels. **Figure 2** shows the first test circuit, a simple peak detector using an HP2826 Schottky diode. The right-hand plate of the input capacitor is clamped at the diode knee voltage below ground on negative input swings. If this knee voltage were actually zero, the average voltage on the diode would equal the peak of the input voltage, but the real diode produces less. The resistor and capacitor filter the rf present on the diode, leaving the dc component. This can be measured by a high impedance meter which reads the peak of the rf voltage minus the diode knee voltage.

A dc forward bias current can be used to improve the sensitivity of the diode detector. If the diode is fed from a high resistance with a current of a few μA, its forward junction voltage will sit around the knee voltage. This potential no longer has to be supplied by the rf, which sweeps the diode's nonlinear characteristic and is detected. Direct current bias is used in the more sophisticated circuit shown in **fig. 3**. Two diodes are used. Both are biased, but rf is fed to only one of them. An op-amp subtracts the diode voltages so that the output of the circuit can be set to zero in the absence of an rf signal. With the diodes connected together, with no rf, the op-amp offset is nulled. The 500-k pot is then adjusted to give zero output with the circuit exactly as drawn. The circuit works best with matched pairs of diodes, since these track well with temperature.

The performance of these detectors is described graphically in **fig. 4**. This shows the improvement in sensitivity achieved with bias. Also shown is the curve for the voltmeter design, which indicates further improvement in sensitivity and linearity, gained by using just one additional technique.

The complete rf voltmeter is shown in the block diagram in **fig. 5**. Two detectors are used. One receives the incoming rf signal, while the other, a "mimic" detector, is fed with a low frequency signal. This signal is an internally generated sinusoid, derived by chopping the dc output of an integrator to give a square wave which is then filtered, leaving the fundamental frequency component.

The integrator input is the difference of the two detector voltages, and its output will change (or "slew") when this difference (the feedback loop's error signal) is other than zero. The action of the negative feedback loop around the mimic detector in fact causes the integrator to try to achieve this zero-error condition, at which point, if the detectors are well matched, the low frequency signal will have the same amplitude as the rf signal. Because the low frequency signal is produced by chopping and filtering the integrator's dc output, this latter voltage is proportional to the rf input voltage, and can be scaled and metered to provide readings in "rf" volts.

Through the use of well-matched and closely spaced

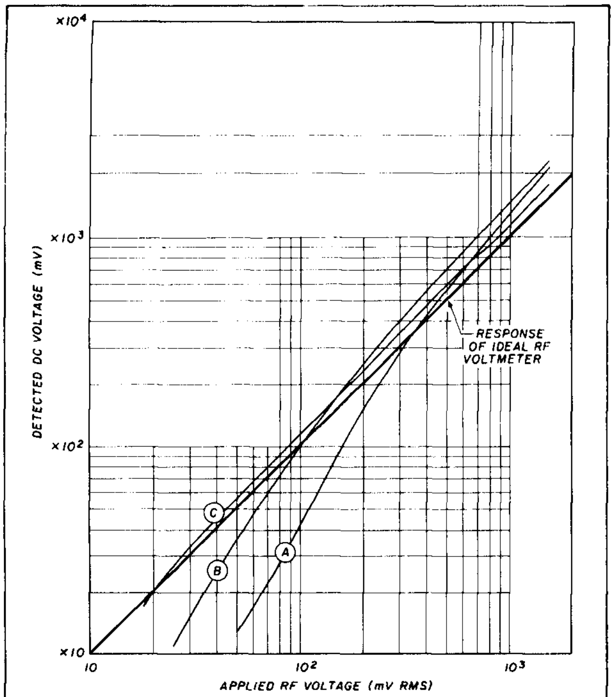

fig. 4. Detected dc output versus applied rf input: (A) single unbiased diode detector; (B) biased diode detector; (C) rf voltmeter; (D) ideal rf voltmeter.

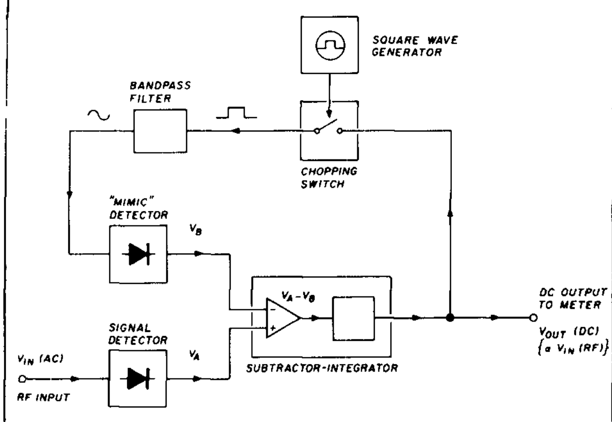

fig. 5. Voltmeter block diagram.

diodes, their temperature and I-V curve variations are minimized. Of course, since the mimic detector measures only a fixed low frequency signal, no frequency response compensation for the input detector is provided. Therefore, it's best to choose the diode that has the flattest possible response.

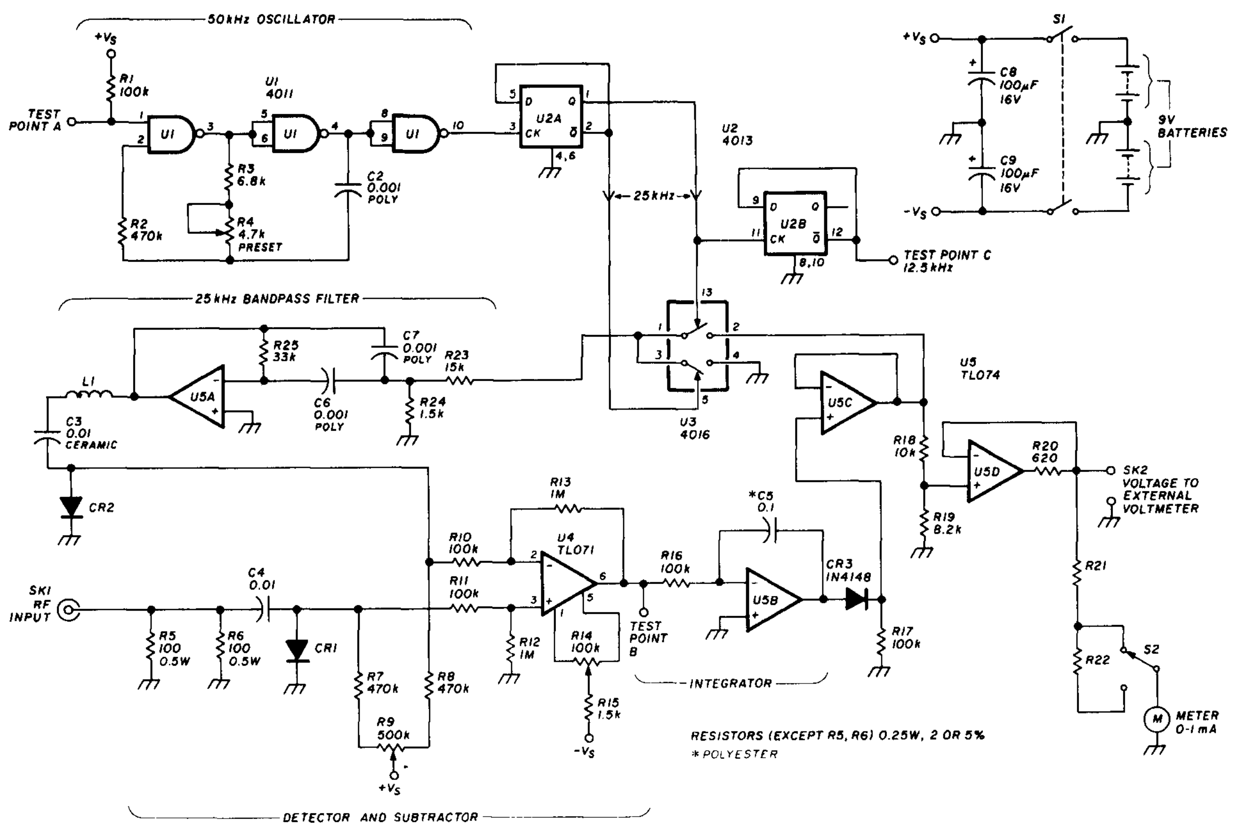

fig. 6A. Voltmeter schematic diagram.

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circuit description

RF enters the instrument via socket SK1. R5 and R6 provide a good impedance match to 50-ohm cable. Two 0.5-watt metal film resistors (or any other combination providing 50 ohms and a 1-watt rating) should be used. Diodes CR1 and CR2 form the detectors: CR1 is supplied with rf, and CR2 with the internally generated 25-kHz sine wave. They should be Schottky diodes and, if possible, should be reasonably well matched in terms of forward voltage at around 10μA. Many types will do, among them the HP2800 and 2826 and the Thomson BAR28. The forward voltage will be in the region of 250 mV, and a pair matched to within a few millivolts can often be found from a small batch. Circuitry around U4 performs subtraction of the two detector outputs with a gain of 10. R14 allows U4's offset voltage to be nulled. Because this is a relatively high gain stage, the remaining stages do not need to be nulled, and can be grouped into a quad package. U5B is wired as an integrator, and CR3 and buffer U5C prevent the CMOS switch U3 from being driven negative. The buffer's dc output is chopped by switch U3, which is operated by 25-kHz square waves from U2, a divide-by-2 flip-flop fed from U1, a 50-kHz oscillator. U2 could be omitted and U1 run at 25 kHz, but U2 does achieve a perfect square wave (1:1 mark-to-space ratio) at small cost. U2B provides a 12.5-kHz output signal so that a rough check can be made on the switching signal with a crystal earpiece at test point C. The chopped dc from U3, now a 25-kHz square wave with a peak-to-peak amplitude equal to the U5C dc output, is filtered by active filter U5A (second order bandpass) so that CR2 receives a fairly sinusoidal signal. L1 was included to stop hf oscillations in the output stage of U5A when the connection to the detector was completed via several inches of ribbon cable (the adjacent wire being grounded). It consists of four turns of enameled wire on a single-hole ferrite bead (I used an FX1115), and has no measurable effect at 25 kHz. R18 and R19, buffered by U5D, attenuate the dc by a factor of 0.45 (see **appendix** for derivation), which provides scaling to units of volts rms. An external voltmeter plugged into SK2 will then read the rf input voltage. R21 and R22 with switch S2 allow the use of a meter to read 1 and 10 volts full-scale. With 9-volt supplies from batteries, the maximum voltage that can be read will be around 3 volts.

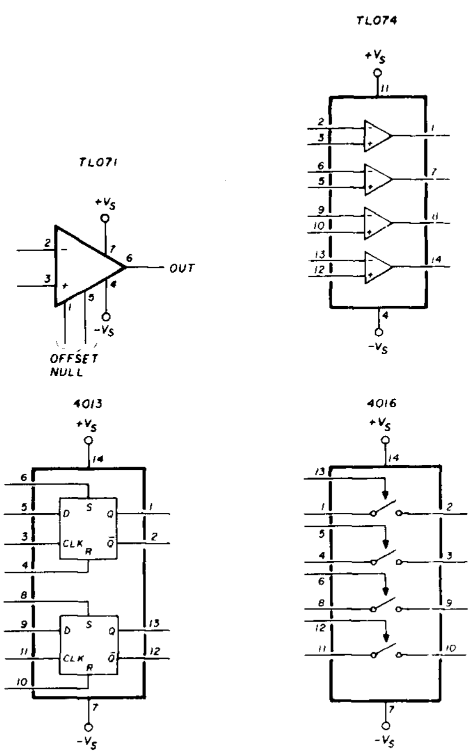


fig. 6B. Voltmeter integrated circuit pinouts.

construction

I have built three instruments according to the design described in this article (see **fig. 7**). The outer two are battery powered (internally); the "economy model" in the center uses an external power supply and meter. Details of the unit on the left are shown in **figs. 8A, 8B,** and **8C**. Construction is straightforward and can be done with ordinary hand tools. The only critical area is the detector, which carries rf. The other areas involve only low frequency circuitry. As shown in **fig. 8C**, I built the input circuitry, consisting of the two detector diodes and U4, on a small piece of double-sided, copper-clad glass-fiber board, using a counterbore tool to provide pads for the components. Frankly, this method of construction — with the components mounted rats-nest style above a copper ground plane, will work at least as well and probably better than a pc board, and is certainly much faster. Those willing to make a pc board for the voltmeter are welcome to do so — I'm afraid I'm too lazy!

The rest of the circuit is wired on a perforated breadboard with copper strips on the underside (known in the UK as "Veroboard"). As the photo of the detector board shows, the signal connection from the front panel socket was made using RG-178 coaxial cable. The only place it's important to keep leads as short as possible is in the detector area. Try to mount the two diodes close to each other for good thermal tracking.

alignment and testing

Check the 50-kHz oscillator and divider by placing a crystal earpiece or high impedance audio amplifier between test point C and ground. The frequency can now be adjusted to coincide with the bandpass filter.

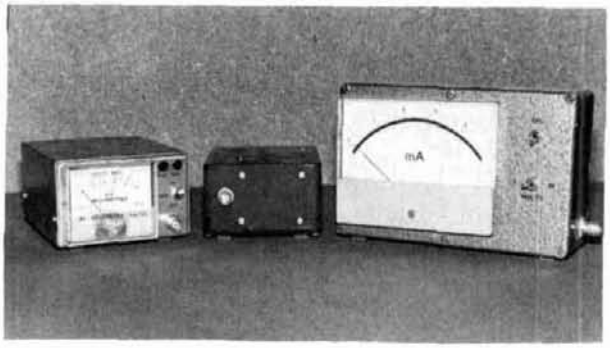


fig. 7. Three different packaging approaches for the same voltmeter.

Turn R9 or R14 so that U4 output is slightly negative, which will cause the integrator U5B to slew to the positive limit. This should result in a healthy square wave output from U3 (pins 1 and 2). A high impedance meter should read a voltage (dc) here that is half that at the U5C output. If the meter is transferred to the anode of CR2, it should be possible to peak this voltage by adjusting the oscillator frequency control R4.

The detector circuit can now be set up with no input. Ground test point A to stop the 50 kHz oscillator. Connect CR1 and CR2 so that the subtractor sees the same voltage at both its inputs. Set test point B to zero volts with R14. Remove the connection between the diodes, and again zero-test point B, this time with R9.

If rf is now applied, U5C should go positive, and the voltage at the output socket SK2 should be 0.45 of this. The meter is now ready to use.

performance

The absolute accuracy and linearity of this meter is illustrated in **fig. 4**, which was constructed from measurements made at 10 MHz. The flatness with frequency was measured at 1 mW, 224 mV rms (0 dBm), and the results are shown in **fig. 9**, which represents a respectable performance of within ± 3 percent up to 150 MHz. This could no doubt be improved to extend the useful range to 70 cm and beyond.

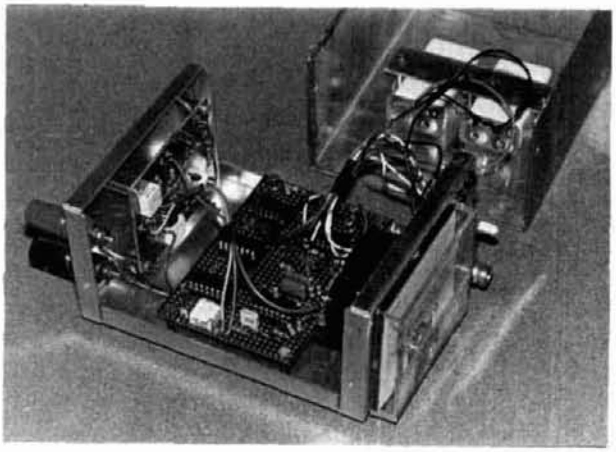
To verify the repeatability of these measurements, I tested the three units against each other using the same source, a 10-MHz crystal oscillator. Referring to the units by position in the photo, the results were:

Unit	Reading on external voltmeter
Left	268 mV
Center	244 mV
Right	258 mV

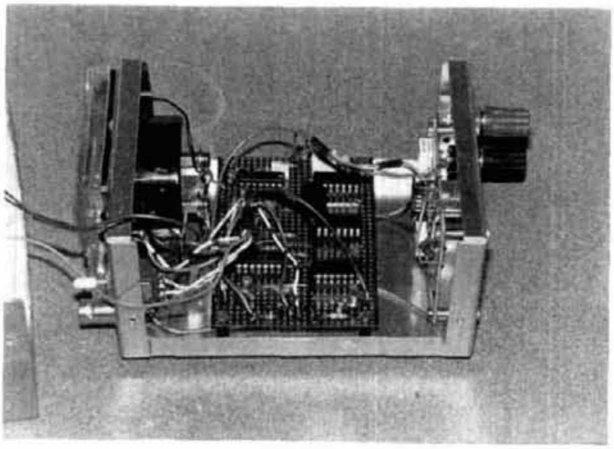
Obviously, three is only a small sample, but considering that the voltmeters had received only the simple dc setup procedure described earlier, I was quite pleased with the outcome, and I hope that this sort of performance will be adequate for your applications.

further suggestions

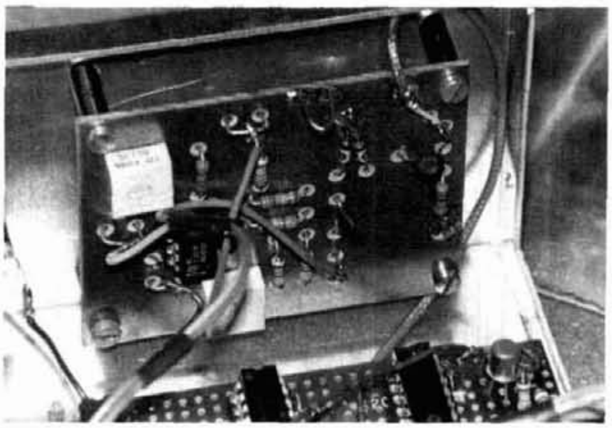
I hope that readers who build this voltmeter will find it a handy instrument to have around the shack. Those who like to experiment and develop their own hardware might enjoy exploring the following options:



A



B



C

fig. 8. Internal views of the rf voltmeter: (A) internal power provided by two 9-volt batteries; (B) "clean" construction enhanced by use of Veroboard; (C) "rf" section of voltmeter.

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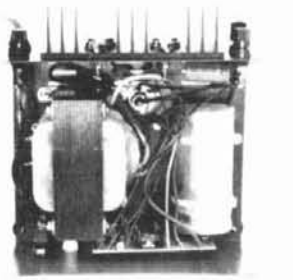

INSIDE VIEW — RS-12A

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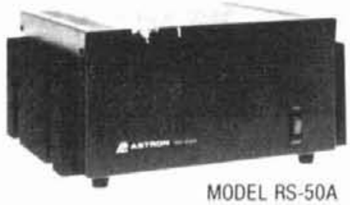

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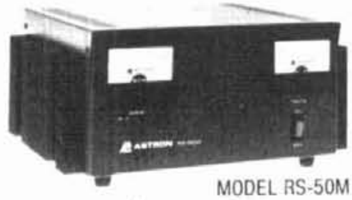
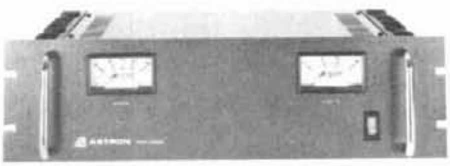


MODEL RS-50M

MODEL VS-50M

RM SERIES



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RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50

RS-A SERIES

MODEL RS-7A

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A	2.5	3	3 × 4¾ × 5¾	4
RS-4A	3	4	3¾ × 6½ × 9	5
RS-5A	4	5	3½ × 6⅛ × 7¼	7
RS-7A	5	7	3¾ × 6½ × 9	9
RS-7B	5	7	4 × 7½ × 10¾	10
RS-10A	7.5	10	4 × 7½ × 10¾	11
RS-12A	9	12	4½ × 8 × 9	13
RS-12B	9	12	4 × 7½ × 10¾	13
RS-20A	16	20	5 × 9 × 10½	18
RS-35A	25	35	5 × 11 × 11	27
RS-50A	37	50	6 × 13¾ × 11	46

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46

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MODEL VS-35M

- Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES

MODEL RS-12S

- Built in speaker

MODEL	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	5	7	4 × 7½ × 10¾	10
RS-10S	7.5	10	4 × 7½ × 10¾	12
RS-12S	9	12	4½ × 8 × 9	13
RS-20S	16	20	5 × 9 × 10½	18

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

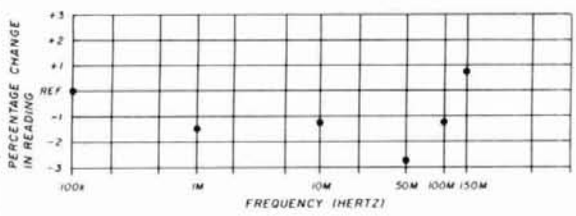


fig. 9. Flatness with frequency at 0 dBm.

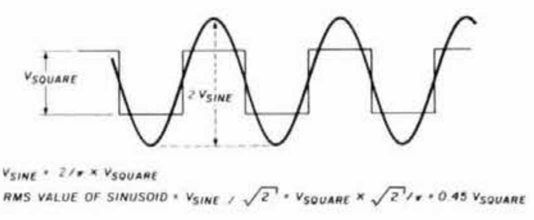
• The design can be simplified by omitting the divider U2. The oscillator could then be run at 25 kHz, or the filter redesigned for 50 kHz. (The choice of frequency was somewhat arbitrary, being high enough to use small coupling capacitor C3 and low enough for active filtering. U2 does guarantee an excellent square wave, but the oscillator alone may well be adequate.)

• The detectors can be built into a high-impedance probe for circuit tracing, rather than a 50-ohm instrument. Keep CR1 and CR2 physically and electrically close together, though.

• By paying attention to the detector matching and circuit offsets, particularly around U4, the useful range could be extended downwards. With attenuators, the range could be extended upwards.

• Careful selection of devices and construction could greatly extend the frequency range.

• The filtering of the square waves could be improved. The units I have built tend to read slightly high, and this could be because the active filter output is not a pure sinusoid, giving a slightly wrong scaling factor. Why didn't I just feed CR2 with a raw square wave? Well, when I tested a diode detector using an accurate function generator, the peak readings were different between sine and square waves — i.e. the diode appeared to clamp at slightly different voltages, depending on the waveform. I wish I knew why; in any case, the results might be worth repeating. If CR2 gave the same response to square waves, the active filter could be omitted. U5C output would then be the peak input voltage, and scaling by 0.707 would give readings in volts rms.



$V_{SINE} = 2/\pi \times V_{SQUARE}$

RMS VALUE OF SINUSOID $= V_{SINE} / \sqrt{2} = V_{SQUARE} \times \sqrt{2}/\pi = 0.45\ V_{SQUARE}$

fig. A1. Relationship between a square wave and its fundamental sinusoidal component.

appendix

how readings are scaled to volts rms

The voltmeter works by making an internally generated sine wave derived from filtering a square wave equal to the rf input. The square wave, then, is generated by chopping a dc voltage. As illustrated in **fig. A1**, Fourier theory tells us that the fundamental (sinusoidal) component of a square wave has a larger peak amplitude than the square wave itself (don't worry, there is less power in this sine wave). If we call the peak amplitude of the sine wave V_{sine} and the peak-to-peak amplitude of the square wave V, then;

$$V_{sine} = \frac{2V}{\pi}$$

But in the voltmeter circuit, the peak-to-peak square wave amplitude is equal to the integrator's dc output voltage V_{dc}, that is:

$$V = V_{dc}$$

We want to make the voltmeter read rms volts. If the applied rf has an rms voltage V_{in}, then the feedback loop makes:

$$V_{sine} = \sqrt{2}\ V_{in}$$

So, the quantity we want to measure, V_{in} is given by:

$$V_{in} = \frac{V_{sine}}{\sqrt{2}} = \frac{\sqrt{2}\ V_{dc}}{\pi} = 0.45\ V_{dc}$$

This is why the dc produced by the integrator is scaled by 0.45.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

low-noise receiver update: part 1

"You can't work 'em if you can't hear 'em" *is an old adage that's still* very true. Building bigger and better *antennas helps, but sooner or later the* antenna size limitation places the burden on the receiver.

Only a decade ago, most Amateurs were using bipolar transistor preamplifiers on the front ends of their VHF and lower UHF band receivers. On the upper UHF and lower SHF bands, diode mixers without preamplifiers were common — with typical noise figures of 6 to 10 dB! That's all changed now, first with the arrival of low-noise silicon bipolar transistors capable of operation into the GHz region, and then with the introduction of GaAs (Gallium Arsenide) FETs in the late 1970s.[1]

Since reference 1 was written, there have been *many new and startling* developments in the area of low-noise devices and techniques. Noise figures are still dropping; device prices have stabilized. So this seems like an appropriate time to update the earlier material and to present state-of-the-art (SOA) information.

This month's column will serve as a quick review and update of the present SOA in low-noise receiver technology. Next month's column will be devoted more to low-noise circuit techniques, recommended devices, testing, and optimization. With all this information in place, you should be right on the *cutting edge of low-noise receiver* technology.

a quick review

The SOA in VHF/UHF and microwave low noise figure Amateur receivers and preamplifiers is now dominated by GaAsFETs, which are technically classified as Metal Semiconductor FETs (MESFETs). The term MESFET is used in the professional community because the gates of a GaAsFET are formed using aluminum, which is a metal that is in direct contact with the semiconductor material. Thus a Schottky barrier diode is formed in the N-type material as shown in **fig. 1A**.

When the lowest noise figure is required above about 100 MHz, GaAsFETs are favored over silicon bipolar transistors because they have up to five times faster electron mobility. Hence, GaAsFETs have much higher cutoff frequencies and gain than silicon bipolar transistors. Furthermore, they typically have much lower noise figures.

Reference 1, *an introduction to low-*noise GaAsFET technology, gave details on preamplifier designs for 144, 220, and 432 MHz, with suggestions for higher-frequency operation. When this material was published in 1984, the GaAsFET was king, but that's no longer true; lower-noise devices and *new breakthroughs in technology* now threaten to decrease noise figures so *far that they will no longer be the* primary limitation to communication capability. Stay tuned.

In addition, during the last few years there's also been a proliferation of "mast-mounted" low-noise preamplifiers using GaAsFETs. These preamps almost completely eliminate the losses associated with feed lines, and virtually eliminate the mismatch loss associated with feed line losses.[2] This is a major problem with low-noise preamplifiers because they often have high input VSWR.

latest developments in devices

GaAsFETs were originally used in commercial and government low-noise amplifiers operating above 2 GHz. Amateurs were in the forefront of developing low-cost GaAsFET preamplifiers to frequencies as low as 30 MHz, but these devices were practical mainly on 2 meters and above, where ambient and sky noise are low.[3]

*GaAsFETs are now being used com*mercially through 40 GHz and possibly

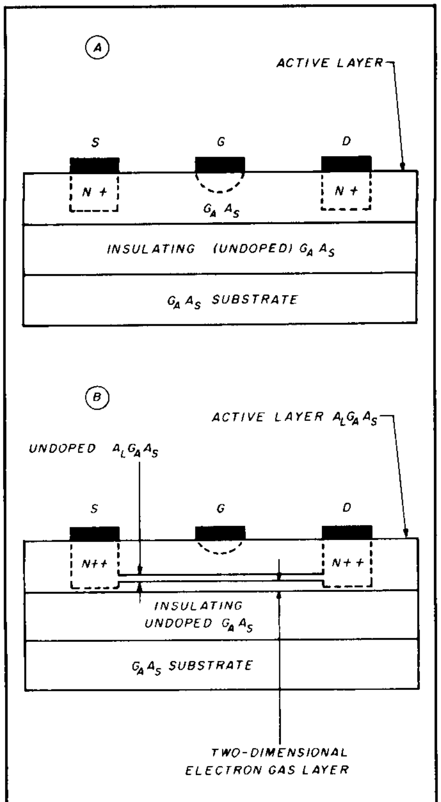


fig. 1. Typical physical structures (not shown to scale) of GaAsFETs and HEMTs: (A) depletion mode GaAs FET; (B) depletion mode HEMT.

higher. New lower noise figure, higher gain, and cutoff frequency devices seem to be appearing almost monthly. Needless to say, if you want to be on the cutting edge of technology, you might as well use any premium-quality devices you have in your desk drawer as soon as possible — before they become obsolete! GaAsFETs with noise figures less than 1.0 dB are now available through 4.0 GHz! SOA GaAsFET noise figures versus frequency are shown in **fig. 2**.

Probably the most important recent improvement in the SOA in low-noise devices is the development of the HEMT (high-electron-mobility transistor).[4,5] Sometimes referred to as TEGFETs (two-dimensional electron GasFETs)[6] or heterojunction FETs (to avoid infringing the copyright on the name HEMT in Japan). Technically speaking, the HEMT is a heterojunction superlattice device that was first described in 1978 and demonstrated by Fujitsu and Thompson-CSF in 1979.[4,5,6] It is very similar in structure to the GaAsFET except for the two-dimensional electron gas as shown in **fig. 1B**.

The HEMT's major feature is, typically, its higher transconductance with a cutoff frequency twice that of a comparable GaAsFET, with higher gain and noise figures as low as half those of typical GaAsFETs! Cutoff frequencies well above 100 GHz have been reported. HEMTs with less than 1.0-dB noise figures are now available through X band (12 GHz). The SOA in HEMT noise figures is shown in **fig. 2**.

Right now, however, most HEMTs are laboratory devices, and the lowest noise devices are very scarce. Only a few HEMT types are available commercially, and these devices are expensive — typically more than $150 each! However, remember that GaAsFETs were in the same price range in the mid-1970s, and better devices are now available for less than $5! HEMTs are known to be manufactured by Fujitsu, GE, Gould-Drexel, NEC, Sony, Thompson-CSF, Toshiba, TRW, and Varian Associates. Other suppliers and even lower noise figures are promised!

Unlike other innovations in technology, the HEMT is compatible with existing GaAsFET dc biasing and rf characteristics. HEMTs usually use the same packages and can be virtual "drop-ins" for GaAsFET circuits. The primary difference is that the HEMT's optimum source impedance is generally higher than an equivalent GaAsFET's. Therefore, an adjustable input-matching circuit similar to the one described in reference 1 is recommended so that the optimum source impedance can be achieved.

One major MESFET anomaly should be stressed. As pointed out in reference 1, GaAsFETs (as well as HEMTs) have a very high noise figure in the so-called $1/f$ or low-frequency region. This

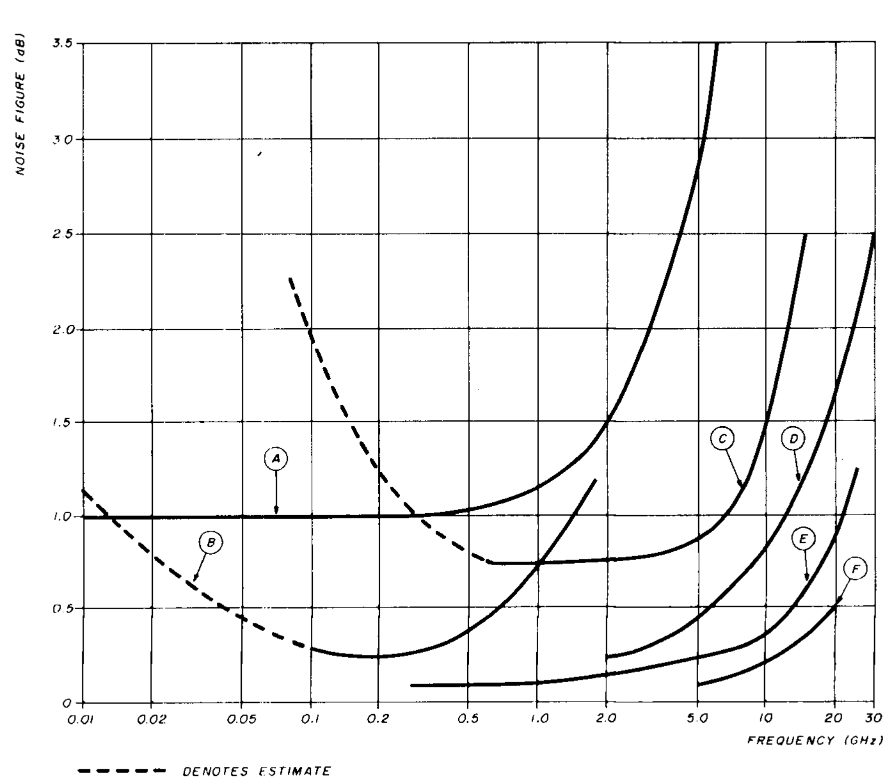


fig. 2. Typical 1987 state-of-the-art noise figures of bipolar transistors, GaAsFETs, and HEMTs: (A) uncooled bipolar transistor; (B) low-frequency GaAsFET at room temperature; (C) high-frequency GaAsFET at room temperature; (D) HEMT at room temperature; (E) GaAsFET cooled to 12 degrees Kelvin; (F) HEMT cooled to 12 degrees Kelvin.

means that the noise figure increases not only as you increase frequency, but also as frequency is decreased! This effect is shown in **fig. 2**.

The amount of noise figure increase and the frequency where it begins to increase (below the normal operating frequency) depends on the device type. Generally speaking, the ideal rf operating region for GaAsFETs and HEMTs is over a one decade-wide frequency range referenced down from the specified operating frequency (not the F_{max}).

For example, a device specified for 1-GHz operation at the top of its operational frequency range will probably be well suited for operation down to about 100 MHz. However, a device specified for 10 GHz will probably have a higher noise figure if it's used much lower than about 1 GHz!

Therefore, don't expect that a very low-noise GaAsFET specified for 10 GHz will be a super low-noise device at 144 MHz. A low-noise 10-GHz HEMT may well have a higher noise figure at 432 MHz than a much less expensive device specified for operation through 4 GHz. This is why so many Amateurs have been able to demonstrate incredibly low noise figures on 2 meters using GaAsFETs costing no more then $5 to $10!

Also, the higher the cutoff frequency of a GaAsFET or HEMT, the narrower the gate; hence, the susceptibility to static burnout increases. Furthermore, higher frequency devices are more prone to oscillate when operated at lower frequencies. So don't "read into the specifications" anything that isn't there. For optimum performance versus cost, operate MESFETs in the frequency range recommended by the supplier.

noise figure limitations

I'm often asked the question, "What limits noise figure?" It should be intuitive that part of the limitation on noise figure is in the actual device itself. Furthermore, for the lowest possible noise figure in a receiver, the gain of the first stage must be high and the second stage should also have a low noise figure. This is shown mathematically by the following equation.[7]

$$F = F_1 + (F_2 - 1/G_1) + (F_3 - 1) / (G_1 \cdot G_2) + \cdots \quad (1)$$

where F is the overall noise factor of the receiver, F_1 is the noise factor of the first stage, F_2 is the noise factor of the second stage, F_3 is the noise factor of the third stage, G_1 is the numeric gain of the first preamplifier and G_2 is the numeric gain of the second preamplifier. Note that noise factor and gains are in numerics, not decibels, so they often have to first be converted from decibels to numeric values before using them in **eqn. 1**. After the final noise factor is determined, you'll probably want to convert noise factor back to noise figure using the following equation.

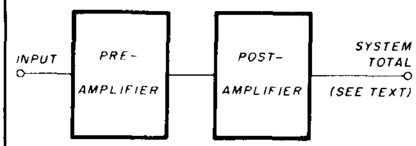
$$NF = 10 \log F \quad (2)$$

For example, refer to **fig. 3**, a block diagram of a typical Amateur front end. In example 1, if the noise figure of the first stage of a receiver is 0.5 dB (noise factor = 1.122), with a gain of 13 dB (gain = 20) and the second stage noise figure is 4.0 dB (noise factor = 2.51), with a gain of 15 dB (gain = 31.6) -- ignoring the third stage contribution and assuming it to be negligible) — the overall receiver noise figure will be 0.78 dB (noise factor = 1.197), a significant 0.28-dB increase over the first stage alone.

Now if we reduce the noise figure of the second stage to 1.75 dB (noise factor = 1.496) [example 2] or increase the gain of the first preamplifier to 8 dB (gain = 63) [example 3], the overall noise figure will be 0.59 dB (noise factor = 1.46), only 0.09 dB above the preamplifier alone, a small penalty to pay.

These calculations are often laborious and prone to error. For this reason, it's best to program **eqn. 1** and **eqn. 2** into a computer or scientific calculator to simplify the calculations and decrease the possibility of human error.[8]

Finally, don't get carried away with gain. Increasing the first stage gain too much may lead to intermodulation distortion or instability, thus limiting the



Example 1:			
Noise figure (noise factor)	0.5 (1.122)	4.0 (2.51)	0.78 (1.197)
Gain dB (gain in numerics)	13 (20)	15 (31.6)	28 (631)
Example 2:			
Noise figure (noise factor)	0.5 (1.122)	1.75 (1.496)	0.59 (1.146)
Gain dB (gain in numerics)	13 (20)	15 (31.6)	28 (631)
Example 3:			
Noise figure (noise factor)	0.5 (1.122)	4.0 (2.51)	0.59 (1.146)
Gain dB (gain in numerics)	18 (63)	15 (31.6)	33 (1995)

fig. 3. Block diagram of a typical Amateur front end showing the effects of tradeoffs in gain and noise figure (see text).

ability to use the inherent low noise figure.[9,10] Therefore, with the low cost of devices today, it's preferable to design for a reasonable first stage gain (15 to 20 dB) and use a similar type second stage with a moderate noise figure (1.0 to 2.0 dB typical). This provides an inexpensive and useful cost/performance tradeoff.

other noise figure limitations

Another noise figure limitation is incurred by operating a preamplifier at room temperature (more on this shortly). However, the major limitations on Amateur receivers attaining very low noise figures commensurate with device specifications are losses associated with the input impedance-matching circuitry.

Amateur preamplifiers are usually designed for a single frequency band. Typically the circuits employ some form of input tuning. This is a preferred technique since the input network will not only allow the device to be optimized for the lowest possible noise figure at the frequency of interest, but will also act like a filter and prevent strong out-of-band signals from entering or causing IMD.

Most Amateur preamplifier input circuits, especially below 500 MHz, use an inductor and capacitor tank circuit similar to those shown in **figs. 4A** and **4B**. **Figure 4A** has one less component, but it also requires the tap to be

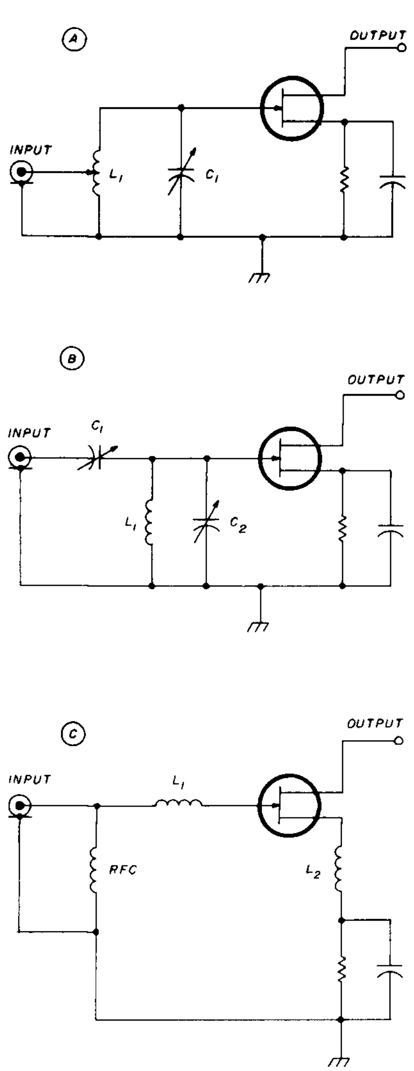


fig. 4. Some typical recommended input matching circuits for Amateur low-noise GaAsFET preamplifiers operating below 2.5 GHz: (A) tapped inductor; (B) capacitor transformer; (C) lossless feedback.

carefully chosen. This can be very tedious and time-consuming, especially if you want to achieve minimum noise figure. Therefore, the input impedance-matching circuit shown in **fig. 4B** is recommended for 500 MHz and below.[1]

Sometimes I see Amateurs and commercial designers alike using an abbreviated type of input matching similar to that shown in **fig. 4B**, but with the shunt capacitor, C2, removed. This is not recommended because if the lowest possible noise figure is wanted, the inductor also has to be tuned, and that can be a tricky job. (And what do you do for tuning if the GaAsFET has to be replaced?)

By now you've probably surmised that the minimum noise figure isn't only a function of tank circuit alignment, but more likely due to losses in the components themselves. All capacitors and inductors have loss, especially as you go above 100 MHz. The higher the Q of the components in the input-matching network, the lower the insertion loss and hence the lower the noise figure.

Probably the "lossiest" component in a low-noise preamplifier is the inductor. A typical inductor in the 100- to 500-MHz range has an unloaded Q (no external components attached) of 300 to 500, depending on wire type, diameter, form factor, and proximity of other components and shielding structure.[11]

As explained in reference 11, there's a definite insertion loss relationship between the unloaded Q_U of an inductor and the loaded or "in-circuit" Q_L of the same as follows:

$$\textit{insertion loss (dB)} = 10 \log [1 - (Q_L/Q_U)]^2 \quad \textbf{(3)}$$

where Q_L and Q_U are the loaded and unloaded Qs, respectively.

How do you determine the loaded Q of the inductor? If the preamplifier is one of the types that uses a broadband output network as described in reference 1, the half-power or 3-dB bandwidth of the preamplifier can be easily measured. The Q of the preamplifier (and therefore the loaded Q of the inductor) is then determined as follows:

$$Q\ preamp = f_O/(f_H - f_L) \quad \textbf{(4)}$$

where f_O is the center frequency in MHz, f_H is the upper half-power frequency and f_L is the lower half-power frequency. For example, if we have a 432-MHz preamplifier with half-power frequencies of 440 and 423 MHz respectively, the loaded Q will be 432/(440 − 423) or 25.4.

Now, if we assume that all other components contribute negligible loss, we can determine the approximate input circuit losses attributable to the inductor's Q. Using **eqn. 3** and assuming a good inductor with an unloaded Q of 500 and a preamplifier with a loaded Q of 25.4, we have an input circuit insertion loss of approximately 0.45 dB.

Typical GaAsFET preamplifiers using input tank circuits of this type have noise figures of 0.5 to 0.75 dB. Therefore, with a 0.45-dB input loss, the overall noise figure of the preamplifier is almost entirely due to the losses in the input network and the GaAsFET itself must be virtually noiseless!

To show the Q_U/Q_L losses more graphically, I've prepared the graph in **fig. 5** and scaled it for low loss and hence low noise figure conditions. (Check **fig. 5** for the 432-MHz preamplifier case above.) For a Q_U/Q_L ratio of 500/25.4, or approximately 20, you'll see that the insertion loss is indeed 0.45 dB.

Also note in **fig. 5** that to get the input losses down below 0.1 dB, the unloaded-to-loaded Q ratio must be equal to or greater than 90. This means that the unloaded Q of the inductor in the preamplifier just described would have to be over 2000! If you want a very low-noise preamplifier, you're going to have to use some pretty low-loss inductors — such as a large (1- to 3-inch diameter) coaxial cavity resonator — and possibly have them silver plated.

Figure 4C is a different input circuit topology which eliminates the tank circuit per se by using a series input inductor and "lossless feedback" in the source lead.[1] This type of circuit definitely has lower input losses and potentially a better input VSWR. However, it's more prone to out-of-band interference and therefore is more appropriate for use on the microwave bands. It will be discussed further in next month's column.

other component losses

Don't forget that there can be other losses besides the input inductor. Tuning capacitors can also have losses. Only the lowest loss, highest Q tuning capacitors should be used in the

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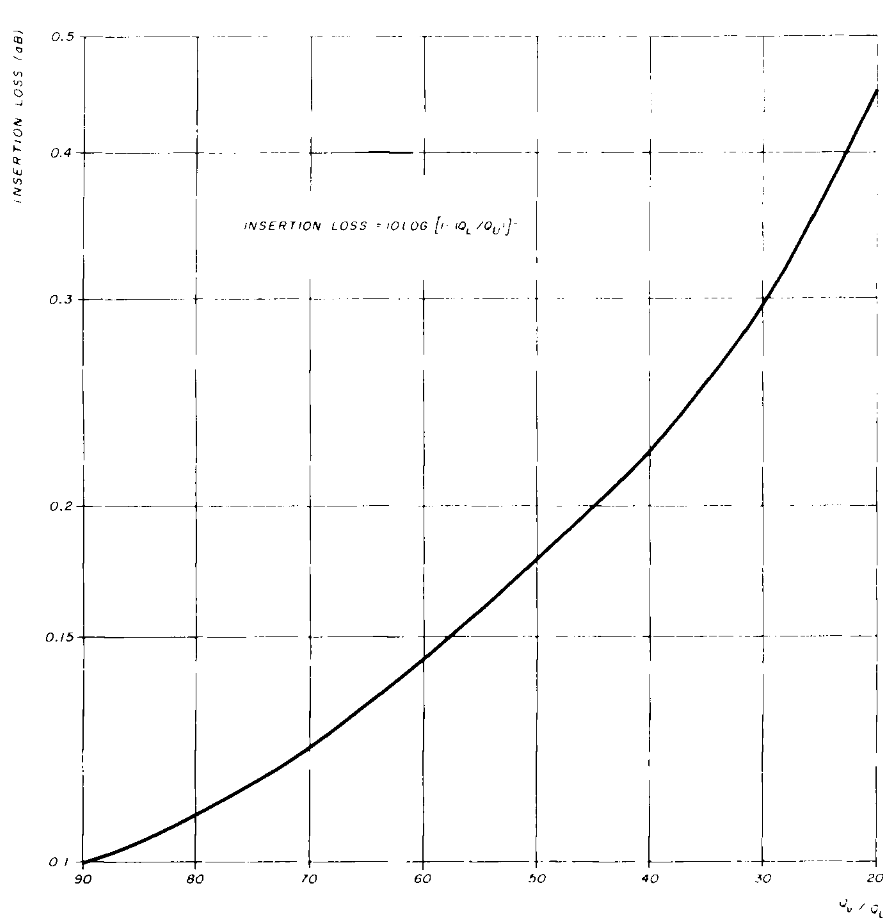

fig. 5. Insertion loss of an inductor (or capacitor) due to the effects of unloaded and loaded Q.

input-matching network. The air-variable type capacitors manufactured by Johanson and others are a preferred type. They not only have low loss and high Q, but also good tuning resolution with little or no backlash. Furthermore, they often have special sealing caps that can be placed over the tuning mechanism to help keep out moisture and prevent inadvertent mistuning.

The minimum Q of a Johanson-type 5200 air variable, one commonly used by Amateurs, is 5000 at maximum capacitance at 100 MHz. This figure decreases rapidly to less than 1000 above 300 MHz! Higher Q types such as the Johanson 5700 and 5800 are recommended, but they have lower maximum capacitance so they're useful only for higher frequencies and for series connections where lower capacitance values are required.

Chip capacitor losses can also be considerable, especially when used in source bypassing or in the rf path. In critical low-loss circuits, the porcelain types are highly recommended despite their higher initial cost. Be careful, too, of resistor types. The older 1/8- or 1/4-watt carbon composition types are recommended. However, the non-carbon or film types that are becoming so popular are usually quite reactive and lossy, and are therefore not recommended.

Finally, coaxial connectors — especially type N, TNC, and SMA — are highly recommended for low-noise preamplifiers because they have low loss and a very positive mating mechanism. On the other hand, BNC- and UHF-type connectors should be avoided because their impedance isn't constant, and they have questionable mating tolerances and known insertion losses. Connector types and losses are discussed further in references 12 and 13.

lower noise techniques

Cooling is probably the last resort when it comes to really low-noise preamplifiers. Bipolar transistors generally don't work well below about 70 to 80 degrees Kelvin. However, many GaAsFETs and HEMTs seem to do quite well when cooled even as low as 12 degrees Kelvin, the temperature of liquified Helium.

The National Radio Astronomy Observatory (NRAO), in Charlottesville, Virginia, has been building low-noise preamplifiers for many years. Their preamplifiers are used in radio telescopes where the sky temperature is as low as 3.5 degrees Kelvin, almost absolute zero. By 1980 they were using GaAsFET preamplifiers cooled to about 13 degrees Kelvin in a Dewar with liquified Helium.[14]

At first NRAO used GaAsFETs because they noticed that the transconductance would often increase — sometimes by as much as 50 percent — as temperature was decreased. At the same time, the noise figure would drop. However, the optimum source impedance changes at low temperatures and oscillations may occur. Consequently the preamplifier has to be optimized at the cold temperature. Recently, NRAO noticed the same effects with HEMTs.

Because the cryogenic coolers used by NRAO cost about $5000 each, they're not really practical for Amateurs. Other less expensive coolers such as the thermo-electric type are available commercially.[15] However, they use diodes that may generate noise, so be cautious if you use them. It should be sufficient to mention that if you have an antenna-mounted preamplifier, especially for EME, you should mount it so that it won't be heated excessively by the sun.

Finally, of the GaAsFETs tested by NRAO, the MGF 1412 seems to have consistently low noise figure at room temperature. Futhermore, at cryogenic temperatures, the MGF 1412 type seems to be one of the most reliable for low noise figures. Since this is one of the most popular types used by Amateurs seeking the lowest possible noise figures, it may be a place to start.

summary

In this month's column, I've attempted to bring you up to date on the SOA in low-noise receivers for VHF and above. Noise figures are still dropping, but at some frequencies can't go lower unless we change the circuit techniques we're presently using. In next month's column, we'll discuss some circuit and device recommendations.

acknowledgments

I'd particularly like to thank Bill Lakatosh, AA4TJ (ex K3QCQ and KJ4QI), of NRAO for their input on SOA noise figures and cooling techniques.

new records

In last month's column we mentioned the outstanding sporadic E occurrence during the ARRL June VHF QSO party and asked for any new record claims. Shortly after the contest was over, I received and authenticated a new North American 2-meter, double-hop sporadic E record. The new record holders are Jim Poore, KD4WF, in Savannah, Georgia (EN92LK) and Jim Frye, NW7O/7, operating portable from Mount Potosi, southern Nevada (DM25GV). Their contact took place on June 14, 1987, at 1704 UTC and extended the existing record by almost 90 miles for a new record of 1980 miles (3186 km). Congratulations to both Jims.

The North American 10-GHz DX record has also been broken; more on that in next month's column.

important VHF/UHF events

November 3	Predicted peak of the Taurids meteor shower at 2200 UTC
November 3	Predicted peak of the Cassiopids meteor shower at 2200 UTC
November 14-15	ARRL EME Contest (second weekend)
November 17	Predicted peak of the Leonids meteor shower at 1500 UTC
November 21	New moon
November 24	EME perigee
December 13	Predicted peak of the Geminids meteor shower at 1900 UTC
December 20	New moon
December 21	± month. Winter peak of sporadic E propagation
December 22	Predicted peak of the Ursids meteor shower at 2200 UTC
December 22	EME perigee

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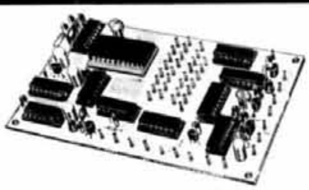

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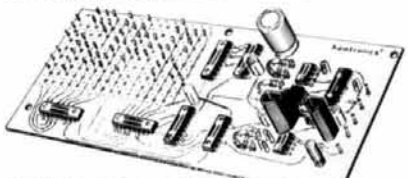

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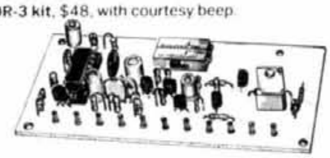

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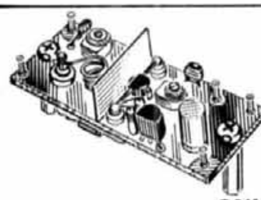

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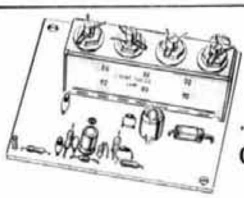

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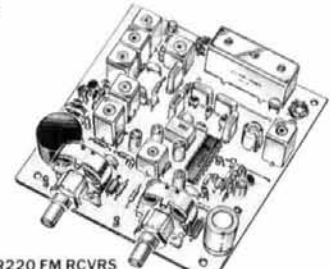

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Here's a collection of 30 super programs written by ON4UN. Just about every interest or need is covered—from antenna design and optimization to general operating programs. Antenna programs include: shunt and series input L network design, feedline transformer, shunt network design, SWR calculation, plus 11 more! General Ham programs include: sunrise/sunset, great circle distances, grayline, vertical antenna design program, sunrise calendar plus 9 more! Phew. When you sit down to use these programs you'll be amazed at what you have. The best value in computer software available today. © 1986

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☐ UN-CPM/Kaypro	$19.95
☐ UN-C-128 (COMMODORE)	$19.95
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by John Devoldere ON4UN

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by Victor Misek, W1WCR

Recognized around the world as the definitive work on Beverage Antennas. Misek delves deep into the secrets of the single wire Beverage and SWA (Steerable Wave Antenna) with helpful hints and tips on how to maximize performance based upon wire size, height above ground, overall length and impedance matching. Also includes information on center fed Beverages constructed out of several wire types. ***CITY LOT OWNERS*** Note: Misek has developed a Beverage for you too! Called the Micro-SWA, it is just 60 ft. long. You get excellent directivity and null steering capabilities. Transformer design information for both termination and feedline matching is completely revised. © 1987 80 pages 2nd Edition

☐ VM-BAH — Softbound $14.95

HAMLOG (MS-DOS Computer)
by Rick Martin, WA3YLD

Harness the power of your MS-DOS computer with this tested and proven program. HAMLOG is fully menu driven and features fast record retrieval and display. Can be edited at any time and allows hard copy print-outs by callsign, date, or a number of other parameters. Also includes helpful frequency allocation tables by license class and third party agreements for traffic handlers.

RM-HL (MS-DOS) — $49.95

DIGIPAC II (MS-DOS Computers)
NEW!

by Kalt Assoc.

Digipac II combines a full featured computer communications program with a powerful message editing and formatting program designed for traffic handling. The message program features address presets, complete subscrolling form, prompts and pop-up help and selection menus, full user defined help system plus more. The communications program has plenty of features too: multi-level alarm, multi screen (full, split or recall,) user programmable function keys to name just a few. No matter what your interest in packet — from rag chewing to traffic handling — this program is for you.

KA-DPII (MS-DOS Computers) — $49.95

QSO-PRO (MS-DOS)
NEW!

by Mark and Mary Morland

Not just another logging program. Written with the active Ham in mind, this MS-DOS program has been professionally designed to unleash the potential of the MS-DOS computer. Uses a logical standardized data entry format and allows editing, sort and selection of data. Powerful program at a very reasonable price.

MM-QP (MS-DOS Computers) Reg. 39.95 Special $34.95 (special expires December 1, 1987)

ULTRA-HAM Contest Logger
NEW!

(C-64 COMPUTER)

ULTRA-HAM is a powerful program. It includes: local and foreign contest logs, a disk formatting routine and a printing program. Easy-to-use and full of helpful prompts, Hams from Extras to Novices will find this to be a great addition to their software library. Other features include: automatic numbering, super fast dupe checker, automatic UTC clock, automatic disk saving every eight QSOs, running score totals plus much more. Also includes editing capabilities.

RC-UH (for C-64) — $39.95

RF-CAD ELECTRONICS DESIGN PROGRAM — Version 3.7 Includes Intermodulation and Distortion Program
NEW

by Joe Reisert, W1JR and Gary Field, WA1GRC

For IBM PC and compatable computers

This software package has been written by electronic engineers and contains nearly 40 tested and proven programs that will help the Radio Amateur or engineer design many common types of radio circuitry. Emphasis has been placed upon ease of use. Wherever possible, menus of choices with examples are displayed. Should the user be computer literate, the programs are not copy protected so they can be modified to meet your specific requirements. (full documentation is also provided.) Programs include: Filters, LC, active- LP,HP,BP; Inductor design, torroid, solenoid, straight wire; Matching networks; Crystal oscillators; Microstrip; Transmission lines; Antennas, Yagi-Uda, helix, dish, horn, element scaling; Pi and T attenuators. Also included: Radio Path calculations; FM modulation analysis, Miscellaneous conversions; Geostationary satellite pointing; Moon tracking aids; Receiver noise figure calculations and Spurious receiver response prediction. Requires IBM-PC with at least one floppy drive and 128k of RAM. © 1985

☐ RF-CAD (MS-DOS) — $39.95

SHOPPING IDEAS

RADIO HANDBOOK 23rd Edition
by Bill Orr W6SAI

Here are some of the highlights of this exciting new edition: New easy-to-use charts for Chebyshev and elliptic filter configurations, new data on power MOS-FETS, how to use state-of-the-art OP-AMPS, and home computer RTTY to name just a few examples. New projects include: GaAsFET preamps for 902 and 1296 MHz, easy-to-build audio CW filter, Economy two 3-500Z, 160 meter amplifier, multiband amp using two 3CX800A7's, and a deluxe amplifier with the 3CX1200A7 tube. New antenna projects include: efficient Marconi design for 160 and 80 meters, computer generated dimensions for HF-Yagis, and a 2 meter slot beam. Get your copy today. 23 edition © 1986

☐22424 **(Reg. $29.95)** **Hardbound $26.95**

Re-CODE-nition CLASS (Apple II)

NEW!

Sound recognition has been proven time after time to be the most efficient method for learning the Morse code. re-CODE-nition uses a computerized derivation of the famous "Farnsworth" Hi speed — slow spacing code teaching method. Using a word game approach to learning, this program strives to eliminate the drudgery of learning the code. You can custom design code practice by entering text or use data files on disk.

MD-CC (Apple II, IIe/c) **$39.95**

HAM RADIO LICENSE STUDY COURSE

NEW!

by Diamond System (MS-DOS Computers)

Use your home computer to study for your Amateur Radio License. Questions and answers are taken from the VEC standard syllabus *for each license class. You can either* study the whole question pool or one of the individual sub-elements. Novice, General and Extra programs also include a Morse code program that will allow you to bone up on your code. The program can either generate code randomly or from a text file. State-of-the-art learning. MS-DOS only.

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DS-A	**Advanced**	**$49.95**
DS-E	**Extra (with code study)**	**$49.95**

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☐**HR-LB** **Spiralbound $2.95**

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RADIO COMMUNICATION HANDBOOK
by RSGB 5th Edition

Here's an inside look at Amateur Radio with the British version of the ARRL Handbook. Full of projects, theory and *other helpful technical information.* Also includes antennas, propagation, satellites plus much more! © 1982 23 chapters 1.25" thick!

RS-RCH **Softbound $21.95**

COMPUTER PROGRAMS FOR THE RADIO AMATEUR

Book currently unavailable.

by Wayne Overbeck, N6NB, and Jim Steffen, KC6A

Here are the best computer programs for the Radio Amateur available. Gives you programs that will help you to, determine sunrise/sunset times, track the Moon's path across the sky, use Greyline propagation and set up record systems for WAS, DXCC and VUCC, or any other award. Take full advantage of your computer with well written and tested programs. © 1984

Program Listing

Data Base Mgmt.
Logs, Awards Data Base, Gridlocator

Latitude/Longitude Programs
Data File, Beamheadings, DX Display, Sunrise Chart, Greyline, DX Checker

Contest and Duping
Dupechecker, General Contest Logger, Field Day Logger, Sweepstakes Logger, Log Print

Antenna Programs
Antenna Scaler, Matching Evaluator, Vertical Pattern Plotter

EME
EME System Calculator, Sky Locator, Moontracker

☐**Program disk** **$19.95**
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Disks available for: Apple II (DOS 3.3), IBM PC-DOS, TRS-80 Model III and Commodore C-64. Please mark your order with the program disk you want.

TUNE IN THE WORLD WITH HAM RADIO
by ARRL staff **7th EDITION**

NOW INCLUDES TWO C-90 CODE STUDY TAPES!

This package has been revised to cover new digital & voice Novice requirements and contains **THE** goodies needed by the beginner to get started in Amateur Radio. Assuming that you have no prior knowledge of radio, the reader is *taught how* to pass the Novice exam, both code and theory, and how to set up a station. Unique code study method makes learning the Morse code easy as 1-2-3. And it's full of illustrations to help clarify difficult technical points. 160 pages © 1987 7th edition.

☐**AR-HR** **Softbound $14.95**

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☐**26-01 Advanced Class**	**$4.95**
☐**17-01 Extra Class**	**$4.95**
☐**102-01 Radio Amateur Theory Course**	**$6.95**

HAM DATA

By Chip Lohman NN4U **FOR C-64**

AMATEUR RADIO COMPUTER SOFTWARE

MASTER LOG

Over three years of development went into this program. It creates a file of 2100 individual records with up to 13 different entries per record. Master Log can do a search and select based upon time, frequency, mode or any of the other variable parameters. It keeps track of DXCC and WAS status, prints *QSL labels and can search its whole file in less than 5 se-*conds! Complete documentation isincluded to help you learn and use this truely state-of-the-art logging program.

☐**HD-ML** (For C-64) **$28.95**

DX TOOL KIT

This handy tool kit should be in every ham's shack. Programs included are: Propagation chart, Beam heading, Grayline and Sunrise. The Propagation chart uses the NOSC Minimuf 3.5 algorithm to compute both MUF and FOT for any QTH. The Beam heading allows you to create a customized chart of both long and short path beam headings. Grayline selects and sorts countries on the Grayline for your location. The Sunrise routine will help you predict upcoming DX openings.

☐**HD-DX (For C-64)** **$24.95**

ANTENNA DESIGN

This self-prompting program covers dipoles, Yagis, long wires, rhombics, quads, loops, verticals and other antennas. You simply enter the frequency you are designing an antenna for, the antenna you want to make, and the computer will give you all the initial design specifications you need. The results *can be stored on disk for permanent retention.*

☐**HD-AD** (For C-64) **$9.95**

PROPAGATION CHART

Invaluable addition to any Hamshack. The Propagation Chart allows you to determine the maximum usable frequency (MUF) and the optimum frequency (FOT) between your QTH and any spot on the globe. You can get either a screen display, a printed tabular list over several days, or a printed graph of MUF and FOT. You also get automatic beam headings and distance to the DX station. Covers 160-10 meters and the new WARC bands.

☐**HD-PC** (For C-64) **$16.95**

SUPER LOG

Super log gives you all the advantages of a computerized data base without significantly changing the traditional log format. For contesters, Super Log can be configured to either manually or automatically enter contact number as well as time of contact. Make an error and you can easily go back and edit the entry. Super Log also allows you to print out either selected contents or the whole log. Will print QSLs.

☐***HD-SL*** *(For C-64)* **$19.95**

CONTEST LOG

This disk contains four different contest programs; ARRL Sweepstakes, Field Day, Universal WW Contest log, plus a dupe checking routine. Each program is designed for real time use. It automatically enters date, time, band andserial number for each contact. A 24-hour clock is displayed at the top of the VDT screen. When the contest is over, the program will print your results listing all duped and scored contacts in serial sequence with all the necessary information as well as completed score at the bottom of the page.

☐**HD-CL** (For C-64) **$24.95**

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☐**UFBC-81** **$6.00**

I.D. BADGES

No ham should be without an I.D. badge. It's just the thing for club meetings, conventions, and get-togethers, and you have a wide choice of colors. Have your name and call engraved in either standard or script type on one of these plastic laminated I.D. badges. Available in the following color combinations (badge/lettering): white/red, woodgrain/white, blue/white, white/black, yellow/blue, red/white, green/white, metallic gold/black, metallic silver/black.

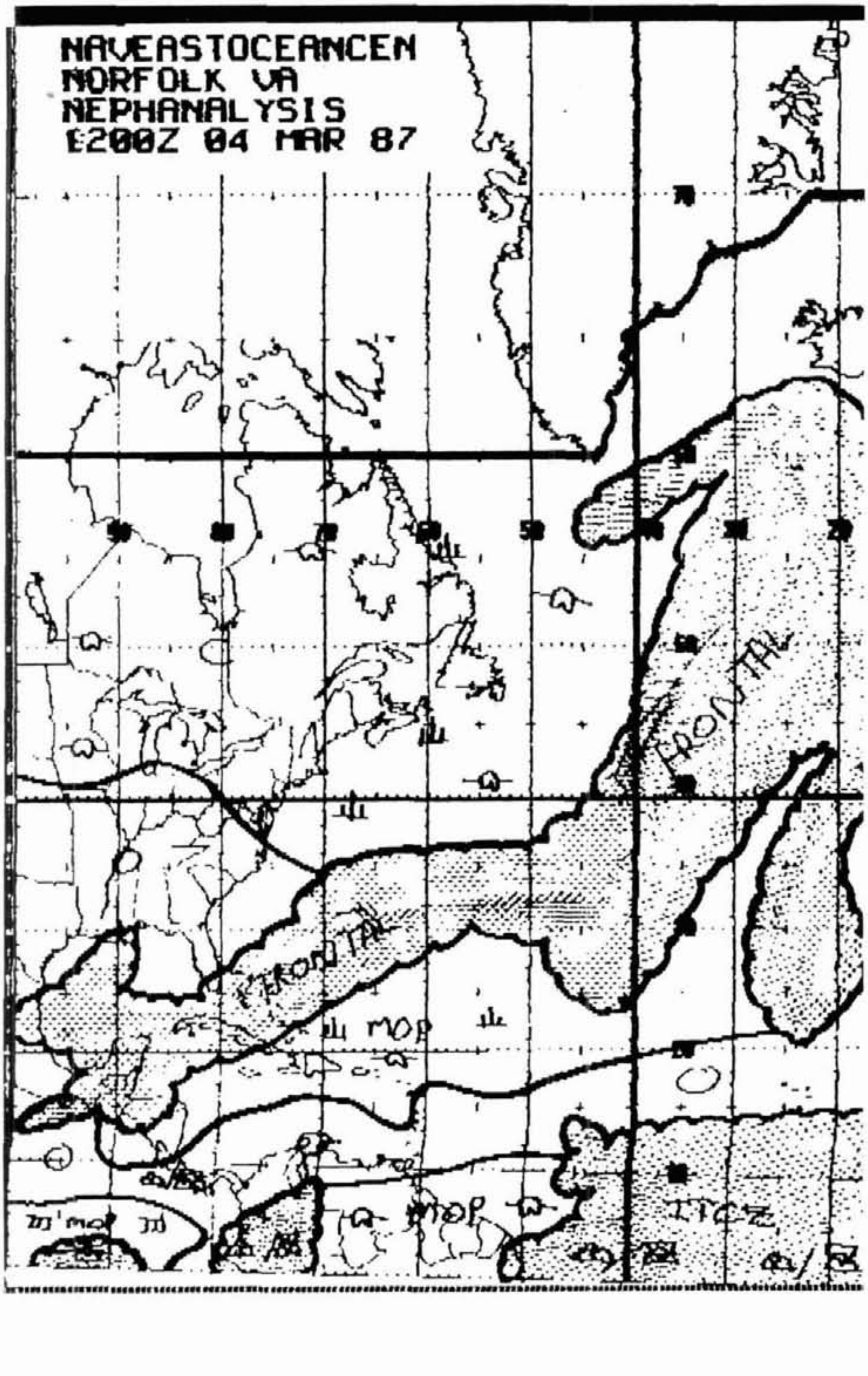
☐**UID** Engraved I.D. Badge **$2.50**

product REVIEW

AEA weather FAX mod for the PK-232

Over the past few years, packet radio has grown from a rather esoteric part of Amateur Radio to one of the fastest growing segments ever. There must be at least ten manufacturers of TNCs, *all selling basically the same product.*

One way of selecting a TNC is to look closely at the features each unit offers. Do you want to go beyond packet? How about RTTY, AMTOR, ASCII, CW and weather FAX (WEFAX)? How about SIAM, Signal Identification and Acquisition Mode? If you want *all* these features in a single unit, your only choice is the AEA PK-232.



All currently manufactured PK-232s have the FAX and SIAM option installed. The unit we reviewed earlier (see page 81 of the July, 1987 issue) was one of the first units off the line and needed to be modified to work on WEFAX and SIAM.

When I learned of AEA's modification for WEFAX, I immediately called and placed an order for the kit. Because demand was enormous, it took a few weeks for my kit to arrive.

Besides the parts and instructions needed to perform the mod, AEA also supplied an addendum for the operator's manual and a pre-made computer-to-TNC to printer cable. The cable alone is more than worth the $40 price of the modification kit.

AEA is currently supporting most parallel graphics, and your dealer will have a complete list of printers that AEA has tested with the PK-232.

Four simple steps were all that were required to make the modification: remove the unit from use, disconnecting all cables; prepare a clean, static-free work area; remove six screws and open the unit; remove and replace EPROM U2 and install EPROM U3, then screw the cover back on. That's all there is to it. You're ready to reconnect and get back on line.

Sometimes it's hard to believe that the EPROM is as powerful as it is. Without EPROMS, the unit can't operate. Yet they can be installed in less than a heartbeat.

operation

I'll break this section into two parts: WEFAX and SIAM.

WEFAX. While WEFAX is basically a service for ships and aircraft, it offers a wealth of information for amateur meteòrologists too. Stations transmit weather maps that show actual conditions and prognostications, satellite photographs of the earth's surface, taken from geosynchronous and orbiting satellites, and plenty of additional information. Of particular interest to me has been the hurricane maps that are sent during the hurricane season here on the East Coast.

Transmitting stations are located around the world, with each transmitting information for its own geographical area. Here in the Northeast, Halifax, Canada and Norfolk, Virginia provide the most reliable reception and information. I haven't had much luck with European or West Coast stations, but this is more a problem of time on the air and propagation.

The PK-232 represents the third generation of WEFAX equipment I've operated. Of the three units I've used, it's by far the easiest to set up and the most convenient to use. No special paper is required. There are no noxious fumes, the unit is easy to transport and install — it's really a pleasure!

From the time the modifications were finished, the cables installed, the computer hooked up and

product REVIEW

booted, to the reception of the first pictures, a grand total of 15 minutes passed.

Because you're using a dot matrix printer, you can't resolve shades of gray and therefore can't reproduce satellite images with pure photographic quality. The images, however, are very good and quite usable. Since maps and charts are black on white, their reproduction quality is excellent.

All you need to do is tune to a station transmitting FAX (with digital readout radios, tuning the PK-232 is no more complicated than subtracting 1.7 kHz from the transmitting station's published frequency). That's a heck of alot easier than it was with the old Hammerlund HQ-110's "coarse" and "fine" tuning controls. Then just configure the PK-232 to FAX mode, turn on the printer, and Bingo! Out come the maps and charts you've been waiting for.

You can also transmit FAX pictures. Frankly, because transmitting FAX requires a special program, I didn't try this option, so I can't comment on the PK-232's capabilities in this area. AEA is currently developing an MS-DOS program for FAX transmission. Details are too sketchy to report. However, early versions include FAX display on screen, transmission capabilities, and a number of other features. Availability is scheduled for late fall or early winter.

SIAM. One of the first things you notice when you tune across the hf bands is the variety of different digital signals. Even if you were an expert and could tell by sound alone, it would take time to configure the PK-232 to receive these signals. With SIAM, the PK-232 analyzes the signal and identifies the type of transmission and its speed. The operator can then decide whether to receive the station or continue on with a band search.

SIAM will decode a number of different digital codes: ASCII, ARQ, and FEC AMTOR and Baudot. It will also decode the Russian Cyrillic and Japanese Katakana codes.

To use SIAM, all you do is type in the command **OPMODE SIGNAL**, confirm that the receive (DCD) LED is lit, and wait approximately 10 seconds. The PK-232 will respond with a baud rate indication and a confidence of mode factor. In another 15 seconds, the PK-232 will identify the signal. To copy the signal, all you do is type in the command *OK*. If the SIAM analysis is correct, you'll start seeing text. If not, the PK-232 will give you a **?bad** prompt.

If the text is decoding but seems to be encrypted, you can try setting **BITINV** to **0 through 31**, if only simple bit-inversion encryption is being used. If none of these 32 settings will decode the station, chances are another more sophisticated encryption system is being used.

conclusion

Typically, AEA has included a well written owner's manual that describes the operation of both WEFAX and SIAM modes fully. In case of difficulty, AEA has listed a number of common faults and the appropriate fixes. They also offer excellent advice by telephone if the manual and a little bit of work fail to solve the problem.

If you have a PK-232 and haven't yet modified it, you're missing out on a treat. I wouldn't just walk — I'd *run* to place the order!

AEA, 2006 196th Street SW, Lynnwood, WA 98036.

Circle #302 on Reader Service Card.

N1ACH

NEW products

miniaturized DTMF encoders

Pipo Communications has introduced the P-7 and P-8 series of miniaturized DTMF encoders designed for custom installation in radios or systems that are exposed to harsh or abusive environments. Built with steel keys and sealed gold dome contacts to ensure reliability and long life, the P-7 and P-8 encoders will fit most radios.

The P-7, a 12-key touchtone encoder, comes in vertical (P-7V) or horizontal (P-7H) formats measuring 2.16 inches by 1.5 inches by 0.20 inches. The P-8, a 16-key touchtone encoder, is available in a vertical (P-8V) format only; it measures 2.16 inches by 1.9 inches by 0.20 inches. Both are available in black or dark brown.

For more information, contact Pipo Communications, P.O. Box 2020, Pollock Pines, California 95726-2020.

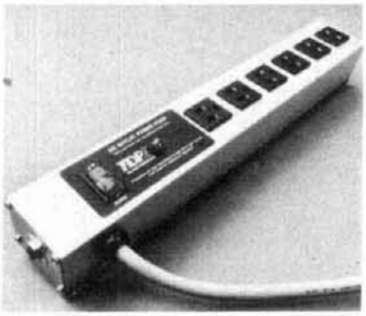
Circle #303 on Reader Service Card.

power strip spike protector

TDP's Model 10PS101 Power Strip Voltage Spike Protector has been specifically designed to guard solid-state circuits against potentially damaging voltage spikes by absorbing and dissipating them through two 50-joule MOVs without interfering with normal current flow. All six outlets are circuit-breaker protected and rated at 15 amps.

In addition, the 10PS101 features a bi-directional noise filter to eliminate both RFI and EFI interference. The noise filter functions over a broad band (100 KHz to 20 MHz); high-frequency signals in this range are attenuated by up to 30 dB for improved equipment performance.

UL-listed and American-made, the 10PS101 has six NEMA-type plug-ins and a heavy-duty,



three-wire grounded 6-foot power cord. MOV working status is confirmed by a built-in indicator lamp.

For more information, contact TDP Electronics, 111 Old Bee Tree Road, Swannanoa, North Carolina 28778.

Circle #304 on Reader Service Card.

all-mode 440-MHz base station transceiver

ICOM has introduced the IC-475A 440-MHz base station transceiver. This deluxe all-mode base receives from 430 to 450-MHz and has 99 tunable full-function memories, passband tuning, a notch filter, noise blanker, built-in SWR bridge, semi or full CW break-in, and a multifunction meter. The new IC-475A also has a velvet-smooth tuning knob and easy-to-read amber LCD readout with variable backlight.

Four scanning systems are available: band, programmable, mode, and memory scan that scans 99 memories in five seconds, with selectable lock-out. The IC-475A features exciting new options such as a tone squelch unit, speech synthesizer, and OSCAR module that allows tracking with a companion IC-275A or IC275H;

NEW products

an FL-83 500-Hz, 10.749-MHz CW filter, and an AG-35 mast mounted pre-amp.

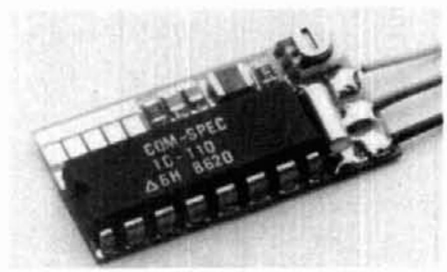
All subaudible tones are built in, and the actual subaudible frequency is displayed. Standard repeater splits are also built in, and odd splits are programmable. The 25-watt IC-475A is similar in design to ICOM's compact base station line: the IC-735, IC-275A, IC-275H and IC-575A.

The IC-475A is priced at $1,999.00. More information is available from ICOM America, Inc., 2380 116 Avenue N.E., P.O. Box C-90020, Bellevue, Washington 98009-9029.

Circle #305 on Reader Service Card.

CTCSS encoder

Communications Specialists of Orange, California has introduced a small programmable CTCSS encoder for use in handheld radios and other size-restricted applications. Dubbed the SS-32SMP, this new encoder measures only 0.53 × 1.00 × 0.16 inches, and offers full-tone versatility and high audio level.



Any 32 tone frequencies between 0.01 and 255.0 Hz may be selected for storage into a 32-bit EEPROM memory. These tone frequencies may be standard or non-standard, and may be changed at a later date if desired. The required tone frequency may then be selected by soldering binary coded jumpers on the tone board. The SS-32SMP may also be ordered to work as a six tone encoder (no switching diodes are necessary) at no extra charge. Multiple tone switching over six tones can be done with switching diode networks or a binary switch. Tone frequencies above 255.0 Hz may be ordered for a slight additional charge.

The SS-32SMP features a low-impedance, low-distortion, adjustable sine wave with adequate audio level to provide sufficient deviation for most handheld radios. It operates on 6 to 15 Vdc so that voltage dropping resistors should never be required.

Priced at $27.95, the SS-32SMP is covered by a one-year warranty and is available for immediate delivery from factory stock. A catalog is available on request; contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #306 on Reader Service Card.

NEWS BULLETIN

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BACK ISSUES HR Magazine from Vol. 1 No. 1 thru 1986, except 2 issues. $150for all postpaid. Also PopTronics, RE, 73 back to 1961. $15.00/full year. Write with your needs. Bill Fossman, 632 Wetmore, Everett, WA 98201.

MARCO: Medical Amateur Radio Council, Ltd, operates daily and Sunday nets. Medically oriented Amateurs (physicians, dentists, veterinarians, nurses, physiotherapists, lab technicians, etc.) invited to join. Presently over 550 members. For information write MARCO, Box 73's, Acme, PA 15610.

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COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC, ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

CONNECTICUT: November 15. SCARA Indoor Ham Radio and Computer Flea Market, N. Haven Park and Recreation Center, 7 Linsley St, N. Haven. Sellers admitted at 7 AM; buyers from 9 AM to 3 PM. Tables are $10 in advance, $15 at the door. General admission $2 per person. Talkin on 146.61 MHz. Reservations for tables must be prepaid by November 4, 1987and no reservation by phone. For information or reservations SASE to: SCARA, POB 81, N. Haven, CT 06473 or call Brad at (203) 265-6478 between 7 PM and 10PM.

ILLINOIS: November. Rockford Hamfest, Forest Hills Lodge, 9900 Forest Hills Rd, Rockford. 8 AM to 4 PM. Tickets $3 advance, $4 at door. Amateur Radio and computer dealers, indoor flea market, outdoor tailgating, radio/computer forums, ARRL speakers, exams, free parking. All on one level. WHEELCHAIR ACCESSIBLE. For advance tickets SASE to Rockford Hamfest, 6514 Swansdown drive, Rockford, IL 61111. For booth/table reservations SASE to Roger Sawvell, KD9MQ, 6514 Swansdown Drive, Rockford, IL 61111 or call (815) 282-1283.

NORTH CAROLINA: November 21 and 22. The 7th annual Greensboro Hamfest Franklin Blvd, National Guard Armory. Sponsored by the MARK IV Radio Club. 9 AM to 5 PM. Tickets $4 advance, $5 at gate. New tailgate area. Ticket plus $2 per space. Walk in FCC exams. Information/registration: Fred Redmon, N4GGD, 3109 Goodall Drive, Greensboro, NC 27407 (919) 852-9244 from 0100Z to 0300Z. Tickets only: Henry Hughes, KA4LPA, 2811 Gwaltney Rd, Greensboro, NC 27407. FCC exams: Hugh Brunson, AE4N (919) 852-1087.

ALABAMA: November 14 and 15. The Montgomery ARC's 10th annual Central Alabama Montgomery Hamfest, Ed Teague Arena, Central Alabama State Fairgrounds near Coliseum. Free admission, free parking. Overnight RV parking with hookups $5/night. Flea market and dealer setups Friday 7 to 10 PM Saturday and Sunday 6 AM. Tables $5 each/day or $7 each/both days. No reservations needed. Doors open to public 9 to 4 PM. FCC license exams both days starting 9 AM. For information: Montgomery Hamfest, POB 3141, Montgomery, AL 36109 or call Randy (205) 832-4598 or Ken (205) 271-0028.

COLORADO: November 29. The Denver Radio Club's annual Hamfest and ARRL State Convention, Jefferson County Fairgrounds, 6th Avenue and Indiana, Golden. 9 AM to 2 PM. Swap tables, seminars, code contests. Non-ham events. Admission $2. Tables $5. Talk in on 147.93/33 and 146.52. Contact Dean Haworth, AC0S (303) 279-4956 for more information.

FLORIDA: November 21 and 22. South Florida ARRL Suncoast Convention sponsored by the Florida Gulb Coast ARC Council, St. Petersburg Hilton and Towers. Hugh flea market. Amateur exams. Saturday QCWA luncheon. Tech talks and demos. Registration tickets $4 to November 13. $5/door. For convention information write FGCARC, 1556 - 56thAvenue North, St. Petersburg, FL 33703.

WISCONSIN: November 14. The Milwaukee Repeater Club is sponsoring the third annual "6.91 Friendly Fest", Serb Hall, 51st and Oklahoma Avenue. 8 Am to 1 PM. Sellers 7 AM. All on ground floor with easy access. Rain or shine. On site Amateur exams. Tickets $3. 4' tables $4. Save $1 per ticket or table–SASE with payment to Milwaukee Repeater Club, POB 2123, Milwaukee, WI 53201 by November 7, 1987. Talk in on 146.91 and 146.52.

MASSACHUSETTS: November 21. The Honeywell Bull 1200 Radio Club and the Waltham Amateur Radio Association will hold their annual Amateur Radio and Electronics Auction, Honeywell Bull plant, 300 Concord Road, Billerica. Snack bar and bargain parts store. Doors open 10 AM. Free admission and parking. Talk in on 147.72/12 and 146.04/64 repeaters. For information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

GEORGIA: October 31 and November 1. Ham Radio & Computer EXPO '87 sponsored by the Alford Memorial Radio Club. Gwinnett County Fairgrounds, Lawrenceville, 20 minutes NE of Atlanta. VEC exams both days, covered flea market, free parking, RV sites with hookups, convenient lodging. $5 admission includes Saturday night cookout. For more information: EXPO '87, POB 1282, Stone Mountain, GA 30086.

MINNESOTA: December 5. The annual Handi-Ham Winter Hamfest, Eagles Club, Faribault. Registration starts 9 AM. Handi-Ham equipment auction. Dinner at noon. Program follows. License exams. Talk in on 19/79. For information: Don Franz, W0FIT, 1114 Frank Avenue, Albert Lea, MN 56007.

OHIO: November 22. The Massillon ARC will sponsor AUCTIONFEST '87, Massillon K of C Hall, off Rt 21. 8 AM to 5 PM. Sellers setup 7 AM. Admission $3.50 advance and $4/door. Tables available $7/8' space. Refreshments available. Free parking. Auction starts 11 AM. Talk in on W8NP, 147.78/.18. For advance registration and information SASE to MARC, POB 73, Massillon, Ohio 44646.

OPERATING EVENTS

"Things to do . . ."

November 8: Armored Forces Amateur Radio Net will commemorate Veteran's Day 0600Z thru Wed Nov 11 2400Z. 80, 40, 20, 15m. For certificate send large SASE to WB1DWR, 16 Berkely Circle, Newington, CT 06111.

November 8: To observe Veteran's Week, members of the Hamfester Radio Club, Chicago, will operate from the Hines VA Hospital's Robert K. "Pappy" Wade,K9CDH, Memorial Ham Shack using Hine's club call K9WFN, 1500Z to 0300Z. 40, 20, 2m FM and 2m USB. For a commemorative certificate, send QSL, QSO number and 9x12 SASE with 39 cents postage or $1.00 to Hamfesters Radio Club, Inc., Chicago, c/o Robert K. "Pappy" Wade Memorial Ham Shack, Bld 8, Hines VA Hospital, Hines, IL 60141.

November 25: Cocos – Keeling Island. Listen for Hans, F6GVD and Victor, G3AAG. For two weeks. No specific frequency or time except first 10 minutes of the hour they will stand by for QRP stations only and at half hour for handicapped operators. QSL via QSL manager VK9YC or direct to F6GVD.

HAM EXAMS: The MIT UFH Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday November 18, 7 PM, MIT Room 1-150, 77 Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann at (617) 646-1641. Exam fee $4.25. Bring a copy of your current license (if any), two forms of picture ID, and a completed form 610 available from the FCC in Quincy, MA (617) 770-4023.

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ELMER'S NOTEBOOK

Tom McMullen, W1SL

receiver buzzwords

Because this is the annual receiver issue, I'll try to clarify some receiver-related terms you might be wondering about. Who knows? This information might even help you choose a rig from among the many available.

One of the biggest problems an Amateur faces is interference from other signals, so I'll emphasize techniques that help reduce that problem.

RIT stands for Receiver Incremental Tuning, a way of tuning the receiver without disturbing the transmitter frequency. This feature is useful if the station you're listening to drifts slightly, or if you have to tune just a wee bit off frequency to minimize the effects of interference. In the earlier days of single-VFO transceivers without RIT, when you moved your receiver dial you also moved your transmitter. The other station then had to move to tune you in. Then you'd move again — and so on. This led to the two stations "walking" each other across the band; if they weren't careful, they could wind up out of bounds. With RIT, you can leave the transmitter alone and move the receiver a few Hz to keep the other station tuned in.

A **direct conversion** or **single conversion** receiver is perhaps the simplest type of heterodyne receiver in use. It's basically a local oscillator (usually a very stable VFO) and a balanced mixer. The rf signal and the VFO are both fed into the mixer, and when the two frequencies are the same (zero-beat with each other), any audio present on the rf signal becomes a product of the mixing process. This audio is fed to audio amplifiers to drive a headset or a speaker, as needed. An rf amplifier is often used ahead of the mixer, and an audio filter after the mixer to prevent hearing the beat notes produced by nearby signals. Many low-power or portable stations use this type of receiver because of its simplicity, light weight, and low power requirements. While **direct conversion** isn't directly related to interference problems, it's worth knowing about this type of receiver in order to understand other discussions to come.

If a single-conversion receiver converts once — from rf to audio frequencies — then it follows that a double-conversion receiver converts twice. The first conversion mixes an rf signal with the first local oscillator, which produces an intermediate frequency signal, abbreviated *i-f.* (Notice the hyphen; it's there so you'll know not to read "i-f" as the word "if.")

The second conversion mixes the i-f signal with a second local oscillator to produce . . . Uh, here's where our nice, neat scheme falls apart!

Older receivers, designed to handle amplitude modulation (a-m), used a diode detector at the end of the i-f amplifier circuit. The second local oscillator would produce another i-f signal, which would then be detected by a diode, to produce audio, etc. In this case, it's a double-conversion receiver. However, newer receivers, designed to handle single-sideband (SSB) signals, needed a different arrangement. Diode detectors are practically useless for SSB detection, so a circuit called a **product detector**, amazingly similar to the single-conversion scheme I mentioned earlier, was developed. This circuit includes a local oscillator that mixes with the i-f signal to produce audio output. The product detector works well with a-m and CW, and really shines with SSB signals.

So whether a receiver is double or triple conversion really depends on the definition of the detector circuit. For example, a receiver that uses a VFO to produce an i-f of, say, 9 MHz, and then a second oscillator and mixer to produce a lower i-f at, perhaps, 455 kHz, which is followed by a diode detector, is clearly a double-conversion receiver.

Does changing the diode detector to a product detector make the receiver **triple conversion**, then? The purists will say yes, but many manufacturers just don't mention it. (And it really isn't important unless you need to know the definition in order to prove a point!)

Double-conversion receivers are important, however, in solving **image problems**. To follow me through this one, **image rejection**, you'll need to look at **fig. 1** and check the arithmetic.

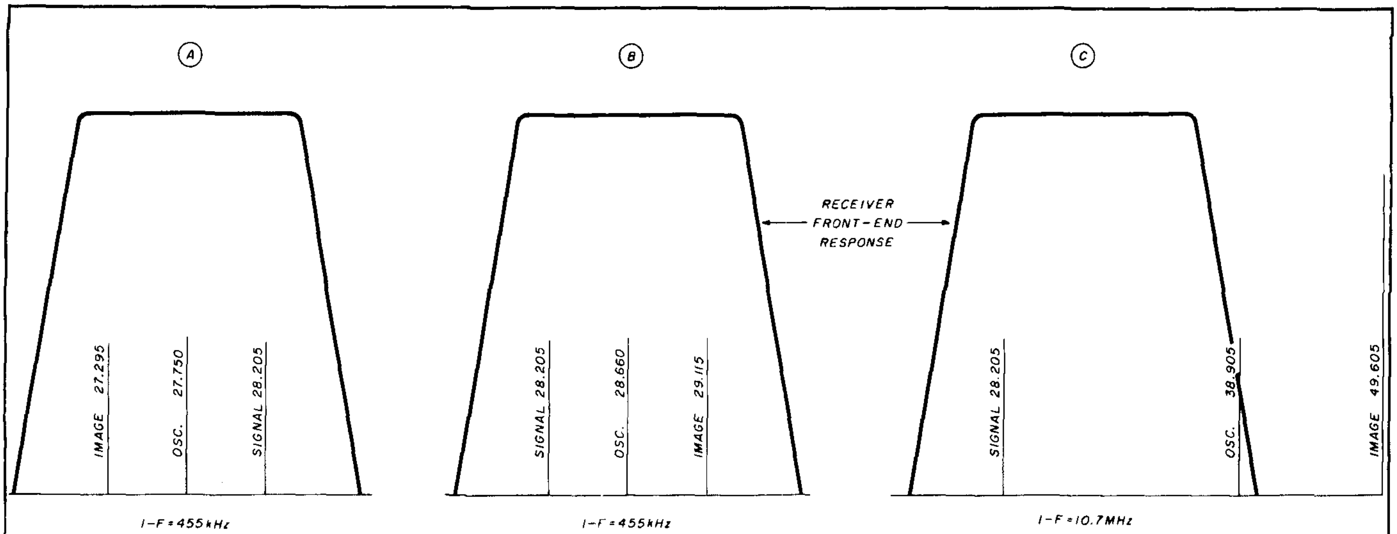

fig. 1. Image rejection is often poor in receivers with a low-frequency i-f, as at A. When the first i-f is high, the image moves away from the receiver front-end response, shown at C.

If the signal you wanted to listen to was the only one on the band, life would be simpler. But there are plenty of signals out there, and most of them aren't the ones you want to hear. A simple receiver (such as a single-conversion type) is capable of receiving all sorts of things you don't want, many of them signals that aren't in the Amateur bands. Let's say you want to tune in a signal at 28.205 MHz. Your i-f is 455 kHz, so your VFO is at 27.750. This produces the right i-f, but if there's a loud signal at 27.295 (as there often is), it too can mix with the VFO to produce 455 kHz.

This intruder is called the **image signal**. An interesting thing is that the image signal is always twice the first i-f away from the signal you want. In the case above, 2 x 455 = 910, and 28.205 − 0.910 = 27.295. If you know that, you can track down suspected image interference.

How do you get rid of it? There are a couple of common tricks: one is to place the VFO above the signal — at 28.660 to receive 28.205 MHz, for example. This would place the image at 29.115, which means that the signal would be from an Amateur station — but that's not much help if it buries your QSO.

A more practical method is to use a double-conversion receiver and make the first i-f high — perhaps 9, 10.7, or even 70 MHz. The high i-f places the image quite some distance away from where the receiver front end is tuned, and it's easier for simple tuned circuits to reject a signal that's far removed from its design frequency. For instance, a 10.7-MHz i-f will place the image of 28.205 at 49.605 MHz, as shown in **fig. 1**. Even a mediocre front end can reject that one.

Another common cure is to build filters (special tuned circuits) that pass only a narrow range of frequencies, say from approximately 27 to 31 MHz for the 10-meter band, but greatly attenuate anything outside that range. Many modern solid-state receivers use this technique on all bands, along with broadband rf amplifier sections, to provide good performance and require minimum attention from the operator.

SINAD isn't a remedy for sinus trouble; it's a test commonly used to determine how well a receiver hears a weak signal. SINAD is an acronym for *signal + noise + distortion* to *noise + distortion* ratio. (Aren't you glad they shortened it?) This is what it means:

If you disconnect the antenna from your receiver and turn the audio gain up, you'll hear a hiss or rushing noise. That's the "noise" part of the formula. It takes a certain amount of signal strength to be heard through that noise. The level of that signal is the "signal" part of the formula. The "distortion" part comes in because a signal can be loud but not clear.

The signal generator used for this test is modulated, and the recovered audio from the receiver is compared with the modulation waveform to see if it has been distorted — and usually, it has. Thus, the test will determine how strong a signal it takes to produce some specific output above the noise, and how badly the i-f filters, audio amplifiers, and even the power supply hum have distorted the audio output.

Fortunately for us, the test results are neatly summed up on a meter on the test equipment, and we don't have to spend a lot of time calculating ratios and such. The test instrument, naturally enough, is called a SINAD meter.

The usual method of rating a receiver under this test is to state a signal strength that's required to meet a particular dB SINAD ratio: for example, 0.5 μV for 12-dB SINAD. The lower the microvolt (μV) number, the better the receiver, and 12 dB is an industry-wide benchmark used in the test.

An i-f notch filter (or i-f notch tuning) is another device used to reduce the effect of interference. It consists of a high-Q circuit that works within the "window" or passband of the i-f amplifier. Instead of acting as a bandpass device, however, it's a band-stop circuit designed to attenuate greatly whatever frequency it's adjusted for. This is most useful if you're listening to a weak signal (or even a moderately strong one) and someone pops up close enough to create an ear-splitting whistle or lots of "splatter" on top of the signal you want to hear. By adjusting the i-f notch, you can often reduce the commotion to a level you can live with.

An i-f notch filter will have limited usefulness, however, because any notch deep enough to really eliminate an interfering signal will also reduce the strength of the signal you want to hear.

Interference suppression can also be handled by an **audio filter** that cuts off all audio tones outside a narrow range. This is great for CW signals, but voice (SSB) tends to sound hollow and distorted if the audio passband gets too narrow. There is an equivalent to the i-f notch filter called the **audio-notch filter** or **audio-rejection filter**. A piece of electronic trickery that uses op amps to create a phase shift that will cancel the offending tone, it works quite well, and usually consists of one or two integrated circuits, a potentiometer or two, and a few resistors and capacitors. Both the frequency of rejection and the degree of rejection (depth of the notch) are adjustable. The audio-notch filter is a great add-on for direct-conversion receivers.

Another neat bit of electronic sleight-of-hand is **i-f shift**. It takes some of study to figure out how it works, but it's really quite simple. In essence, it works this way: when a signal is in the same i-f passband as the one you want to hear, you just move the passband "window" over a bit until that signal is outside it. You can do the same thing by slowly tuning your receiver until the interference is out of the passband, but when you do that, the signal you want is moving also, and might end up as a tone that's not easy to hear (on CW) or as duck-talk (on SSB).

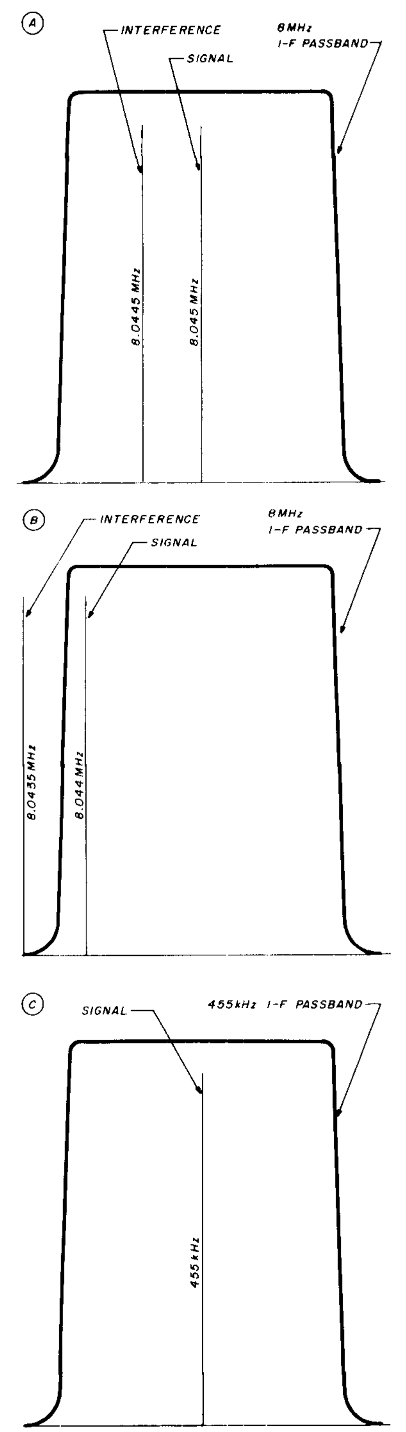

fig. 2. I-F shift, or i-f tuning, can move an interfering signal out of the i-f passband, then place the wanted signal back in the low-freqency i-f window for clear reception. This example uses approximately 8 MHz for a passband filter, then converts down to a conventional 455-kHz i-f for further amplification and detection.

Perhaps **fig. 2** will help clarify this. At **A**, both the interfering signal and the wanted signal are in the i-f passband. By adjusting the i-f shift, you can move the interfering signal out of the passband, as at **B**, and leave the wanted signal in. Now, the only trick is to get the wanted signal back into the center of the window again, as at **C**.

How do they do that? By adding two more conversions in the i-f amplifier chain, and using the same beat-frequency oscillator (BFO) to mix with the signal twice. For example, you can mix a 455-kHz signal with an 8.0-MHz BFO to produce an 8.455 i-f. This i-f signal then passes through a filter and into another mixer. There, the 8.455-MHz signal mixes with 8.0 to produce the original 455 kHz i-f again.

Now to exercise the cranium a bit more — with the help of **fig. 2** — and get more specific: let's say that the signal you want comes through the i-f amplifiers at 455,000 Hz (455 kHz). The one that's bothering you comes through at 455,500 Hz. This produces the normal result at **A**, with both signals inside the i-f passband. By detuning the 8.500-MHz VFO a few Hz to 8.499, and mixing with the i-f signals, you can change the wanted signal to 8.044, and the interference is at 8.0435 MHz, as shown at **B**.

The 8-MHz i-f filter will pass 8.044, but not 8.0435, so the interference is gone! When 8.044 is again mixed with 8.499, the difference is 455 kHz, which is right back in the middle of your i-f window (at **C**), just where you want it.

Of course, this is a greatly simplified example of how it works, and I've made the frequency separations large to make the example easier to follow, but you get the idea. Many manufacturers use more complex circuits to accomplish this, and some use phase-locked loop (PLL) circuits to move the signals around and reject the unwanted ones. No matter how they do it, the results make it worth looking for this feature in a receiver.

conclusion

The degree of interference rejection you need depends upon the type of operating you do. On most of the hf bands, crowding is a way of life, and the best DX is usually buried beneath several layers of loud signals. It seems to be magic when an experienced operator can peel away those layers to leave an S1 or S2 signal standing alone, perfectly readable among S9 + "locals." By carefully choosing your interference-fighting weapons, and with some practice, you too can become a magician.

Next month, I'll explore the possibilities of 1200 MHz.

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