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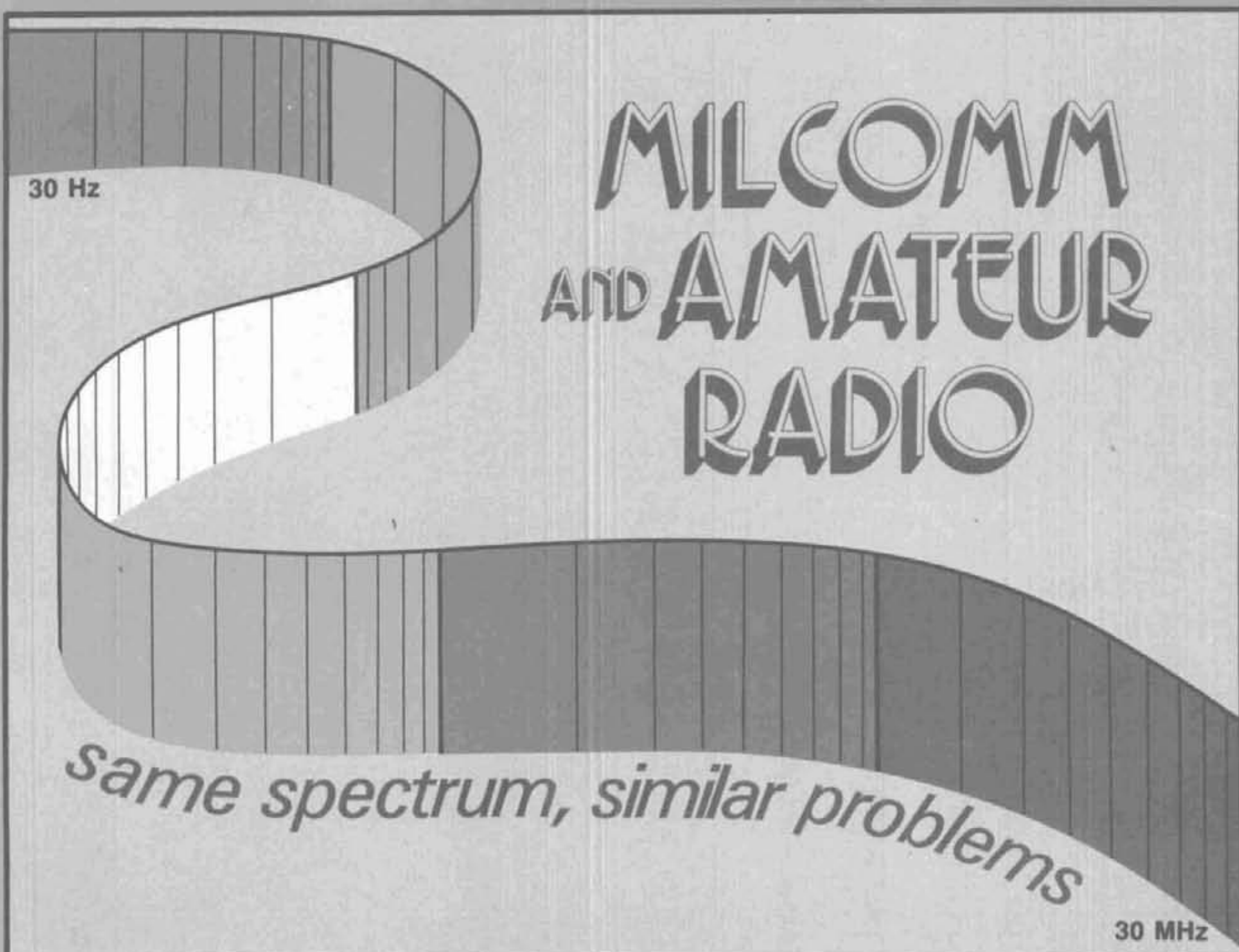

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focus on communications technology

a PSK demodulator for OSCAR-10 • run RTTY on your computer • DC dummy load • carrier-operated relay for VHF amplifiers • controlled vertical radiation rhombics, part 2 • a state-of-the-art electromagnetic jargon generator • plus W6SAI, W1JR, KØRYW, and the Guerri report

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Here's what other hams have to say about the "dream rig."

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John J. Schultz W4FA
CQ Magazine
September 1984

"...we seriously doubt anyone finding a unit superior to ICOM's new 751 HF 'dream rig.'"

Dave Ingram K4TWJ
Computer Trader Magazine
September 1984

"The general-coverage receiver is excellent."

Mark Wilson AA2Z
QST Magazine
January 1985

"The Notch measured 55dB, and is the best ICOM Notch yet."

"The stability of the 751 deserves mention. We measured 10Hz drift in the first hour."

Robert Pohorence N8RT
International Radio, Inc.
September 1983

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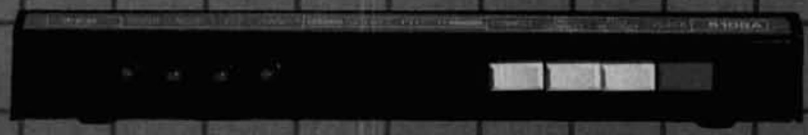
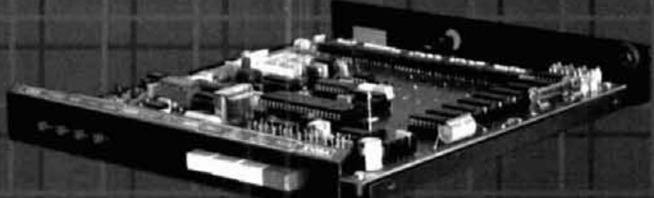
All stated specifications are approximate and subject to change without notice or obligation. All ICOM radios significantly exceed FCC regulations limiting spurious emissions. 751285

With SMART PATCH You are in CONTROL



With CES 510SA Simplex Autopatch, there's no waiting for VOX circuits to drop. Simply key your transmitter to take control.



SMART PATCH is all you need to turn your base station into a personal autopatch. SMART PATCH uses the only operating system that gives the mobile complete control. Full break-in capability allows the mobile user to actually interrupt the telephone party. SMART PATCH does not interfere with the normal use of your base station. SMART PATCH works well with any FM transceiver and provides switch selectable tone or rotary dialing, toll restrict, programmable control codes, CW ID and much more.

To Take CONTROL with Smart Patch – Call 800-327-9956 Ext. 101 today.

What To Look For In A Phone Patch

The best way to decide what patch is right for you is to first decide what a patch should do. A patch should:

- Give complete control to the mobile, allowing full break in operation.
- Not interfere with the normal operation of your base station. It should not require you to connect and disconnect cables (or flip switches!) every time you wish to use your radio as a normal base station.
- Not depend on volume or squelch settings of your radio. It should work the same regardless of what you do with these controls.
- You should be able to hear your base station speaker with the patch installed. Remember, you have a base station because there are mobiles. ONE OF THEM MIGHT NEED HELP.
- The patch should have standard features at no extra cost. These should include programmable toll restrict (dip switches), tone or rotary dialing, programmable patch and activity timers, and front panel indicators of channel and patch status.

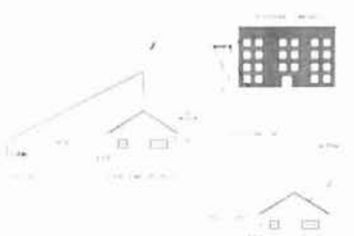
ONLY SMART PATCH HAS ALL OF THE ABOVE.

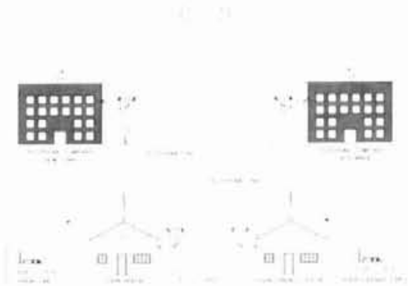
Now Mobile Operators Can Enjoy An Affordable Personal Phone Patch. . .

- Without an expensive repeater.
- Using any FM tranceiver as a base station.
- The secret is a SIMPLEX autopatch, The *SMART PATCH.*

SMART PATCH Is Easy To Install

To install *SMART PATCH*, connect the multicolored computer style ribbon cable to mic audio, receiver discriminator, PTT, and power. A modular phone cord is provided for connection to your phone system. Sound simple? . . . IT IS!





How To Use *SMART PATCH*

Placing a call is simple. Send your access code from your mobile (example: *73). This brings up the Patch and you will hear dial tone transmitted from your base station. Since *SMART PATCH* is checking about once per second to see if you want to dial, all you have to do is key your transmitter, then dial the phone number. You will now hear the phone ring and someone answer. Since the enhanced control system of *SMART PATCH* is constantly checking to see if you wish to talk, you need to simply key your transmitter and then talk. That's right, you simply key your transmitter to interrupt the phone line. The base station automatically stops transmitting after you key your mic. *SMART PATCH* does not require any special tone equipment to control your base station. It samples very high frequency noise present at your receivers discriminator to determine if a mobile is present. No words or syllables are ever lost.

SMART PATCH Is All You Need To Automatically Patch Your Base Station To Your Phone Line.

Use *SMART PATCH* for:

- Mobile (or remote base) to phone line via Simplex base. (see fig 1.)
- Mobile to Mobile via interconnected base stations for extended range. (see fig. 2.)
- Telephone line to mobile (or remote base).
- *SMART PATCH* uses SIMPLEX BASE STATION EQUIPMENT. Use your ordinary base station. *SMART PATCH* does this without interfering with the normal use of your radio.

WARRANTY?

YES, 180 days of warranty protection. You simply can't go wrong.

An FCC type accepted coupler is available for *SMART PATCH.*

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contents

matrix operation

It started innocently enough. First there was a single response from Europe. "Yes, K2RR, you are being heard here and your signal is 5 and 9." Then someone broke in from the States asking if he could just exchange a signal report with the G station. Before you could say "phased 4-element vertical array" five times, the size of the group on frequency had swollen to 10 Europeans, 1 African, 1 New Zealander, and 400 U.S. stations, give or take a few.

What ensued is what I like to call "matrix operation," whereby a number of stations across the Atlantic get an opportunity to talk to at least *scores* of U.S. stations in very short order. (Consider the phenomenon as a 12 by 400-term matrix and look at the large number of possible combinations and permutations — i.e. QSOs.) This is exactly what happened the other morning on 75 meter sideband during a 2-1/2 hour period in which a number of stations from Western Europe worked rare (for them) states like Nevada, Oregon, and Washington. It also provided many stations in the Western states the opportunity — possibly their first — to talk to Europe on 75 meters.

So what does this prove?

It proves that we **ARE** civilized, contrary to the impression some people might come away with after casually scanning the lower HF bands these days.

These days, more often than not, it appears that some normally decent, law-abiding Amateurs are willing to jam and curse each other out over a piece of precious modern "real estate" — our frequency spectrum — each claiming that he was there first. Others decide to make "critical" adjustments to their transmitters at full power while established QSO's are in progress. The overall result is that what was supposed to be a relaxing hour or so of operating turns into the nearest thing to bedlam, creating ill will — not only among ourselves in the States, but because of the long-distance nature of propagation on the low bands nowadays, other countries as well.

May I be so bold as to share with you some simple suggestions on how to improve low-band operation and keep the collective blood pressure down?

LISTEN *before* transmitting. A local conversation might be taking place on frequency — or a DX contact might be occurring even though you can hear only the closer station.

ASK whether the frequency is in use *after* you listen a bit. It's quite possible that even though you can't hear the stations clearly, they can hear you very well.

ADAPT your operating procedure to the conditions at hand. For example, if a rare DX station is working four stations a minute and you manage to get his attention, don't monopolize him *and* the frequency with a long soliloquy.

ADJUST your equipment for best performance (minimum distortion, spurs, chirp, drift, etc.).

USE the *minimum* power necessary to establish and maintain the contact. Save money: leave your amplifier off as much as possible.

BE CREATIVE. If conditions permit, enhance the rubber-stamp type of DX contact with information of genuine interest — but only if conditions permit (see "ADAPT").

***DON'T* BE CREATIVE.** It is both amusing and sad to hear so many one-way conversations on frequency. The old adage, "If you can't hear them, you can't work them," still holds.

CALL the *least* number of times. Don't try to be the last one heard. Listen to how some top operators do it; sometimes they drop their call in only once — but at the appropriate moment.

DON'T BROADCAST, *communicate.* There are plenty of interesting people out there who have something worthwhile to say.

ENHANCE your knowledge of the band by listening, reading articles on propagation, and noting relationships between the WWV forecasted indices, geomagnetic field status and band conditions.

OBSERVE established "windows." On 75 meters, for example, 3790-3800 is still the international DX window. Your signal at 3800.0001 (LSB), even if perfectly clean, will wipe out the possibility of DX contacts from 3800 down several kilohertz.

LEARN at least a few words in another language. The joy of communicating is yours for the asking.

Though my experience is based on years of operating on 75/80 meters, in general these principles apply to other bands as well. What about your favorite bands and modes? I am very interested in learning about the operating habits, procedures and standard and anomalous propagation modes specific to your band. Drop me a line. Who knows? A cumulative set of notes from these responses could evolve into a pamphlet useful to all.

(Reader's responses to the February editorial "One Million Years of Experience," in which we asked for your suggestions about how the growth of Amateur Radio might be encouraged, continue to pour in. Many thanks to all who've written — a detailed summary of your varied ideas will appear in a forthcoming issue.)

Rich Rosen, K2RR
Editor-in-Chief

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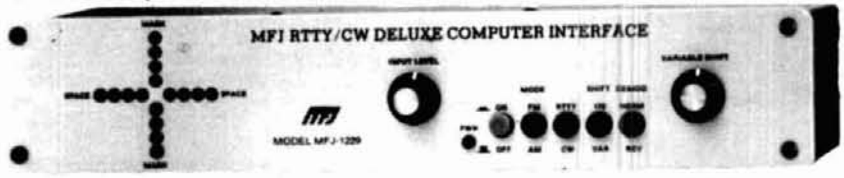
MFJ'S MOST ADVANCED RTTY/ASCII/AMTOR/CW COMPUTER INTERFACE HAS FM, AM MODES, LED TUNING ARRAY, RS-232 INTERFACE, VARIABLE SHIFT TUNING, 170/850 Hz TRANSMIT, MARK-SPACE DETECTION.



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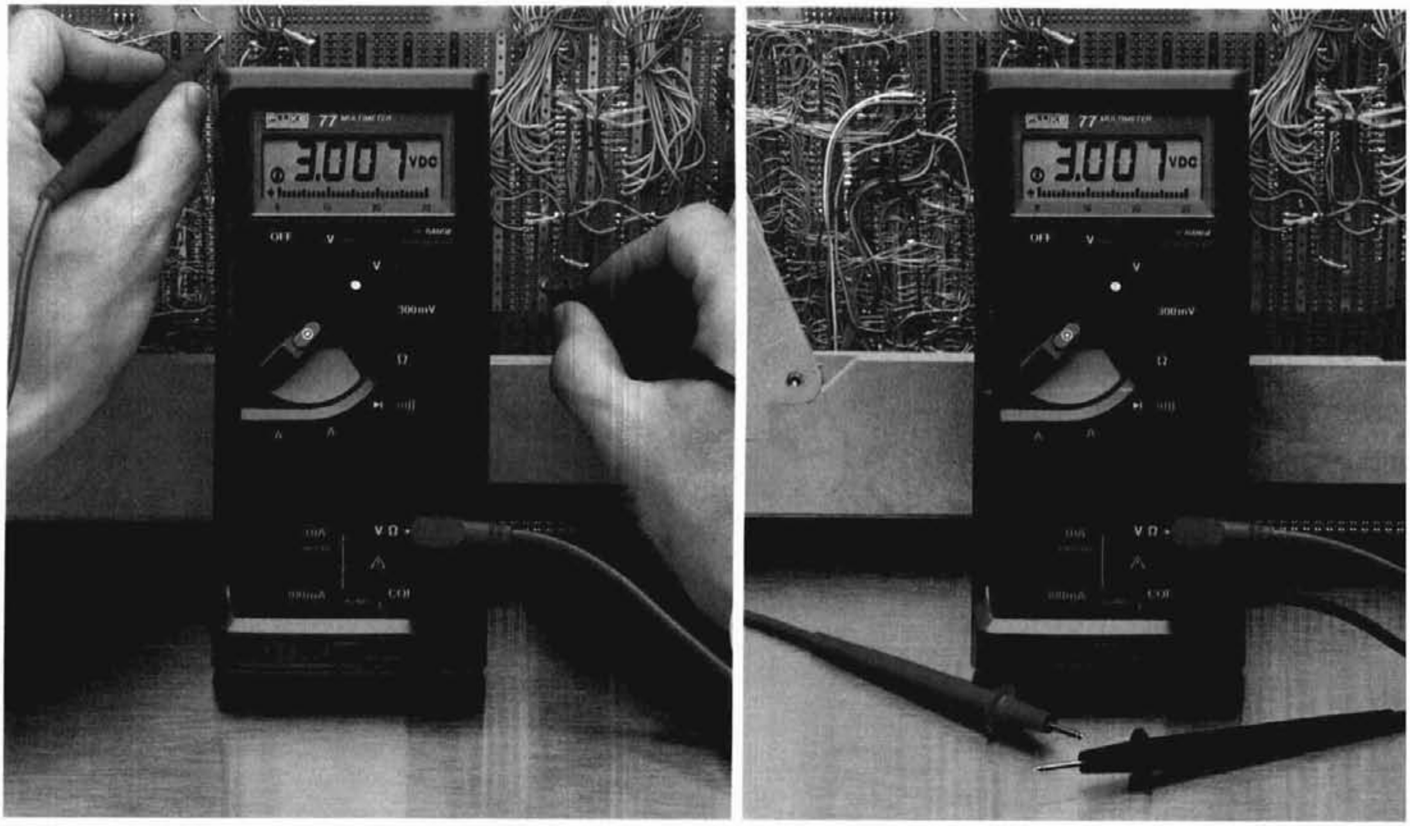

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** Patent pending.

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AMATEUR RADIO DOES HAVE A FUTURE, BUT A STRONG EFFORT MUST BE MADE to assure that future...that was the consensus of the all-day industry meeting held in Miami January 30, just prior to the Tropical Hamboree. About 40 people, representing many major manufacturers, several distributors, the principal Amateur publishers, and the ARRL attended the session, chaired by former HR Report editor Joe Schroeder, W9JUV.

Amateur Radio Does Have Some Grave Problems, those attending agreed: the static U.S. Amateur population, an uncomfortably high rate of unrenewed licenses, the on-going problems of both manufacturers and dealers, and awareness that our influence in Washington seems to be waning, were just some of the symptoms that were cited. However, during the "off-the-record, no attributions" discussion, a number of really worthwhile suggestions were made.

An Aggressive Program To Attract Junior High School Youngsters was one of the key ideas —junior high science teachers can provide access to this group. Free passes to hamfests for youngsters, their teachers and/or parents was one suggestion, to be complemented by an industry-sponsored and staffed "This Is Ham Radio" booth to provide both an introduction and even "hands-on" experience for newcomers. Supplementing this effort will be an Amateur Radio "comic book" highlighting its "fun" aspects in an entertaining way. Also planned are "sales pitches" to the general public, directing a special effort toward responsible CB groups such as REACT and scanner organizations. The possibility of making the entry level more attractive, by adding limited Novice data and/or voice privileges, was also discussed. Greater dealer involvement, such as hosting training courses and club meetings, was also considered. A proposal that industry representatives sit on the ARRL board, to encourage closer ARRL-industry coordination, was broached to ARRL representatives.

Overall Attitude After The Grueling Session Was Upbeat, with those attending feeling that some very real progress had been made in overcoming Amateur Radio's current inertia. Volunteers from the group are already actively working on a number of the suggested programs, and progress reports on their efforts plus a discussion of future plans is scheduled for a Dayton meeting the Thursday evening before the Hamvention.

FCC'S REPEATER FREQUENCY COORDINATION PROPOSAL will place prime responsibility for resolving repeater conflicts squarely on operators of uncoordinated repeaters. In the case of a dispute with two coordinated machines, both operators would share responsibility for its resolution equally. In its Notice of Proposed Rule Making, the FCC cited the rapidly increasing number of repeater interference complaints it's received, and that the bulk of these problems seem to involve uncoordinated repeaters.

The FCC Asks Amateurs To Consider A Number Of Key Questions in this potentially far-reaching NPRM. For example, should coordination be mandated in major urban areas? As an alternative, should narrow-band technologies and tone squelch be required to minimize the interference problem? Should the Commission recognize a "single national frequency coordinator," either on a national basis or as an "advisor" to local coordinators?

A Blanket Moratorium On New Repeaters Initially Distracted Attention from the real issues until lifted by the FCC February 19. There had been some comment that the moratorium had simply confused the issue and was probably unenforceable anyway, and the ARRL (supported by the Tri-State Repeater Council) had petitioned the FCC to rescind the ban.

Comments On The Repeater Coordination Docket, PR Docket 85-22, are due at the FCC July 1; Reply Comments will be due September 30.

TEXAS HAS JOINED THE SHIFT TO 20 KHZ 2-METER SPACING, adopting the change by an 8:1 margin at the February 16 meeting of the Texas VHF FM Society. The decision, which makes Texas the second state east of the Rockies to make the move, has been under consideration for some time. As yet, no timetable for the move has been established.

Northern California And Minnesota Are Also Looking At The 20 kHz Plan. In California the Northern Amateur Relay Council has invited Clay Freinwald, K7CR, to its April 13 meeting at Concord to discuss the 20 kHz plan. K7CR is considered one of the fathers of the plan in the Pacific Northwest. Minnesota has 20 kHz on its April meeting agenda, too.

The Escalating Move To 20 kHz Is Causing Concern In Europe where 2-meter FM operation on their 144-146 MHz band has traditionally been on 25 kHz centers. Some Europeans are worried that the Japanese, whose domestic 2-meter band is also on 20 kHz spacing, would drop 5-kHz step capability from their rigs if the U.S. market no longer needs it.

TWO ESTABLISHED REGIONAL VECS ARE GOING NATIONAL. Both the DeVry Amateur Radio Society and Metroplex have applied for national status, based on their experience and success in operating regional programs in the Ninth and Second call areas, respectively. The DeVry proposal is particularly interesting, as DeVry has campuses in seven other call areas that could serve as nucleii for its efforts in those areas. As the FCC's experience with both organizations has been excellent, their proposals should be accepted quickly.

WARC BAND EXPANSION IS STILL HANGING FIRE, with some GMRS representatives looking into the possibility of carving a slice for personal radio from the new 902-928 MHz band.

The Commission's Recent Turndown Of A Personal Radio Service at 900 MHz has also been challenged by the same group, who've filed a Petition for Reconsideration with the FCC.

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HRA-432	420-450 MHz	$59
HRA-( )	150-174MHz	$69
HRA-( )	450-470 MHz	$79

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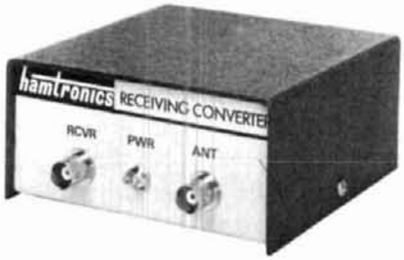

Models to cover every practical rf & if range to listen to SSB, FM, ATV, etc. NF = 2 dB or less.

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Kit with Case $49	50-52	28-30
Less Case $39	50-54	144-148
Wired $69	144-146	28-30
	145-147	28-30
	144-144.4	27-27.4
	146-148	28-30
	144-148	50-54
	220-222	28-30
	220-224	144-148
	222-226	144-148
	220-224	50-54
	222-224	28-30
UHF MODELS	432-434	28-30
Kit with Case $59	435-437	28-30
Less Case $49	432-436	144-148
Wired $75	432-436	50-54
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Model XV2	28-29	145-146
Kit $79	28-30	50-52
Wired $149	27-27.4	144-144.4
(Specify band)	28-30	220-222*
	50-54	220-224
	144-146	50-52
	50-54	144-148
	144-146	28-30
For UHF,	28-30	432-434
Model XV4	28-30	435-437
Kit $99	50-54	432-436
Wired $169	61.25	439.25
	144-148	432-436*

*Add $20 for 2M input

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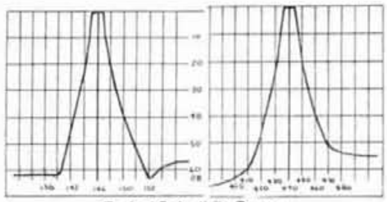

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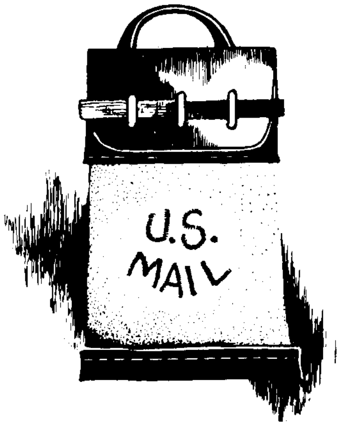

comments

Amateur Radio — not what it used to be?

Dear HR:

As a lifetime subscriber, I have a few comments about K2RR's account of his maritime experiences ("Reflections," November, 1984, page 5). I was in the Merchant Marine during World War II, but as a fireman/water tender/oiler. I was discharged the day I was to get my commercial ticket and never went to sea again.

K2RR only hinted at what is missing from Amateur Radio today. Gone is the clean language, the exchange of technical information, the invitations to visit (especially to mobiles), the hospitality we used to have, and in short, good manners on the air.

I'm glad I was born sooner and had the opportunity to enjoy Amateur Radio for the last 48 years. I had it at its best.

Albert Kaufman, W1JVQ
Bridgeport, Connecticut

user group agrees on 23-cm band plan

Over 100 users of the 23-cm band (1240-1300 MHz) reached agreement on an updated regional plan to serve the needs of the southern California area for the next three years at a meeting in Orange County, California. Present and participating at the meeting, sponsored by the Southern California Repeater and Remote Base Association (SCRRBA), were representatives of all users' modes currently operational, or likely in the future to require spectrum, on 23 cm.

Each of the 23-cm band users' groups (ATV simplex, ATV repeaters, weak signal/experimental, FM voice repeaters/links, digital, satellite/AMSAT, VRAC, VUAC, and SCRRBA) selected a representative to participate in a four-hour roundtable and negotiation session. The resulting plan — basically a modification of the existing plan — was prepared specifically for use in their region. Key stipulations provide that the plan will remain in effect for the next three years, after which time a similar meeting will again be held to review the existing band utilization patterns, and that the frequency allocations of existing users' groups will change only as new users' groups begin to operate on the band.

Because the region leads the nation in terms of 23-cm band activity, SCRRBA suggests that the band plan shown in **table 1** may be useful to other coordination councils in preparation of their own regional band plans.

— SCRRBA

table 1. A comparison of present and modified 23-cm band utilization plans (courtesy SCRRBA, P.O. Box 5967, Pasadena, California 91107).

band segment	present usage	new usage (initiated as needed)
1240 - 1246	ATV repeater (Channel 1) (1241.25 video carrier, VSB filtering required)	ATV repeat (Channel 1) (1241.25 video carrier, VSB filtering required)
1246 - 1248	narrow-band FM point-to-point links (voice)	same plus narrow-band digital (<50 kHz BW)
1248 - 1258	ATV repeater (Channel 2) video carrier 1253.00	
1248 - 1251.5		wideband digital (>500 kHz BW)
1251.5 - 1252		guard band
1252 - 1258		ATV repeater (Channel 2) video carrier 1253.25, VSB filtering required
1258 - 1260	narrow-band FM point-to point links (voice)	same plus narrow-band digital (<50 kHz BW)
1260 - 1270	satellite uplink ATV repeater (Channel 4) 1277.00 video carrier	satellite uplink, plus non-coordinated simplex: *experimental* wideband, no repeater inputs/outputs
1270 - 1272	FM (voice) repeater inputs; 1271.000 "test pair" input	FM (voice) repeater inputs; 1271.000 "test pair" input
1272 - 1282	ATV repeater (Channel 4) 1177.00 video carrier	
1272 - 1275.5		FM repeater future expansion, ACSB systems, linear translators

grid dipping

Dear HR:

Even though George A. Wilson, Jr., W1OLP, in "Matching Dipole Antennas," (May, 1984, page 129) made at least 24 separate references to GDO (Grid Dip Oscillators) and Grid Dipping, someone is certain to try substituting a solid-state dipper, (such as the Heathkit HD-1250 or one of several factory assembled versions) when exciting the RF Bridge discussed in the article. In fact, with the solid-state dipper far more prevalent today than the old vacuum tube grid dip oscillator (and interchangeable in most applications), no doubt a large number of hams who build the RF bridge will end up frustrated and with no discernible "dip."

While the solid-state dippers can be

used to determine resonance, per the first part of George's article, it is not likely to provide enough excitation to obtain a reading with the RF bridge unless overcoupled, with sensitivity set at maximum, and with an extremely sensitive μA meter used as the detector. Even a 50 μA meter will probably not allow a discernible "dip" to be obtained!

A rough idea of a dipper's suitability can be obtained by connecting a germanium diode and a small 2 to 3-turn link in series across the μA meter's terminals. Coupling the link to the dipper's coil should easily produce a full-scale reading. If it does not, the dipper cannot be used to excite the RF bridge.

Robert G. Wheaton, W5XW
San Antonio, Texas

receiver input temperature

Dear HR:

Amateur Radio literature never mentions (or I've never seen) the temperature contributed by a transmission line to receiver input temperature. None of this mattered with hot, noisy amplifiers, but it could matter significantly with the low-temperature amplifiers coming into use. If someone can come up with a convincing argument that the transmission line adds no temperature of its own, I'd like to see it. Such a line would make an ideal cold-source for noise measurements.

Picture the line as a string of small attenuators generating their own thermal noise, which is sent in both directions along the line. The noise energy has an equivalent temperature at the line's Z_0, which is some fraction of the line's ambient temperature.

I derived the temperature output for a small attenuator (0.1 dB, 6.769 degrees K) and summed the cumulative temp (with cumulative losses), generating a chart of temperatures for various losses. It closely follows the equation

$$T = 298 \left[1 - \text{antilog} \left(\frac{-\text{dB}}{10} \right) \right] \text{degrees}$$

I deliberately ignored the possibility of any energy reflected from the antenna.

This sugggests that 30 feet of RG-59B/U could add 127 degrees K at the input to a 432-MHz amplifier, and 6 feet would do the same at 1296 MHz (approximately). 15 feet of 1/2 inch hard-line could add 36 degrees to the input at 1296 MHz.

I'd like to hear if anybody has any thoughts or can suggest any references on this.

Martin Sample, WA6JTD
PO Box 1245
Tuolumne, California 95379

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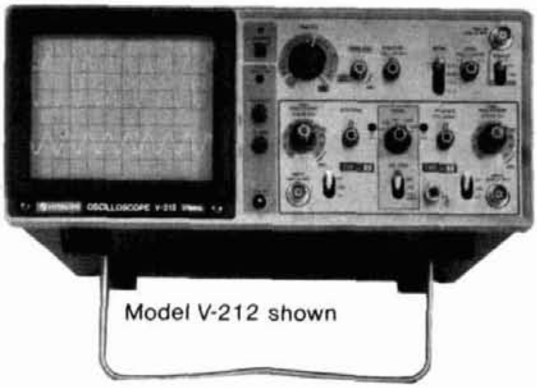

Model V-212 shown

Model V-1050F shown

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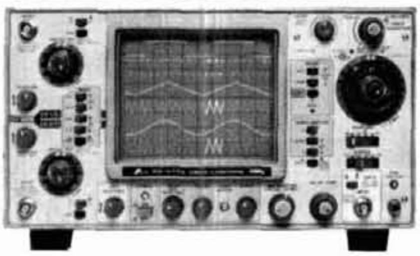

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Vertical and horizontal deflection accurate within ±2%. CRT acceleration voltage 12KV. 3 channels, 6 traces. High precision calibrator (±1%). Fastest sweep rate: 10 ns.
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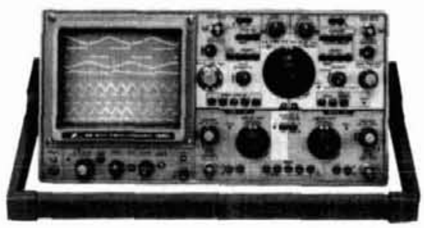

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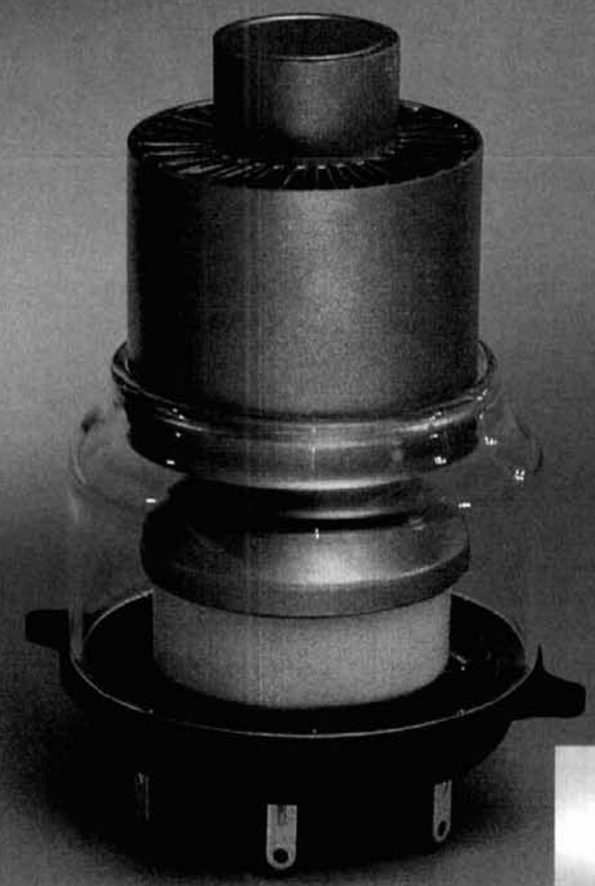

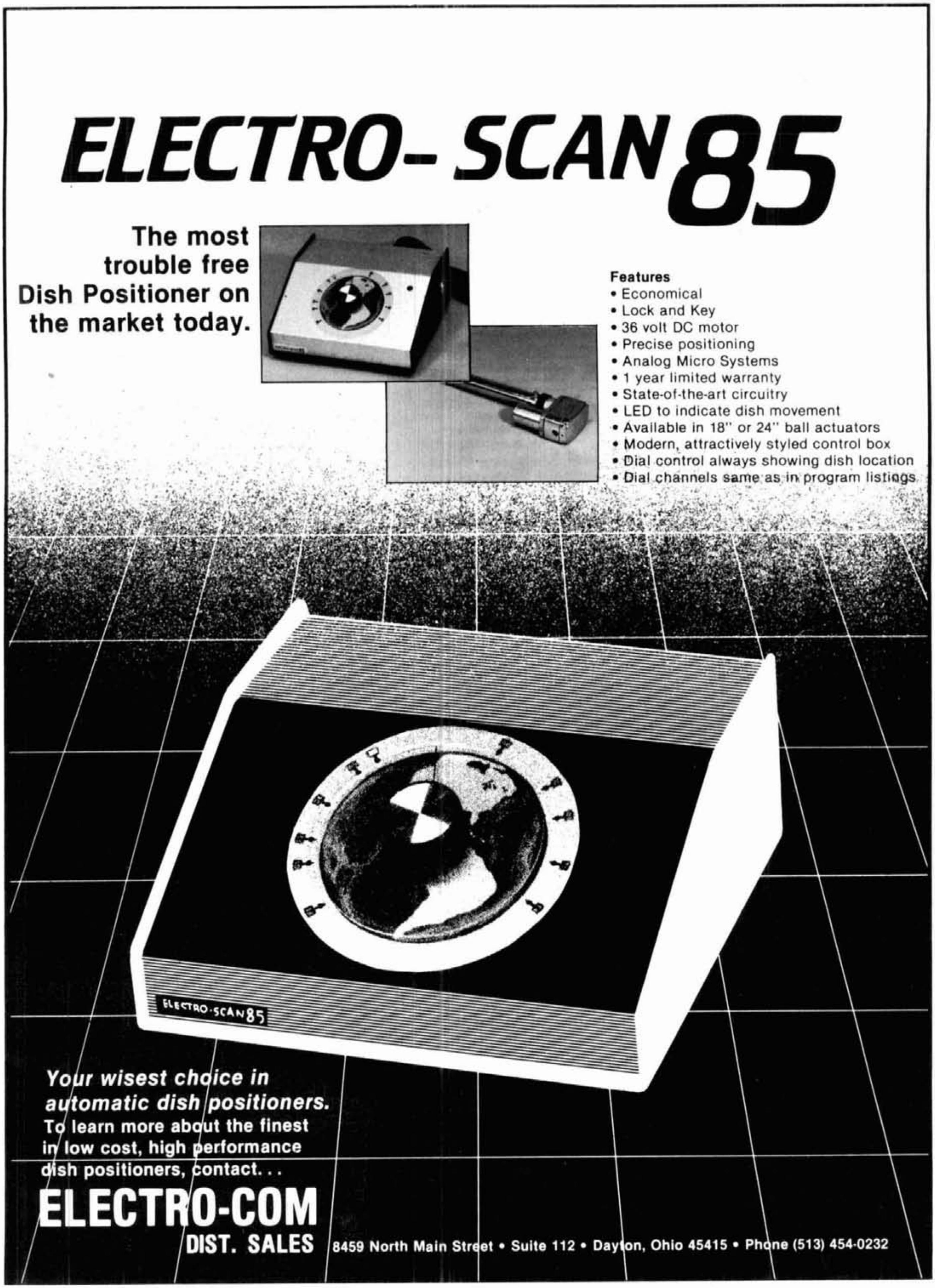
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digital HF radio: a sampling of techniques

Tomorrow's technology is at work in the military today

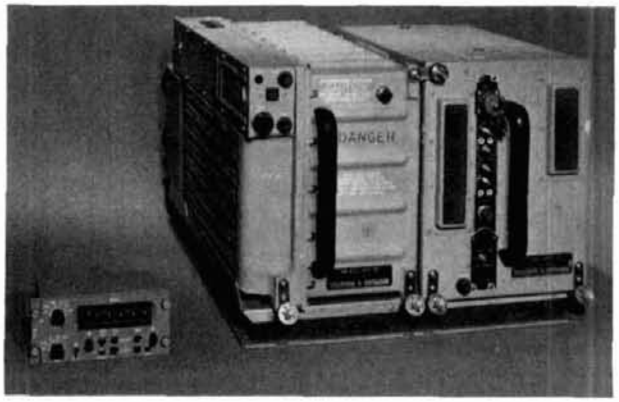


fig. 1. P-3C AN/ARC-161 HF radio set.

Thinking of military communications equipment, one conjures up the image of sturdy, olive drab-colored, compact all-inclusive designs that meet stringent operating requirements. On the other hand, when one thinks of Amateur Radio equipment, a different picture comes to mind. Instead of the predictable "military" designs, one imagines a diversity of commercial homebrew items with little uniformity of design or form.

But which spectrum environment — military or Amateur — is more congested? Which demands more stringent operating requirements? The answer is not obvious.

From his unique vantage point as president of Communications Consulting Corporation, Dr. Rohde, an experienced receiver designer and active ham provides insight into how military equipment designers solve their problems of congested spectrum using the most sophisticated techniques and materials. Some circuits will doubtless appear in Amateur Radio applications — perhaps a few may already have done so.

— Editor.

Until recently, HF radios for the military (fig. 1) have been designed and built in the traditional analog way, with selectivity obtained through the use of LC or crystal filters in the IF section and active filters in the audio frequency section. These radios have been used for point-to-point operation where only infrequent change of operating frequency was required. In addition, these point-to-point connections were used with constant output power.

In order to meet communications goals for 1985 and beyond, particularly in military applications, modern HF equipment must be adaptive, frequency agile and capable of supporting secure digital voice communications. It must be capable of operation on both a point-to-point and networked basis as well. Adaptivity is needed both to control transmitter power to a level no greater than required for the connection, and to select frequencies which provide good propagation with a minimum of interference.

Frequency agility is desirable in the event of deliberate jamming or rapid change of propagation conditions. Transceivers must have the ability to *change frequency rapidly enough to adjust to changes* in environment. While propagation conditions normally change relatively slowly, avoiding a jammer requires a repetitive fast change of frequency called *frequency hopping*. For a jammer to be effective in disrupting communications, it must either be extremely powerful, use a high-gain antenna, and cover a wide frequency range, or operate narrowband and try to predict or detect the frequencies on which the frequency hopper is operating and jam only those.

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

Generally, the optimum hop rate for communication is determined by a trade-off between implementation cost and a combination of operating considerations which include expected propagation delays (related to distances between jammer and communicators), capabilities of enemy direction-finding and jamming equipment, and required communications distances, bandwidths, and reliabilities. Taking all of the above into account, recent Army, Air Force, and Navy requirements have focused on hop rates in the range of several hundred to several thousand hops per second as needed to satisfy most military situations (actual rates are classified information). Frequency-hopped signals faster than 1 millisecond or so are difficult to locate with currently-deployed tactical direction-finding hardware and would require a costly "Fast-Follow" jammer to track and jam the communications on a hop-by-hop basis.

other processing techniques required

Often frequency hopping alone is insufficient to defeat deliberate jamming. Modern HF transceivers must be equipped with sophisticated signal processing techniques that provide several levels of redundancy and/or error correction capabilities. These techniques, involving more complex circuitry and larger instantaneous bandwidths, allow recovery of desired signals even in the presence of high levels of natural noise or deliberate jamming. The modern HF transceiver has an RF portion which must be "transparent" to the real brains of the transceiver, the digital signal processing circuits.

Finally, military HF radios must be designed to support the transmission of secure voice and data. Although many techniques exist for manipulating analog voice signals to provide privacy, the U.S. military has settled on the encryption of digitized voice as being both easier to accomplish and more secure. Also, to overcome frequency selective fading characteristics of HF, techniques for digitizing voice have included bandwidth compression as well. This is because narrowband signals suffer less distortion. What is sought is the lowest possible bit rate that produces acceptable speech quality when converted back to its analog form.

voice encoding technique

Linear predictive coding (LPC is the compression scheme currently favored by the U.S. military for HF links. Used at a bit rate of 2.4 kbps in the ANDVT (U.S. Navy Advanced Narrowband Digital Voice Terminal), LPC encodes the voice as numbers derived from the instantaneous spectral characteristics of the voice. The numbers themselves have no relationship in an analog sense to the original voice signal, but they are used in an inverse process to produce an approximate analog signal resembling the original voice. The process is not unlike that used in children's toys that speak (e.g., "Speak and Spell"™).

Present military standards have settled on LPC-10, a linear predictive coding/decoding algorithm which compresses 3 kHz speech to a 2 kilobits per second data stream. The algorithm uses a linear mathematical relationship to predict the value of each successive sample it is digitizing and hence the name.

To make the compressed data secure, it is encrypted by combining it in a unique mathematical fashion with a string of numbers generated by a "key" generator. The result is a new succession of data carrying voice information encrypted by the "key" and the method of combining it with the voice data.

At the receiver, demodulation depends on synchronously detecting the transmitted bits even though they cannot yet be converted to intelligible voice. Then, by proper mathematical application of the same "key" used to encrypt the original data, the decrypted bit stream is applied to the inverse of the LPC process, and the spectrum-related numbers are converted back to analog speech.

The entire process depends heavily on maintaining good channel quality so that accurate bit timing at the receiver may be achieved; the most important aspect is to preserve timing relationships. A decision at the receiver as to whether a received symbol (character) is a "one" or a "zero" must not be made during a symbol transition period. As the data rate increases it becomes more difficult to insure that the receiver is making its decisions at the proper time. Decision errors, the result of distortion that occurs between characters (inter-symbol distortion) cause degradation in the overall system performance. Timing relationship distortions are the result of frequency selective phase shifts that arise from differences in path length as the signal component at each frequency passes through and reflects from different layers of the ionosphere. The narrower the instantaneous bandwidth of the radiated signal, the fewer the perturbations. Consequently, the best results are achieved by using the narrowest instantaneous bandwidth design.

With simple forms of modulation such as a Binary FSK, Bi-Phase FSK, or even AM, 6 kHz or more bandwidth would be needed to send the 2.4 kbps LPC signal. To fit the signal into a standard *3 kHz* voice channel requires more sophisticated modulation schemes. For example, to achieve 1 Hz/bit packing density, ANDVT uses a 39-tone parallel modulation scheme. The 2.4 kbit data stream is split into a number of parallel data streams transmitted at a lower rate. According to the ANDVT algorithm, each of the slower streams modulates one or more of the 39 tones.

✓ 112

The resulting tones are summed in an analog summer and applied as conventional analog modulation at the exciter.

This parallel modulation approach trades complexity and transmitter efficiency for narrow bandwidth. The transmitter (at maximum power) has to be able to handle the case in which all tones add in-phase (maximum power out) and, as a result, must operate well below peak output power most other times. Complexity results from having to handle parallel channels both at the transmitter and receiver. At the receiver, careful tracking of frequency and phase is needed to insure that demodulation of each tone occurs properly. Further, for the frequency hopping case, differences in path length at each frequency must be compensated for on a hop-by-hop basis. Nevertheless, these techniques do work and are being incorporated, primarily in software-based modems into new military HF equipment.

An alternate approach under study at RCA is a serial modulation scheme which processes blocks of data taken in sequence from the data stream. Each block is encoded as one of several tones. If 6 bits of data are taken at once and the modem has 64 (2^6) available tones in a 3 kHz audio bandwidth to pick from, then each 6 bits of data determines which one of those tones is to be transmitted. The selected tone is transmitted with a duration of roughly six times the original bit duration; the transmitter operates at full output power; and the instantaneous bandwidth, once the system is synchronized, is one-sixth that of the parallel scheme. This approach has been implemented and tested at RCA. Its penalties include a complex synchronization and demodulation algorithm and a rather unique (and objectionable) on-air "signature."

key digital building blocks

Traditional analog circuit techniques are inadequate to meet the previously stated requirements. The modern HF transceiver must be based not only upon new intelligence, but on a new set of "building blocks" as well. Many of these building blocks are digital functions performed entirely by software routines; others, involving high computational rates, are better implemented by using dedicated digital hardware. Some of the more obvious building blocks include the following:

Modems. Modems, in general, are required to process voice or data and to provide the necessary waveforms, usually at IF, to the exciter and receive them from the receiver. Selection of the waveform is critical since performance of the entire link depends on the features it possesses. The waveform is said to be "robust" if it has at least two levels of redundancy in its synchronization scheme. It must also have built-in error detection and correction as well as preambles, which allow channel quality measurements. In addition, the waveform may have to support a variety of data rates to maintain high quality information transmission as link conditions degrade.

Digital filters. For reasons of flexibility and performance, digital filtering is preferred in the IF and audio frequency sections of the transceiver. The ability to choose an appropriate filter type and time constant greatly simplifies modem design. Also, under software control, it is possible to have a digital filter assume any of the classic shapes or performance characteristics as required. Although either Chebyshev or Bessel characteristics are generally used, others can be selected. On the negative side, use of sophisticated digital filters sometimes requires correction of group delay effects of any remaining LC or crystal filters in the transceiver. Also, real time digital filters require correspondingly fast high-resolution A-to-D and D-to-A converters as well as high throughput microcomputers. Two devices ideally suited to digital filter implementation are the Texas Instrument TMS-320 and the RCA (internally developed) high performance ATMAC II CMOS/SOS Processor.

Frequency synthesizers. Sufficiently fast-switching frequency synthesizers cannot be easily implemented in the traditional analog form using multi-loop PLL synthesizers; instead, the direct digital synthesizer (DDS), to be described later, is required. The DDS features arbitrarily fine resolution and uses a cosine look-up table together with a microprocessor, a D/A converter, and a lowpass filter to generate such waveforms. The DDS can be built with switching times between 1 μs and 50 μs, with the actual switching time dependent on the D/A converter and the number of glitches produced by the sampling and integration process. A limiting signal-to-noise ratio of about 75 dB, which has to do with the sample rate, is the current state-of-the-art.

The direct digital frequency synthesizer output is mixed with the output of a conventional single-loop PLL synthesizer with wide loop bandwidth. This PLL then determines the settling time of the overall system.

Agile antenna couplers. Since wideband antennas aren't commonly used, the transmitter or receiver has to be matched to the antenna on a hop-by-hop basis. The switching speed of the antenna coupler is thus part of the overall system switching time. Conventional relay couplers can be built with 10 ms switching speed relays, and there are claims for future reed relays that will provide 1 ms switching times. However, if 10-100 (or more) frequency changes or hops per seconds are required, the lifetime of the mechanical devices will be soon depleted. The typical lifetime of these devices is about 2 million operations or, at 100

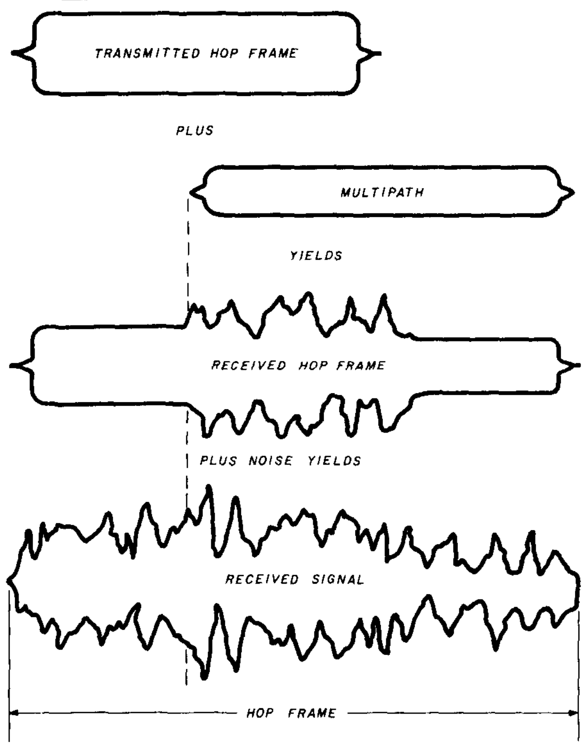
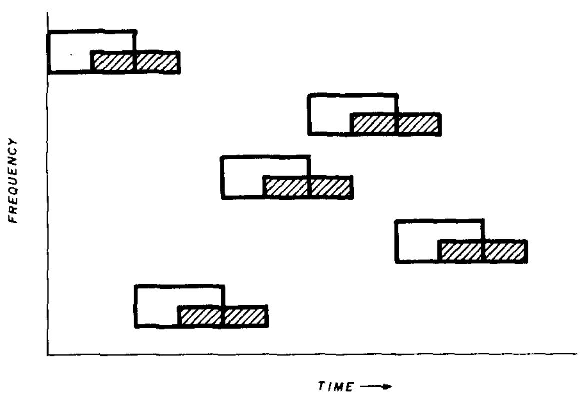


fig. 2. Frequency hopping multipath problem.



Note: Illustration assumes multipath delay equals one-half hop frame. Amplitude is one-half of desired signal.

fig. 2(A). Frequency hopping with multipath.

switches per second, 20,000 seconds — only three hours of continuous use.

Modern couplers, then, must employ solid-state switching, for which the use of PIN diodes is the obvious choice.* Such a coupler uses a quasi-binary coded arrangement of inductors and capacitors that are switched in and out of the matching network by PIN diodes. Because antenna couplers have to be built to handle power levels of up to 1 kW, as much as 8,000 volts DC is required for reverse biasing; forward currents of up to 2 amps are necessary. In order to generate these voltages and currents, special dedicated switching power supplies have to be provided, and a mechanism is needed to bring the voltage or current to the switching diodes without introducing parasitic stray effects.

*See "High Power RF Switching with PIN Diodes," by J.R. Sheller, KN8Z, *ham radio*, January, 1985, page 82.

Having summarized both the basic requirements and some of the key building blocks for modern digital HF radios, we'll now examine some specific design approaches.

frequency hopping receiver design

The need to frequency-hop at HF, especially in the presence of multipath, places the most stringent requirements on the modern HF transceiver. The following section discusses frequency hopping, design considerations, and introduces the unique problems created by frequency hopping. Several means of solving these problems are considered, and a novel solution illustrating the use of digital techniques is presented.

Frequency hopping in a multipath environment. The effect of multipath distortion is depicted in **fig. 2.** A transmitted hop frame represents a burst of signal energy radiated at one frequency and received by a frequency hopping receiver (synchronized to the transmitter). A *multipath* is a burst of identical signal energy, delayed in time and reduced in amplitude. The number of multipaths, amount of relative time delay, and attenuation of the received burst depend on link characteristics such as frequency, separation, and time of day.

What the receiver actually "sees" during its dwell time (window) is a combination of all the energy contained in all the multipath bursts on the frequency. This is illustrated in **fig. 2**, both with and without noise. This burst distortion will be different for each frequency while frequency hopping. The receiver must be able to correctly demodulate the transmitted information in this distorted signal.

Intersymbol interference reduction techniques. If a transmitter is sending a succession of symbols (such as ones and zeros), the receiver's job is to identify those symbols and convert them back to intelligible information. When a signal includes additive noise, errors occur because the difference between the two levels is not as clearly defined. When multipath corruption occurs, symbols actually overlap and intersymbol distortion occurs. This is a special type of distortion which requires more sophisticated processing to overcome. Inter-symbol distortion caused by multipath is generally what limits the throughput capability (maximum data rates) of HF links. A number of techniques have been tried to reduce inter-symbol in-table 1. Current HF signal processing techniques.

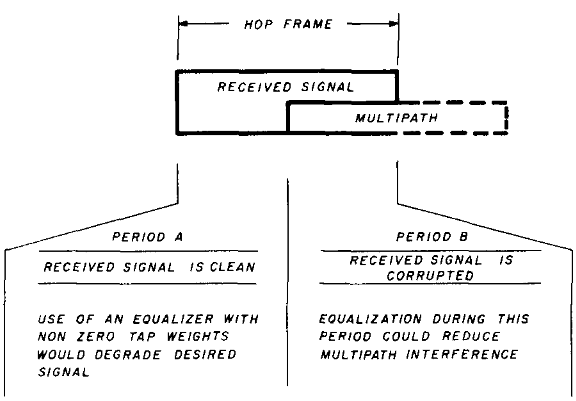
technique	concept	remarks
parallel tone differential quadriphase shift keying (DQPSK) modulation	long bauds on adjacent frequencies — nonadaptive	selective fading causes high error rates high peak-to-RMS ratio transmitter
parallel matched filters or rake processing	pulse matched filter (correlation)	good for low data rates or when WT > 30
linear equalization	minimize distortion by filtering the received signal	severe multipath causes high error rates requires training sequence and updating
decision feedback equalization	minimize distortion by filtering the received signal and past decisions	tracking problem, requires training sequence and continuous update
maximize likelihood estimation	message matched filter viterbi decoding algorithm	exponential growth with multipath delay tracking problem

terference, with the most common method shown in **table 1**. We will concentrate here on equalization.

Equalization techniques. Figure 2A illustrates a simplified model of frequency hopping with multipath. It shows a frequency-versus-time plot of a pseudorandomly hopped signal. Each frequency is represented by a received signal component and a crosshatched multipath signal of one-half amplitude and delayed half the hop frame. Let's look at the problem of implementing an effective equalizer in this hopping environment.

When a single frequency hop frame is examined (assuming receiver synchronization), as in **fig. 3**, the received signal is present for the entire hop frame, while the multipath signal may be observed only during part of the hop frame, identified as period *B*. During period *A*, the first part of the hop frame, the received signal is not corrupted by multipath components, and one would expect only normal atmospheric and receiver noise to be present. Trying to eliminate multipath during this period using feedback equalization would actually degrade the desired signal. In fact, *no* special processing should take place during period *A*. During period *B*, however, the signal is corrupted by multipath and some processing may be used to reduce the multipath interference.

But there is a further complication in that a single communications link or net is not necessarily the sole user of a frequency band or a family of hopped frequencies. This is especially true in the crowded HF band. Consequently, the simplified model must be extended to include a large number of users who may share the same frequencies. The shared use of frequencies by several synchronized nets is illustrated in **fig. 4**.



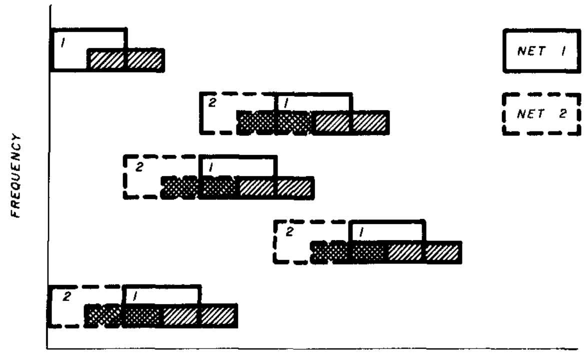
NOTE: ASSUME SIGNAL WITH SINGLE MULTIPATH INTERFERENCE COMPONENT.

fig. 3. Simplified model.

Additional "friendly" stations cause problems. Multiple users in a network present a more complicated interference pattern at the receiver. This is illustrated in **fig. 5**. The hop frame contains the desired signal and two distinct and different interference components. During period *A* there is a new delayed replica of an undesired signal from another user or from another net. This signal has totally random parameters with respect to the desired signal.

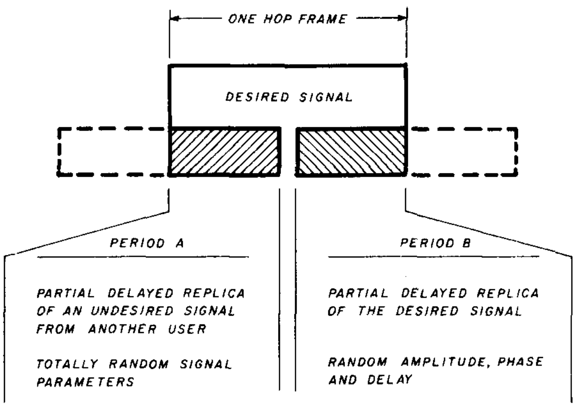
Also if the frequency hopping rate is increased to "outrun" the multipath, interference will be present from the multipath components of other synchronous net users and from others sharing the frequency.

In **fig. 5**, each hop frame is shown as consisting of the desired signal and several different types of interference. During the first part of the hop frame, (period A), random interference is present which does not cor-



Note: Illustration assumes multipath delay equals one-half hop frame. Amplitude is one-half of desired signal.

fig. 4. Two-net frequency hopping with multipath.



Periods *A* and *B* may overlap or have a gap since multipath for each user is unique.

fig. 5. Frequency-hopping interference.

relate with the desired signal. The only known technique to reject this type of interference, which may have the same waveform characteristics as the desired signal, is to use correlation processing, which rejects uncorrelated random noise. This type is processing, achieved by deliberately spreading the spectrum of the original signal at the transmitter and compressing it again at the receiver, cannot be easily achieved at 2.4 kbps data rates. This is because the usable channel bandwidth is already fully occupied by the non-redundant digitized and compressed voice signal. To maintain occupied bandwidths at 3 kHz or so while transmitting data at 2.4 kbps, time-gated adaptive equalization may be used. The name time-gated is applied because it should only be active during period *B*, when the multi-path burst *is* a replica of the desired signal. During period *B*, even a high speed dedicated processor will be taxed. For each multipath component, the individual tap locations must be determined. When frequency hopping over a 10 percent bandwidth, tap location variations corresponding to relative multipath delays may be small, but the amplitude and phase weights differ for each frequency hop frame. Weights may also have to be varied from hop to hop due to doppler shifts encountered on the channel — the result of ionospheric variation, as well as motion of the user. Because of these effects, amplitude and phase of the tap weights cannot be determined only once and revised each time the same frequency is revisited. They must be determined for *each* hop. Prior data may be helpful as an initial estimate, but cannot be relied on for adequate equalization.

sample design approach to multipath processing

An equalizer is actually a matched filter that attempts to model the corrupted channel as a function of time. The input to the equalizer is a composite waveform consisting of wanted signal, noise plus distortion components. The approach to be described is based on an RCA-developed adaptive algorithm which repeatedly calculates the ratio of desired signal to signal plus noise plus distortion. Maximizing the ratio by rapidly adjusting characteristics of the equalizer comprises the adaptive process. When the ratio is maximized, the equalizer is said to be "converged to the value best representing the inverse of the corrupted channel." Passing the corrupted input signal through the adjusted equalizer essentially removes the corruption and gives the best possible signal. The process is very much like that used to equalize trans-Atlantic telephone cables, except that in this process the equalizer characteristics must be adjusted each time the transmitter-receiver pair hops to a new frequency.

The basic operation consists of storing all signal samples received during the dwell on a particular frequency. The digital processor uses the samples to compute the equalization measure described above until a final value is found. In reality, the equalizer resembles a tapped delay line with complex weights applied to the signal developed at each tap. The taps are summed and, using an appropriate algorithm, added to the original input signal. An equalization value is computed for each set of tap weights examined. The tap weights producing the highest equalization value are then applied to the equalizer and the original set of signal samples filtered and passed on for further processing.

For binary FSK signals, the equalization value can be found from spectral energy measurements using a Fast Fourier Transform (FFT). The process consists of taking the spectral energy in the two carrier frequencies and comparing it to the energy in the remaining in-band spectral region after the carrier frequencies have been deleted. The equalizer is optimally adjusted

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when the desired carrier frequency energy, less residual energy, is maximized. Thus, the adaptive process would vary the complex tap weights until that difference is maximized.

Appropriate techniques can be found for other forms of modulation and will be discussed subsequently, but in general, they all compare signal energy in a known region of the spectrum (where the desired signal should have the majority of its energy) to the remainder of the spectrum where multipath components lie.

The RCA approach is unique in that it does not require a known signal to be transmitted at the outset of each new hop frame. Such a signal (termed a "training" signal) reduces the usable throughput of the system since it occupies a portion of every hop frame.

A second attractive feature of this equalizer approach is that any arbitrary signal possessing the desired waveform characteristics (i.e. modulation type and rate) may be automatically equalized by applying the same algorithm. This allows a receiving station in a network to automatically equalize signals from any other station in the network without first having to identify which station it is or decode a special "training" signal.

A block diagram of the multipath processor is illustrated in **fig. 6**. Input signal samples provided by the frequency hopping receiver are stored in an input buffer. When an entire hop frame has been accumulated, the samples are transferred to a second buffer, which stores all the signal samples associated with two hop frames. While the input buffer is collecting new data from a different carrier frequency at the next hop, the *output or hop frame buffer* is used repetitively to perform the equalization. For a frequency hop rate in the low hundreds, there is sufficient time to recirculate the stored data through the equalizer and adjust the appropriate tap weights for highest channel quality.

The output hop frame buffer supplies digitized signal samples to two separate processing functions — the time-gated equalizer and the tap locator. The time-gated equalizer is controlled by the tap locator and the tap weight calculator. After an "equalization complete" signal is provided by the tap weight calculator, the signal samples stored in the receiver signal buffer are passed through the now adjusted equalizer and the corrected data is stored in the output buffer. The signal samples are supplied on demand to the next processing stage, which may be a spread spectrum despreader or data demodulator.

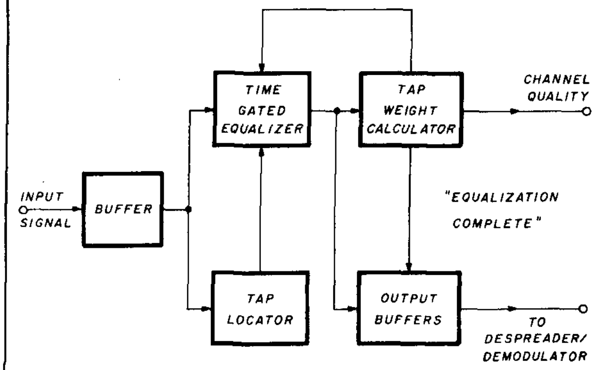
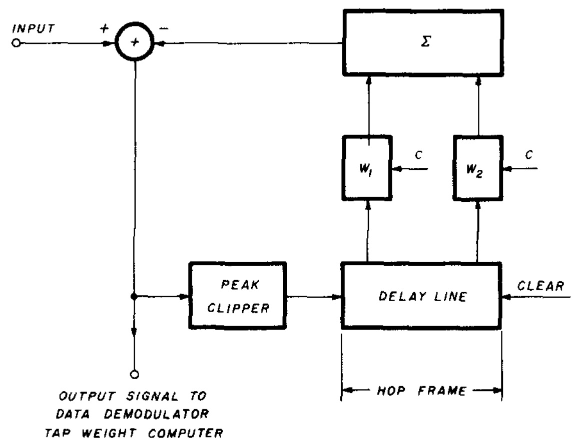


fig. 6. Multipath processor.



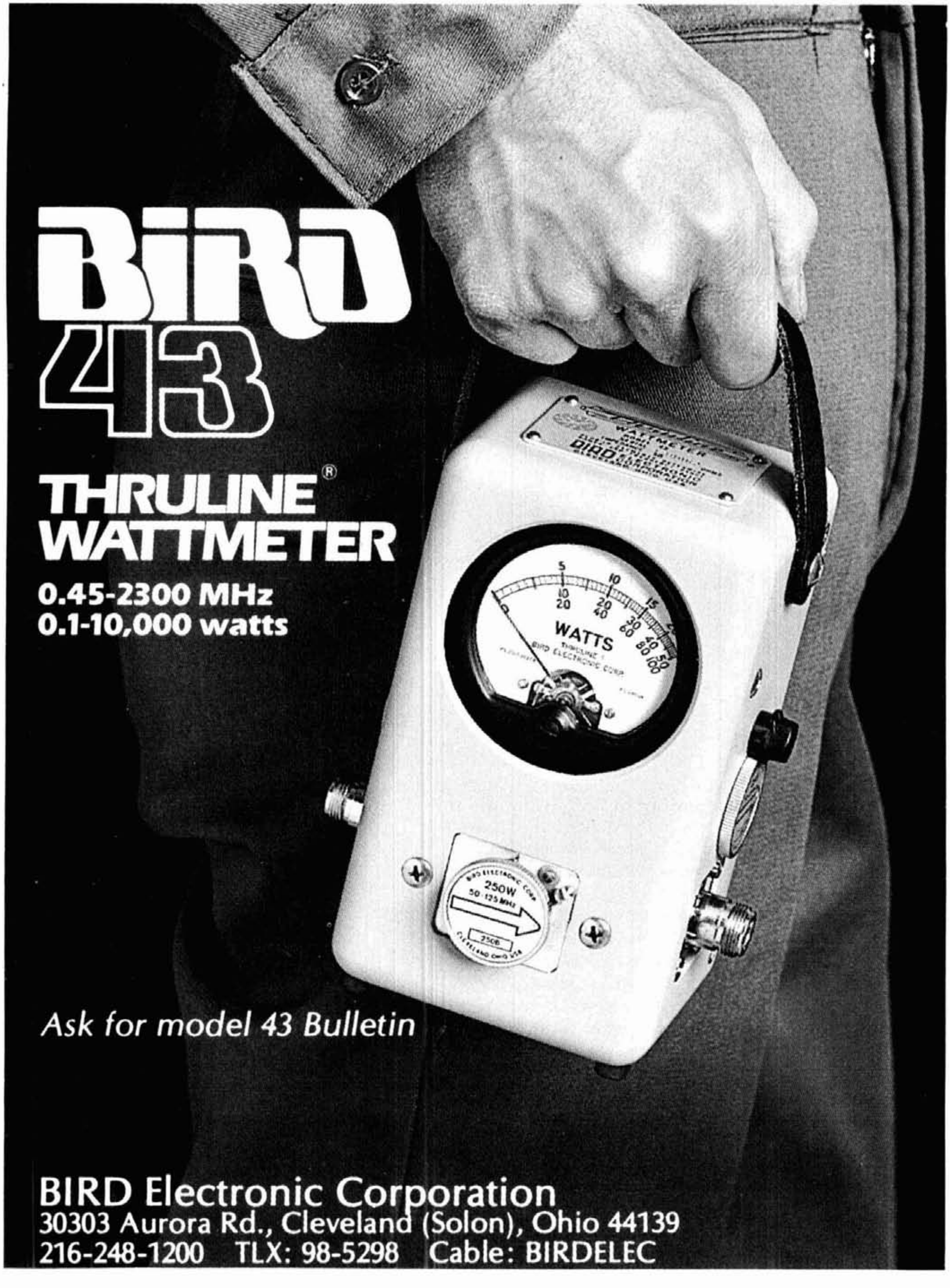
NOTES:
1. CLEAR DELAY LINE AT BEGINNING OF EACH HOP FRAME (STORE ALL ZEROS).
2. CLOCK OUTPUT SIGNAL INTO DELAY LINE.
3. WHEN OUTPUT SIGNAL REACHES FIRST DELAY LINE TAP, WEIGHTS W_1 (AMPLITUDE AND PHASE) BECOME EFFECTIVE.
4. SUBTRACT MULTIPATH COMPONENT FROM INPUT SIGNAL.
5. REPEAT FOR SECOND MULTIPATH COMPONENT USING WEIGHTS W_2.

fig. 7. Time-gate feedback equalizer.

Time-gated feedback equalizer. A time-gated feedback equalizer is shown in **fig. 7**. Its structure is very similar to that of any other feedback equalizer. It has an input signal, a differencing circuit, and a weighting network. The difference is supplied as an output signal to the data demodulator. The delay line is a multilevel shift register equal in length to the hop frame. The tap spacings, corresponding to minimum resolvable multipath delay, are equal to the signal sampling interval, which is considerably smaller than one bit in duration.

The delay line in **fig. 7** is tapped in two places. Two multiplying weights, W1 and W2 are shown; these are complex and represent both amplitude and phase weights. The resulting outputs are summed and fed back to the input differencing circuit. The peak clipper is incorporated to prevent positive feedback for certain data sets at high multipath levels.

Time-gating the feedback equalizer is achieved by clearing or resetting the delay line to zero at the beginning of each hop frame. Consequently, when a new hop frame starts, there is no feedback. In fact, there will be no feedback until the first samples reach the first tap. At that point, feedback of the appropriate amplitude and phase will begin cancelling the first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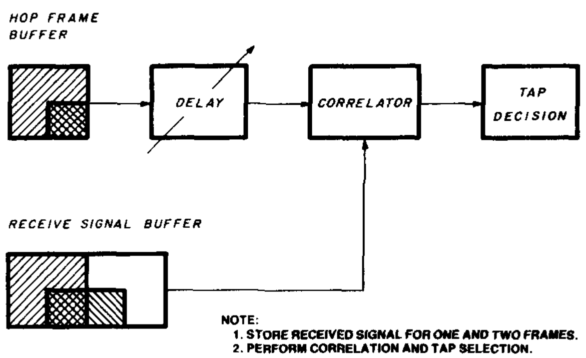

fig. 8. Correlation process.

multipath component. (A second multipath component will not become bothersome until the output signal reaches the second tap with amplitude and phase weight W2. Up to that time the finite values of amplitude and phase weights of weight W2 will have been multiplied by the zeros advancing in the delay line and the tap will not have been effective.) After that time, feedback cancels the second component, and so on. The impact of a large number of taps is that additional time is required to optimize the amplitude and phase weights. It is reasonably assumed from HF path predictions that only three to five multipath components will significantly affect a signal. The hardware necessary to process this data is available right now.

Tap locator. The second function in the multipath processor is the tap locator. The receiver provides two signals for use in an auto-correlation function which determines multipath tap locations. Once tap locations are known, the tap weights for the filtering function can be adjusted. This correlation is illustrated in **figs. 8** and **9**. The original transmitted frame is transferred to two buffers. The contents of the two buffers are then fed to a correlator.

The output of the correlator is examined for correlation peaks above a predetermined threshold, which indicates that a multipath component is present at that particular value of delay. This auto-correlation process determines the multipath delay without depending on any external timing. It is not affected by timing uncertainty or by jitter from one frequency hop to another.

Tap weight calculator. The third major function in the multipath processor is the tap weight calculator, which consists of a channel quality measurement unit and a tap weight programmer. The tap weight programmer alters the amplitude and phase of the tap weights in response to instantaneous channel quality measurement. Because hop frames are buffered, the quality of the channel can be measured repetitively as the equalizer converges and used to control the tap

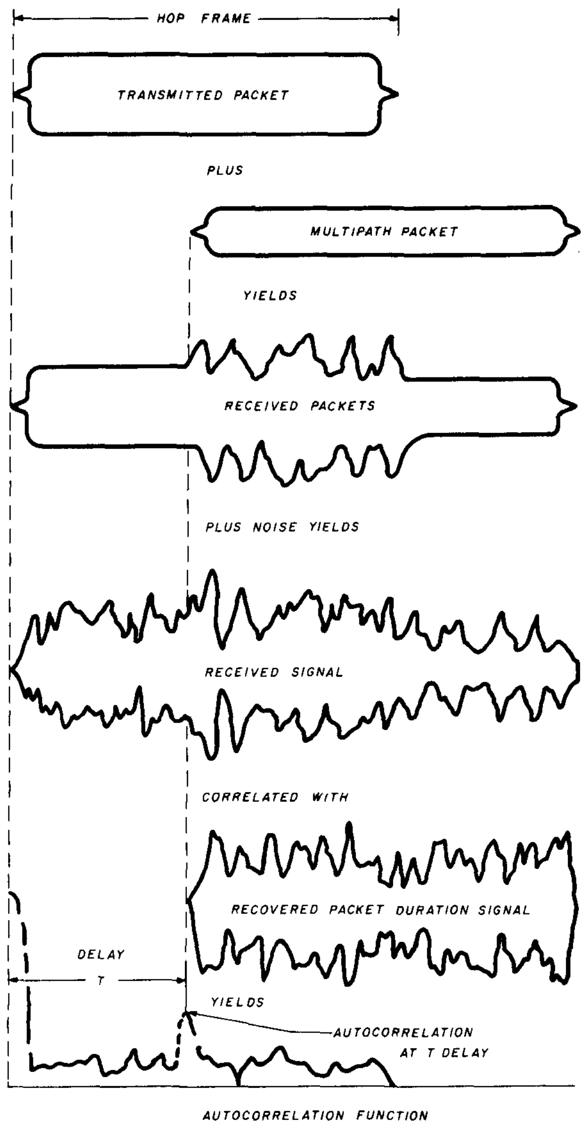

fig. 9. Time relationships in simulated signals.

weights in a closed loop fashion. The quality measurement is based on an RCA-patented quality monitoring technique particularly applicable to a variety of angle modulated signals. It does not require any special training signals, but instead operates directly on the data signal.

The principle behind the RCA technique is illustrated in **fig. 10**. Spectral analysis is performed at the output of the power law device (i.e., a frequency doubler or quadrupler). The RMS value of all the distortion products (that is, all spectral lines except the desired carriers) is *directly dependent* on the multipath amplitude and phase and are independent of the data contents. Therefore, no training signal is required to determine the channel quality.

When a biphase PSK signal, for example, is received, the frequency doppler will change a 180 degree phase shift to 360 degrees. The data now appears as 0 degrees or 360 degrees, which are equivalent angles. The data modulation has effectively been stripped off and converted to a carrier component; this is precisely how many coherent receivers track phase modulated data. However, the carrier component may not be the only spectral energy that remains after doubling. Any distortion products, including the intersymbol interference produced by the multipath delay, produce modulation sidebands. *A straightforward measurement of carrier to sideband energy is then used to determine the amount of multipath.*

When QPSK signals are transmitted, the standard carrier recovery technique in coherent receivers quadruples the signal, thereby eliminating the data modulation leaving only an unmodulated carrier signal. Tracking is then accomplished on what has essentially become a reconstituted carrier.

NOTES:
1. DATA MODULATION IS REMOVED FROM THE BPSK SIGNAL WHEN DOUBLED, THE QPSK SIGNAL WHEN QUADRUPLED, AND FROM THE MSK WHEN DOUBLED.
2. CARRIER, DATA CLOCK AND DISTORTION PRODUCTS REMAIN.
3. DISTORTION PRODUCTS (RMS VALUE) ARE DIRECTLY DEPENDENT ON MULTIPATH AMPLITUDE AND PHASE.
4. BASED ON RCA PATENT 4317206, (ON LINE QUALITY MONITORING).

fig. 10. Distortion measurement approach.

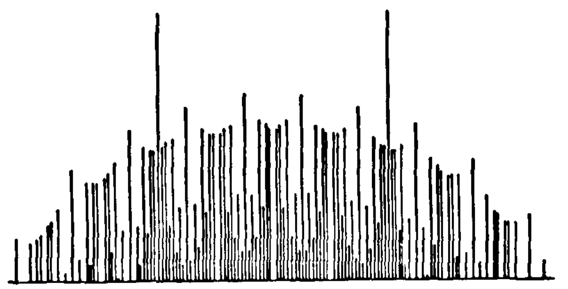

fig. 11. Spectral density MSK after squaring.

When MSK signals are sent, the standard approach to carrier extraction is to double the frequency in the squaring circuit to produce two carriers separated by the clock frequency. This is illustrated for continuous phase FSK in **fig. 11**. In the case of MSK, after squaring, one half the power is distributed between the two carriers. It is standard practice to phaselock a loop to one of the carriers or use the data clock to collapse both carriers into a single component and then phaselock for coherent demodulation of the data. Channel distortion results in less energy in the carrier components and more in the sidebands. Again, the ratio of carrier energy to sideband energy provides an estimate of the channel quality measurement.

The tap weight calculator uses the Channel Quality (CQ) measurements to progressively reduce the uncertainty of the phase and amplitude weights applied to each delay line tap. The CQ is the ratio of the carrier power to the RMS value of all other frequency components after frequency doubling. In the case of MSK, the power in both of the reconstructed carriers is added to achieve a higher CQ and therefore a better measurement. As the tap weights are changed, the CQ value will change dramatically and reach a maximum when best equalization has been achieved.

The simulation example shown in **fig. 12** illustrates rapid tap weight convergence for an assumed multipath amplitude of 0.9 and phase angle of 169 degrees. While seven complete iterations are shown to illustrate the sequence, after the first iteration the phase error is only 11 degrees and the amplitude error 0.4 volts. For purposes of achieving minimum bit error rate performance, this represents complete equalization. Since there is no significant change in the results at the end of the second iteration, the convergence procedure would be halted. If the CQ is high enough, depending on noise and the remaining multipath level, the convergence procedure may actually be halted after the first iteration.

fig. 12. Simulation example of equalizer convergence.

data parameters:		multipath parameters:		equalizer parameters:	
sequence	1	delay (bits)	3.00	tap location	3.000
frame length	64 bits	amplitude	0.90	threshold	1.000
samples/bit	8	phase	169.00	input SNR (dB)	5

iteration	amplitude	phase	amplitude error	phase error	channel quality	data errors
0	0.000	0.0	−0.900	−169.0	25.6	9
1	0.500	180.0	−0.400	11.0	16.2	0
2	0.500	180.0	−0.400	11.0	16.2	0
3	0.500	157.5	−0.400	−11.5	16.4	0
4	0.830	157.5	−0.070	−11.5	212.3	0
5	0.995	157.5	0.095	−11.5	191.5	0

Acquisition process. Because the equalizer does not require training signals, a frequency hopping receiver can achieve initial acquisition and convergence of its equalizer without requiring a cooperative transmission. **Figure 13** shows a transmitter hopping at a normal rate. The receiver is trying to acquire and determine the tap locations, but this is not apparent to an outside observer. Initial acquisition is accomplished by letting the receiver dwell twice as long on each frequency compared to the transmitter. This means that the receiver will miss alternate frequencies; but, at the same time, it will include in its output data the desired signal as well as the multipath signal. As shown in **fig. 13**, the transmitter hops through six frequencies, but the receiver only sees three of those frequencies. Once acquired, the receiver commences hopping at the same rate as the transmitter. Adding a second receiver can provide coverage of the missed alternate frequencies. The additional expense of a second receiver and a frequency synthesizer are not warranted in most applications.

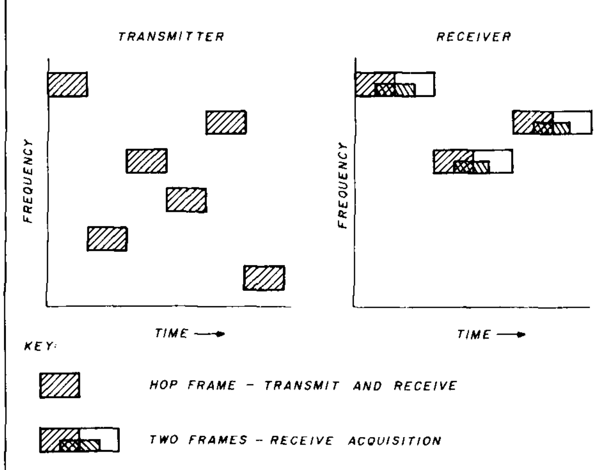

fig. 13. Acquisition frequency hop patterns.

multipath equalizer summary

When frequency hopping is employed to achieve protection from jamming, the problems posed by multipath distortion on HF radio links increase. The novel approach discussed here illustrates what can be done with digital processing. It is well matched to HF hopping systems and supports a very robust form of MSK modulation with efficient error correction coding and is readily implemented with today's technology.

digital SSB tuning

Another example of the successful application of digital techniques to HF radio is the problem of automating SSB tuning. **Figure 14** illustrates a block diagram of a typical digital HF radio. The digital sec-

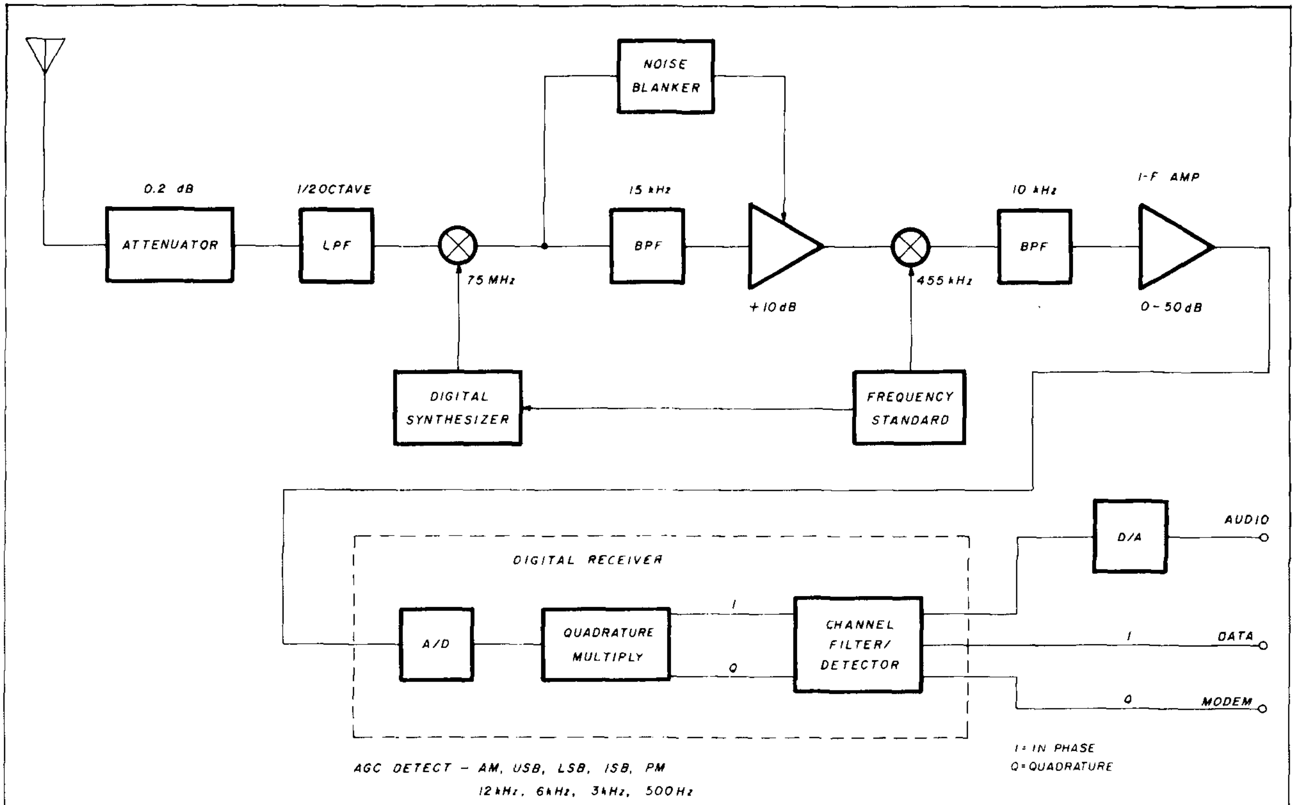

fig. 14. Block diagram of typical digital HF radio.

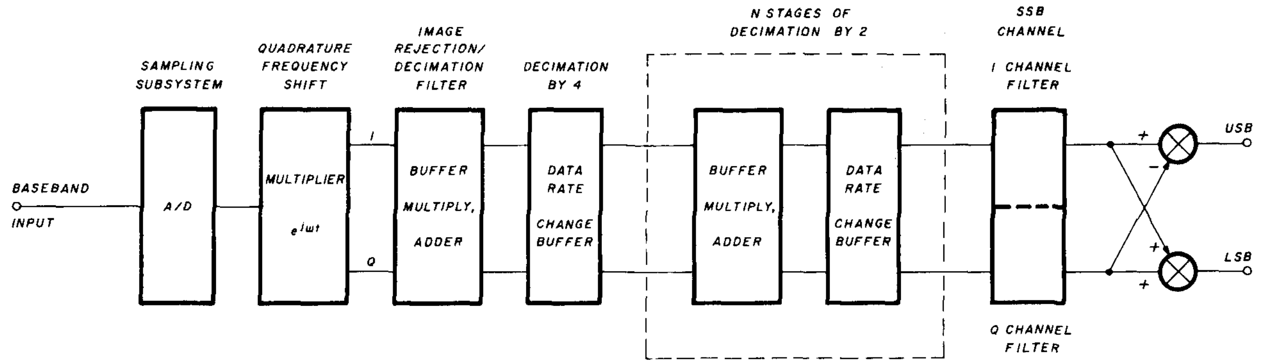

fig. 15. Digital receiver processing (SSB model).

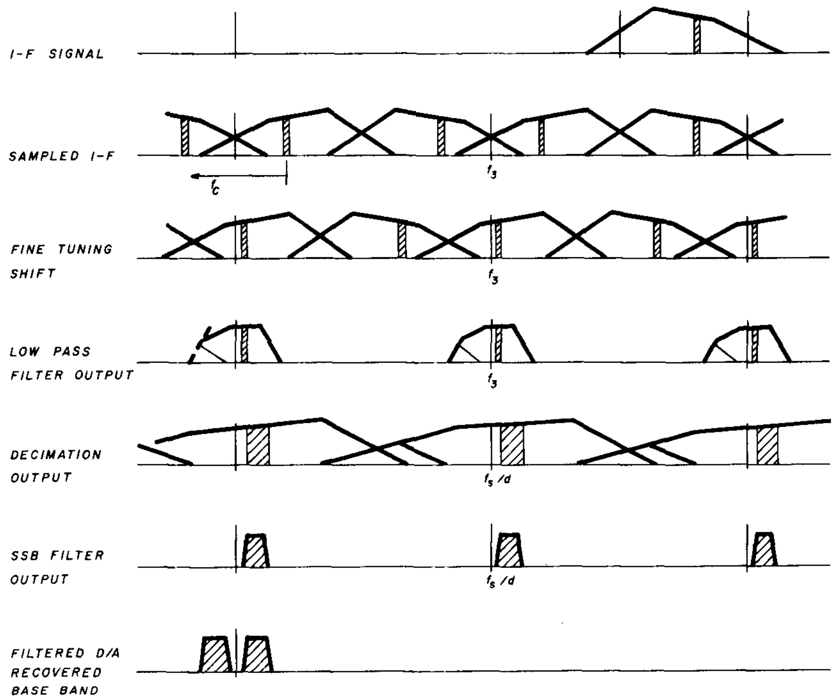

fig. 16. Digital receiver process spectra.

tion of the receiver is indicated within the dotted lines. **Figure 15** shows, again in block diagram format, how SSB detection is performed. **Figure 16** illustrates what signals occur during processing.

The method shown in **fig. 17** can be used to develop a software solution to automatic SSB tuning. The voice bandwidth signal is transformed into a voice power spectrum, which is analyzed to detect harmonic relations and complex correlations of the voice spectrum. Digital signal processing then determines the amount of frequency offset relative to a hypothetical center frequency, and a digital signal then tunes the frequency synthesizer to correct any offset. This "center frequency" can then be fed to a processor and stored together with the demodulated data.

Alternatively, this process can be used to regenerate the suppressed carrier. If we look at the lowest signal line in **fig. 18**, we see what appears to be pure noise. If enough samples with sufficient bit resolution are collected, it is possible not only to discover the actual suppressed carrier, but also to find 60 cycle hum sidebands. Spectral detail is useful in uniquely identifying specific pieces of radio equipment. Such techniques are known as "fingerprinting."

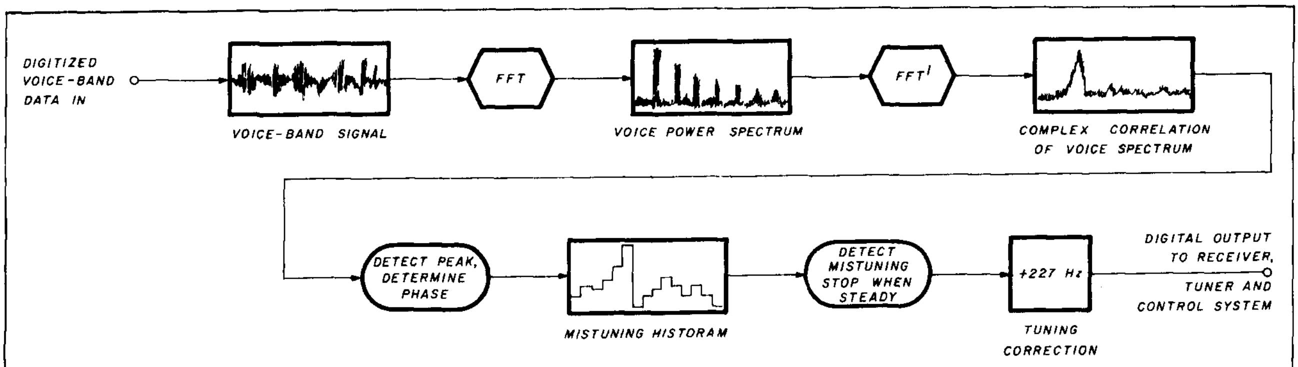

fig. 17. Automatic SSB tuning algorithm overview.

digital waveform generation

Generating analog waveforms from digital signals is another interesting application of this new technology, which allows straight-forward implementation of RCA's "Ampliphase"® system, a modulation scheme used in high-power AM broadcast transmitters.

In its analog form, Ampliphase combines outputs of two individually phased modulated carriers to produce a single AM output signal. Using efficient, nonlinear solid-state amplifiers, it is possible to generate many different forms of modulations — rather than only AM — simply selecting the desired mathematical algorithm. A list of modulations appears in **fig. 19. Figure 20A** shows a digital arrangement that generates AM signals; for explanation, **fig. 20B** shows the amplitude and phase relationships between the two channels. Depending on the phase shift, different modulation and sideband levels are achieved.

The flexibility of Ampliphase is shown in **fig. 21**, which illustrates a method of digitally generating four-channel SSB suppressed carrier modulation. **Figure 22** shows the digital implementation of constant envelope independent SSB suppressed carrier generation. In each case, the hardware stays the same — but a different algorithm is applied.

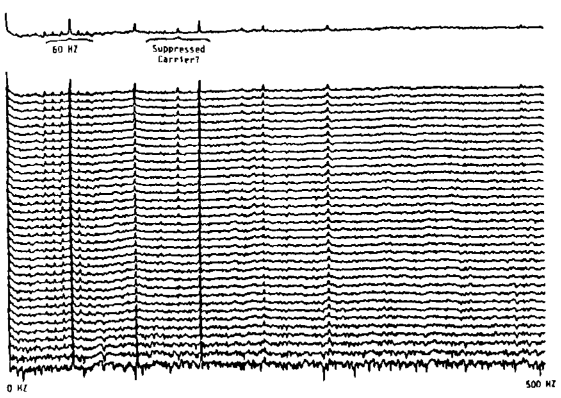

fig. 18. Exponentially-averaged SSB spectra ($\alpha = 0.9$) **mistuned up 100-200 Hz.**

digital filter implementation

Perhaps the most familiar application of digital

fig. 19. RCA's Ampliphase® system generates many different forms of modulation.

modulation:	$x(nT) = Re\,[Z(nT)\ EXP\ (j\omega nT)]$
DSB-AM	$Z(nT) = \frac{1}{2} + \frac{m}{2}\, y(nT)$
DSB-SC	$Z(nT) = y(nT)$
SSB-SC	$Z(nT) = y(nT) \pm j\,\hat{y}\,(nT)$
PM	$Z(nT) = EXP\ [jK\ y(nT)]$
FM	$Z(nT) = EXP\ [jK\ \Sigma\ y(nT)]$
demodulation:	
DSB-AM	*envelope detector* $y(nT) = L + 0.3\ S \approx \sqrt{I^2(nT) + Q^2(nT)}$ *synchronous detector* $y(nT) = Re\ [(I(nT) + jQ(nT))\ EXP\ (j\omega_o nT - \theta_o(nT))]$
DSB-SC	*same synchronous detector as DSB-AM*
SSB-SC	*product detector* $y(nT) = Re\ [I(nT) + jQ(nT))\ EXP\ (j\omega_c nT)]$
PM	*angle detector* $y(nT) = \theta(nT) = ARCTAN\ (Q(nT)/I(nT))$
FM	*discriminator* $y(nT) = \theta(nT) - \theta((n-1)T) \approx \tilde{\theta}$

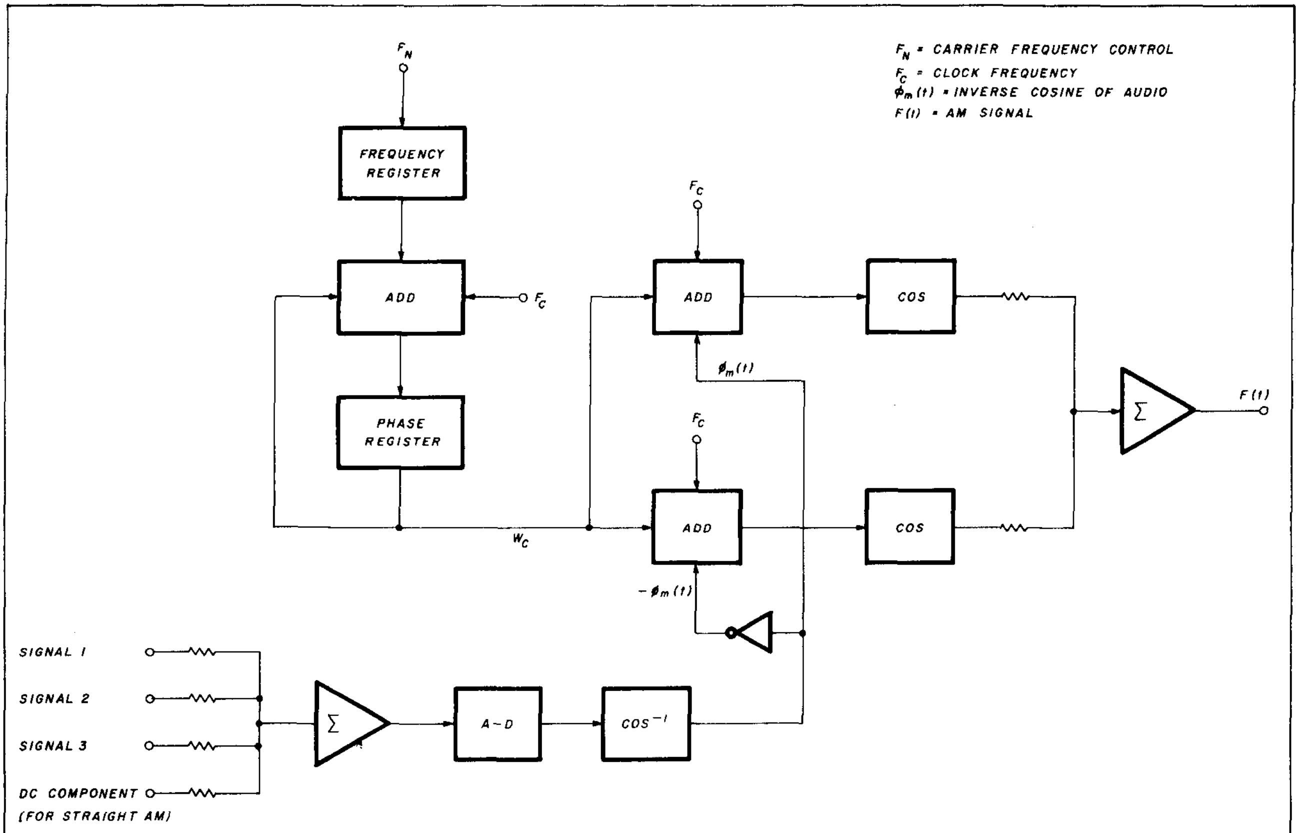

fig. 20A. Ampliphase® AM simulator block diagram.

techniques today is the use of digital filters to replace analog components in audio sections of radios. Often, the selectivity of a digital receiver is determined by the quality of the implementation of its digital filters. These digital filters, typically non-recursive finite impulse response filters, are characterized by no more than sufficient performance, although this deficiency is offset by the simple hardware requirements for implementation. The impulse response of these filters can be computed by using the Parks-McClellan FIR design program. **Figure 23** shows predicted performance data for such filters. The filter algorithms are usually implemented entirely in microcomputer software when time allows, or, alternately, using high speed arithmetic processors to enhance throughput. Special arithmetic logic units can also be employed at the output of a filter to generate either the inverse or the magnitude of the output value, depending upon the filter requirements. Finally, the filter output can be either numeric (digital) or analog via a D-to-A converter.

Figure 24 illustrates the predicted amplitude response for a typical FIR filter.

frequency synthesizers

Digital frequency synthesizers, common in today's radios, typically have slow frequency switching speeds

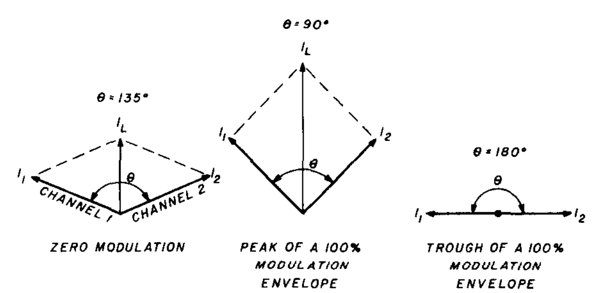

fig. 20B. Vector representation of phases between channels.

and variable noise performance. But modern frequency synthesizers can be built with almost infinite frequency resolution and very fast switching speed. (The typical design trade-off is noise sideband performance vs. switching speed.) Analog PLL frequency synthesizers are typically limited in their switching speed by the loop bandwidth of the system and hence offer only limited resolution at high speed.

The best approach to building fast frequency synthesizers with fine resolution is to combine either a wideband analog loop with a digital direct frequency synthesizer or a wideband analog loop with a fractional

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The fact that the Computer Patch Interface unit by Advanced Electronic Applications, Inc. is known as the best value on the market is no accident. The CP-1 was designed by Al Chandler, K6RFK (PHD-E.E.), an active RTTY user since 1963.

Given a cost per unit budget for the CP-1, Al designed as much performance as possible into the Computer Patch, including a unique new tuning indicator, referred to by one of our customers as the "Dead Eye Dick" tuning indicator. This indicator is ideal for RTTY and CW, in that it is both fast to tune and (within 10 Hz) as accurate as scope tuning. It also performs under poor signal to noise conditions in which other indicators provide no useful data.

Al's variable shift tuning was designed to move the space filter center frequency from 2225 Hz to 3125 Hz without changing the bandwidth (by varying the Q of the filter). All this is accomplished using a precision ganged potentiometer to assure proper tracking of the multiple filter stages. We could have used a pot costing a tenth as much by simply using a two-pole filter design, but we feel the advantage of a sharper filter reduces the noise bandwidth significantly and allows the variable shift control to be used like passband tuning for extra elimination of adjacent channel interference.

Some manufacturers are concerned that amateurs might try calibrating their own equipment and, therefore, have used non-adjustable components, which results in sub-optimal performance. Although more costly, trimpots used in AEA equipment allow factory adjustment for performance to design specifications. Competently designed active filter circuits need not be adjusted after leaving the factory; however, for specialized use the owner can easily change filter parameters.

Mindful of the fact that many of our customers are new to RTTY, Al made the CP-1 tuning as forgiving as possible, while providing the most critical operator a piece of equipment in which he could be proud. Even old "pro's" are surprised at the poor signal conditions under which the CP-1 will still provide good copy.

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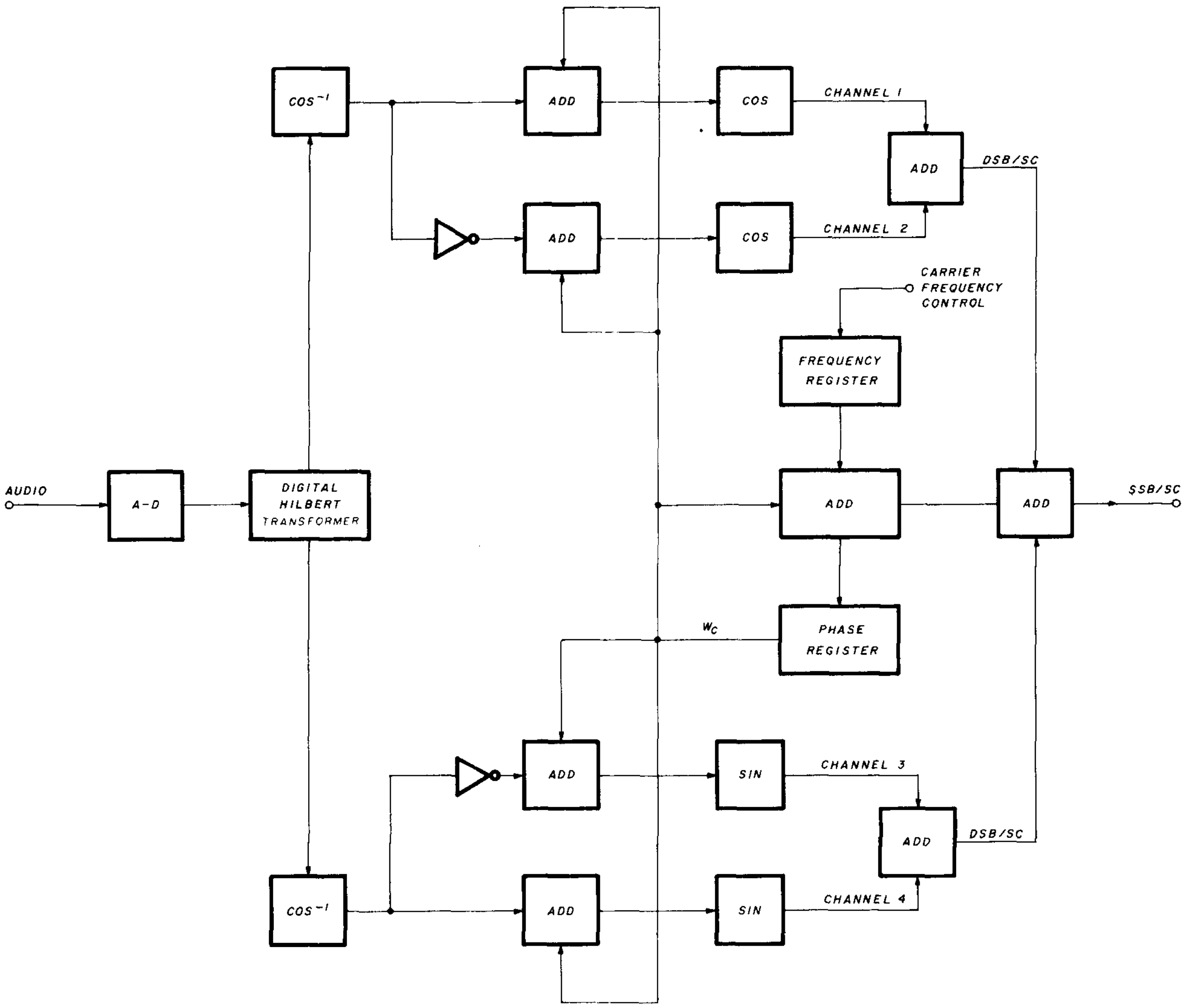

fig. 21. Four-channel SSB/SC simulator block diagram.

divider and synthesizer. **Figure 25** shows one implementation of the former. It is possible to get switching speeds of several hundred microseconds and excellent phase noise at the same time using this technique.

antenna coupler technology

One interesting item not often considered part of the radio is the antenna coupler. Antenna couplers now used provide good RF power transfer between the radios and the antenna. At a single frequency their reliability is not a serious problem. New couplers, with a capability for fast frequency hopping, require new technology to provide fast frequency changes and extended lifetimes — i.e., a predictable period of use without wear-out or breakdown. Because couplers must also be driven by the digital processing system "brain" of the radio, they require a digital interface and, usually, their own processor and training scheme as well.

available hardware

Antenna couplers presently used on aircraft adapt to the driving point impedance peculiarities of their respective antennas. Though tuning methods may differ in detail (depending on the matching network configuration), tuning always involves the use of motor-tuned and relay-switched reactive elements driven toward 50-ohm convergence by an error signal from magnitude and phase discriminators. Tuning accuracies are quite good (VSWR 1.3:1), and coupling efficiencies range between 40 and 85 percent depending on antenna type and frequency of operation. The disadvantage of this class of antenna coupler, however,

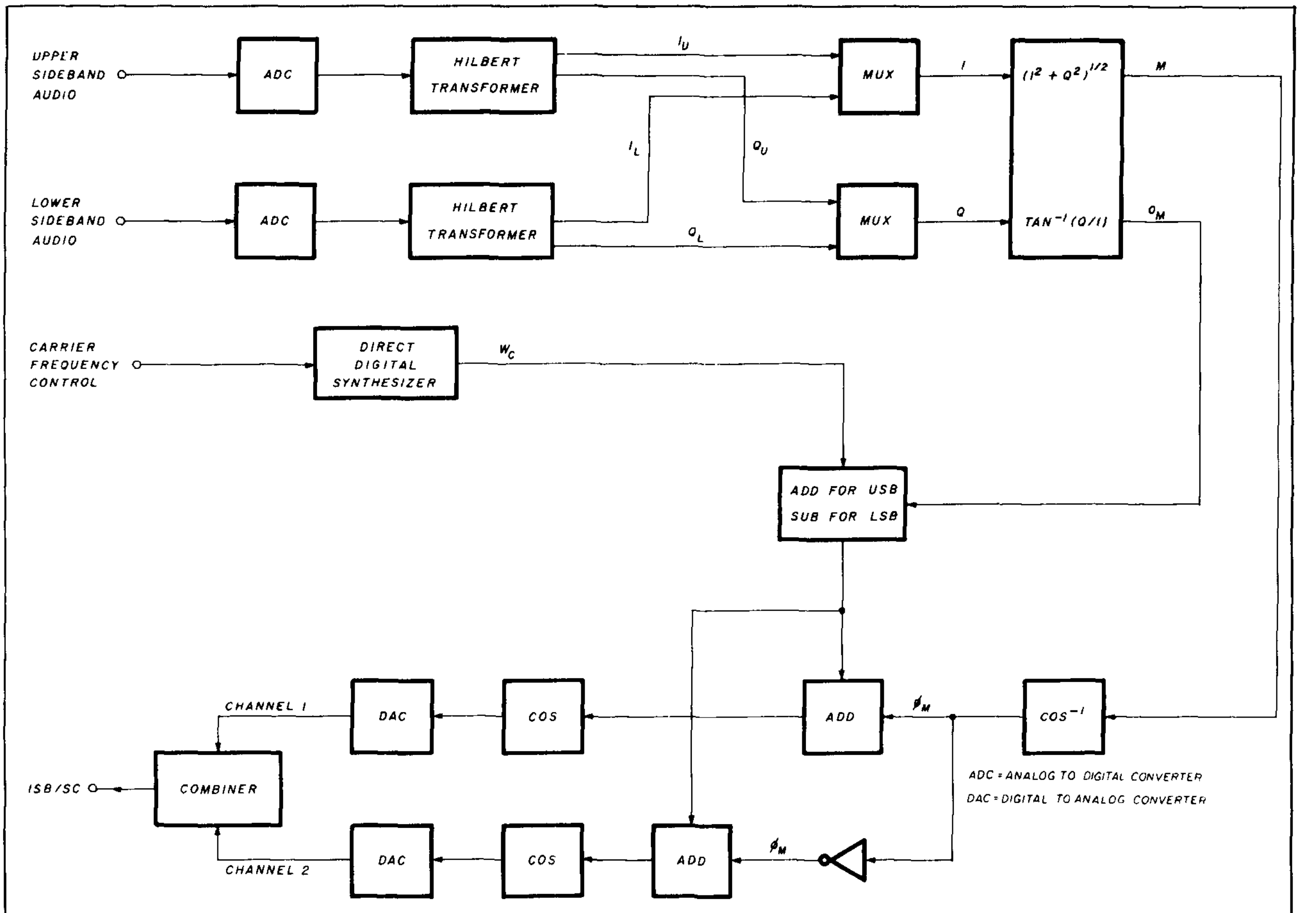

fig. 22. Digital constant envelope ISB/SC simulation block diagram.

fig. 23. Predicted performance data for digital filters in audio applications.

filter function	input sample rate	output sample rate	2-sided passband bandwidth	ultimate rejection	2-sided BW at ultimate rejection	shape factor	impulse response length (μsec)
FM receive	100 KSPS	50 KSPS	25 kHz	>80 dB	36 kHz	1.4:1	62
AM receive	100 KSPS	25 KSPS	12 kHz	>80 dB	17.2 kHz	1.4:1	124
AM receive	100 KSPS	12.5 KSPS	6 kHz	>80 dB	8.8 kHz	1.5:1	248
SSB receive	100 KSPS	12.5 KSPS	2.7 kHz	>80 dB	5.4 kHz	2:1	248

is relatively long tuning time (from 3 to 15 seconds), and a limited lifetime resulting from the failure of electromechanical drives and switches.

Advanced tuning elements. Given the insufficient durability of electromechanical components, the future of tuned couplers with both fast-tuning times and operational reliability will depend on the use of electronic tuning and solid-state devices. Of the solid-state devices that could be considered for this application, two have been given varying degrees of attention: the saturable reactor and the PIN diode.

A saturable reactor is an RF inductor wound on a suitable ferrite core whose permeability is varied by an *orthogonal or parallel DC excited magnetic field.* Inductance variations of 4:1 or greater have been obtained with this kind of device, which can be configured as a memory element by the inclusion of a permanent magnet bias. The main difficulties associated with the use of the saturable reactor are:

- Response time is slow due to excitation time constants.
- Large ferrite volumes are required to overcome the inherent nonlinearity and heat dissipation problems.

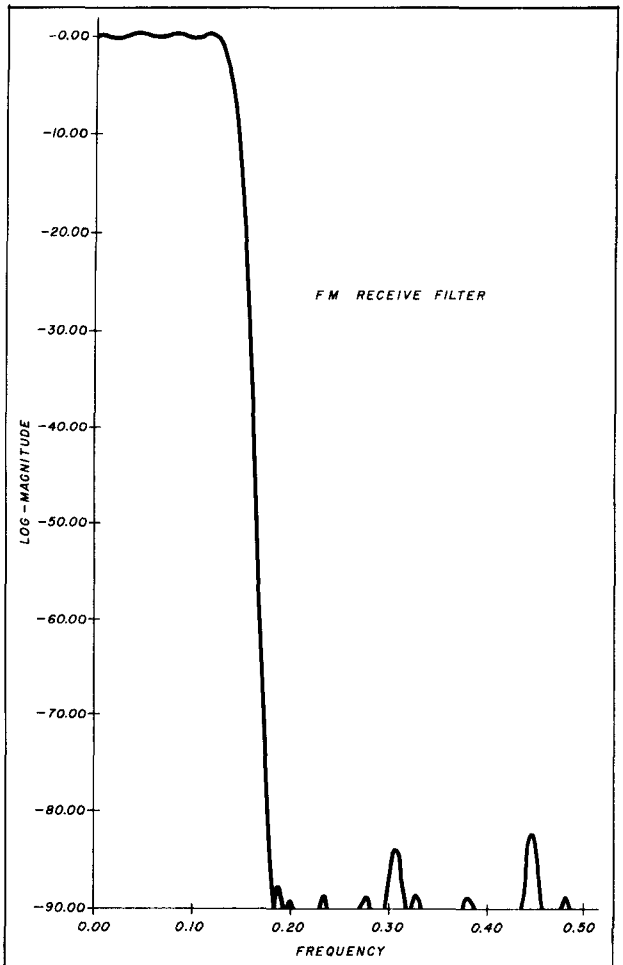

fig. 24. Predicted amplitude response for a typical FIR filter.

• The ferrite material is temperature-sensitive.

• Permanent (irreversible) structural changes can take place at some specific excitation levels.

The device may find application in coupler schemes where a limited degree of adaptive tuning over a narrow frequency range is desirable.

The PIN diode, a solid-state device, offers many interesting possiblities for switching if optimized for operation in the HF frequency range. The keystone of advanced HF Coupler design at RCA, its use permits reactive elements of the matching networks to be varied by switching discrete values of inductors and capacitors in a digital manner. Such a configuration is readily adaptable to a microprocessor control interface. The diode parameters most critical for this application are forward bias resistance, reverse-bias breakdown voltage, minority carrier lifetime, and power-handling capability.

The reverse breakdown voltage of the PIN diodes in the HF frequency range must be high enough to withstand the peak RF voltages developed at the high Q end of capacitive antennas. As an example, an aircraft probe antenna with a Q of 500 at 2 MHz, when matched may develop an RF voltage at the base of 15 kV or more with 1 kW input, depending on the losses encountered in the matching network. Thus a PIN diode, or rather a PIN diode package, must be able to withstand at least half that voltage.

Agile antenna couplers using PIN diodes which quickly respond by microprocessor control can meet today's needs of high hopping rates. Frequency

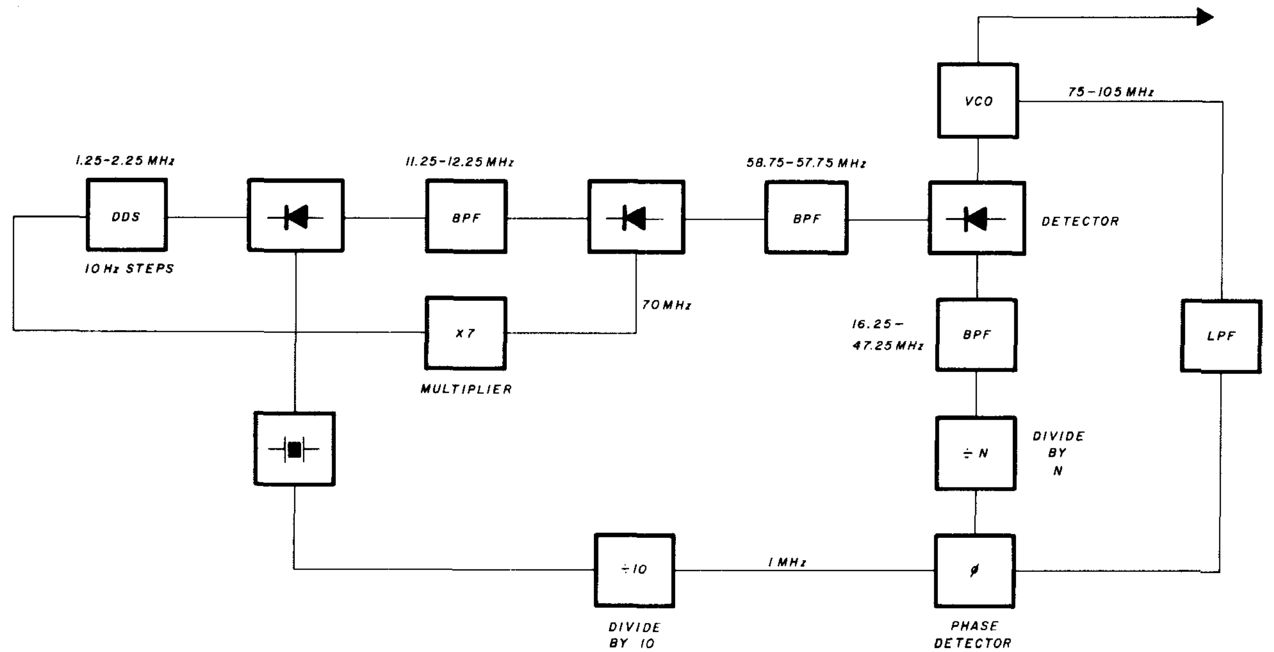

fig. 25. High-speed synthesizer.


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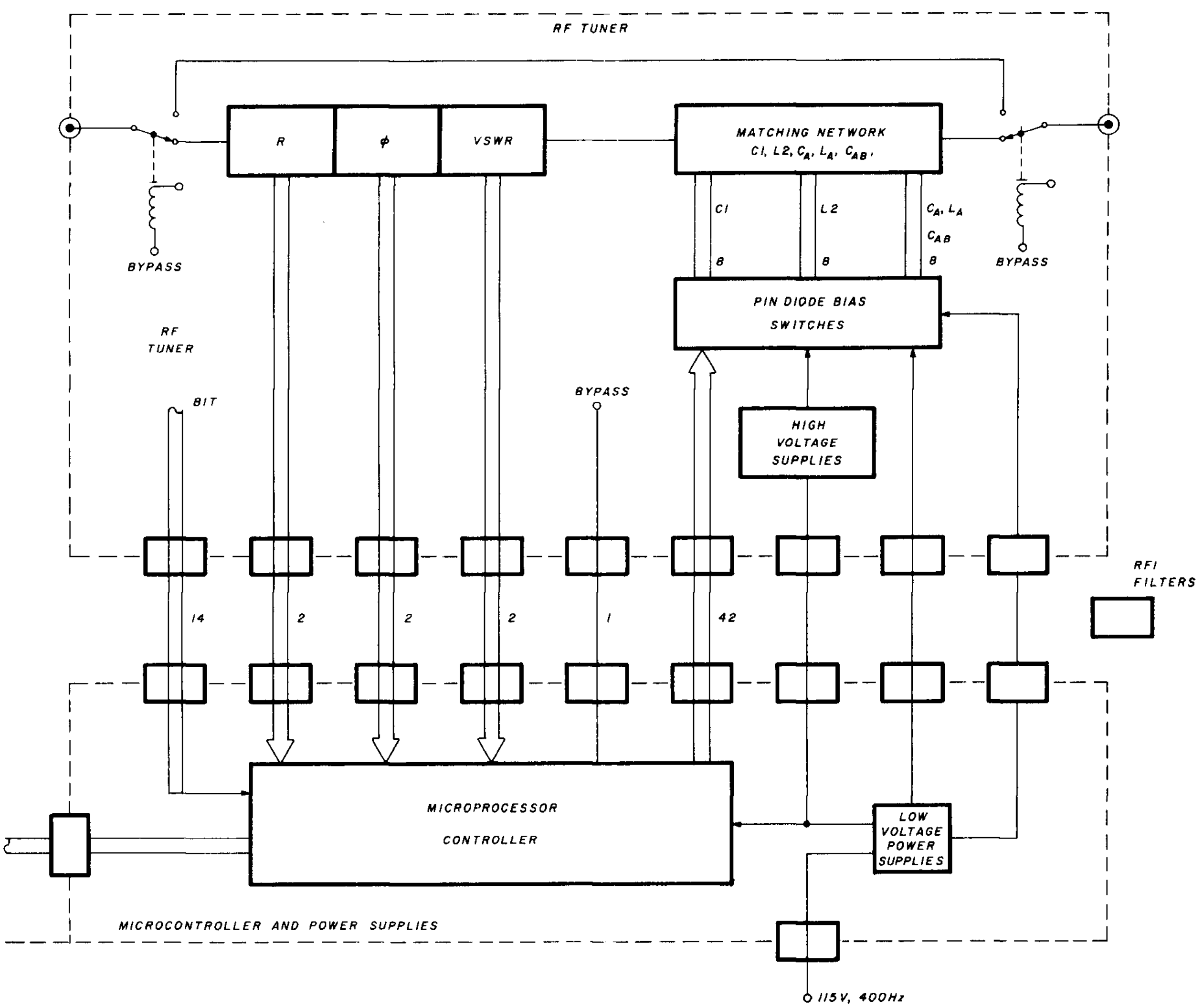

fig. 26. HF agile antenna coupler system block diagram.

changes can be made in microseconds, thus obsoleting older techniques. Advantages of agile antenna couplers with digital processing are high speed, reliability, and adaptability. Through self-test techniques and learning and storing of new antenna characteristics, the coupler can adapt to changes and continue to implement the best impedance match, resulting in good power transfer to the radiating antenna. **Figure 26** shows the system block diagram of the HF agile antenna coupler. A switching speed in the order of 200 microseconds is possible.

summary

This article describes requirements and techniques used in the design of modern digital HF radio which fulfill the needs of advanced RF communication systems. Most of the advanced techniques were developed to satisfy digital data transmission and frequency hopping requirements. Digital implementation of many of the system functions, required for both hopping and non-hopping modes, avoids some of the problems analog linear devices introduce. Digital techniques may ultimately prove to be more cost effective as well. The use of microprocessing and signal processing devices provides greater flexibility of the HF transceiver and allows its use in an integrated system.

acknowledgement

Much of the work described in this article derives from the efforts of the Radio Systems' Engineering section at RCA's Government Communications Systems, Camden, New Jersey, without whose help this article could not have been completed. Among the major contributors to this article were Robert M. Lisowski, John B. McMackin, David A. Miller, Edward

J. Nossen, David P. O'Rourke, and Charles K. Vickers. Additional thanks go to Mr. Miller for very carefully proofing the entire manuscript and supplementing several of the more esoteric terms with examples.

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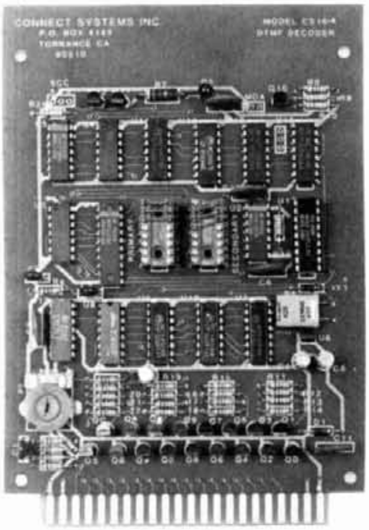

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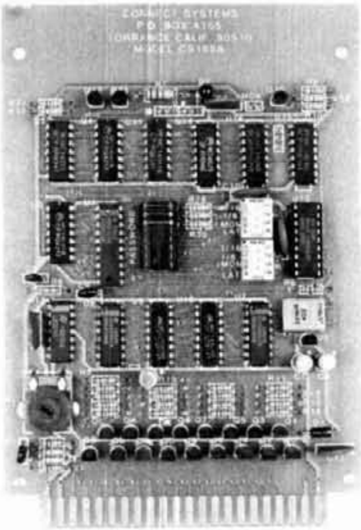

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4.	8 MOMENTARY	and	1 OF 8 SELECT
5.	1 OF 8 SELECT	and	8 MOMENTARY
6.	1 OF 8 SELECT	and	1 OF 8 SELECT
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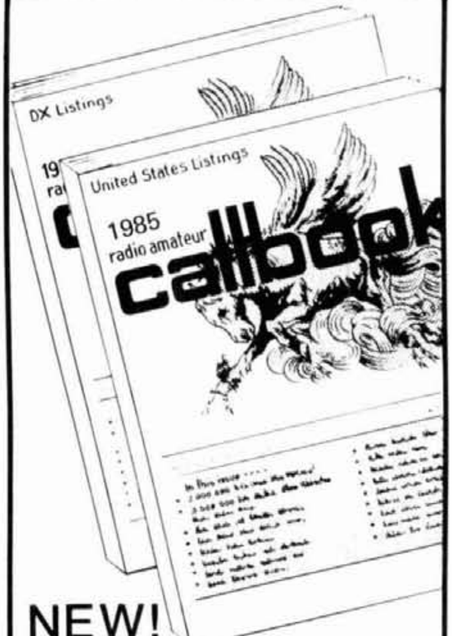

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a carrier-operated relay for VHF amplifiers

simple switch operates reliably over wide temperature range

The availability of low-cost hand-held VHF transceivers has increased the demand for external power amplifiers to overcome the inherent power limitations of these radios. In particular, many Amateurs wish to adapt their hand-held units to mobile use and take advantage of the abundance of low voltage DC power available in this application with an add-on amplifier.

Many application notes detailing the construction of such amplifiers are available.[1] A typical design appeared in past editions of the Motorola *RF Data Manual* as EB92A and is built around the MHW-252 hybrid module. The possibility of obtaining 20 dB or more of gain in a single module prompted us to build these amplifiers with the hope of developing 25 watts from the 200 mW provided by an ICOM 2A operating on low power.

The construction of these amplifiers proved to be a near-total disaster for several reasons, and we are by no means encouraging others to follow in our footsteps as far as the amplifier design is concerned. In fact, the latest edition of Motorola's *RF Data Manual* does not list the MHW-252 at all. However, one significant improvement was made in the Motorola design which will undoubtedly prove to be very useful in future power amplifier projects.

The major improvement concerns the COR circuit. In the original Motorola design, shown in **fig. 1**, it can be seen that a 5-pF capacitor is used to couple RF energy to a transistor that drives a mechanical relay. At 146 MHz, 5 pF represents 220 ohms of capacitive reactance. Because the transistor switch is forward biased somewhat during reception, the 5 pF capacitive reactance is primarily the input impedance to the circuit, and thus seriously affects the receiver signal path. In addition, the transistor circuit was found to be extremely unreliable because of base-emitter threshold changes with temperature, especially bothersome in mobile operation during cold mornings. In fact, the slight variation with frequency of the power output of an IC-2A became very noticeable on some mornings, yielding operation on only part of the band.

some problems occurred

Some history of our construction experiences offers comic relief value and should therefore be expounded for completeness. We built three of these amplifiers using the suggested single-sided PC board. As described in the application note, the circuit purportedly used the lead inductance of the relay and printed circuit traces as elements of a harmonic output filter. Measured with a Bird wattmeter, each of our separately constructed units produced a whopping 17-18 watts of output power. This was about 10 watts lower than expected. After a call to the manufacturer we were certain we had erred somehow. The manufacturers' representatives verified the published performance specifications, which we were obviously not meeting. In order to rectify matters, we delved into the theory of the LPF and tried different capacitor values and types. In spite of these changes, output power remained below 20 watts even with different ICOMs, each of which provided the required 200 mW into a 50-ohm load. Suspecting some type of mismatch condition either at the input or output, we proceeded to construct new boards using 50-ohm microstrip for RF connections. Instantly, 30 watts appeared at the Bird wattmeter load upon test. We concluded that the relay lead inductance was itself not enough to mismatch the input and output circuits but that the

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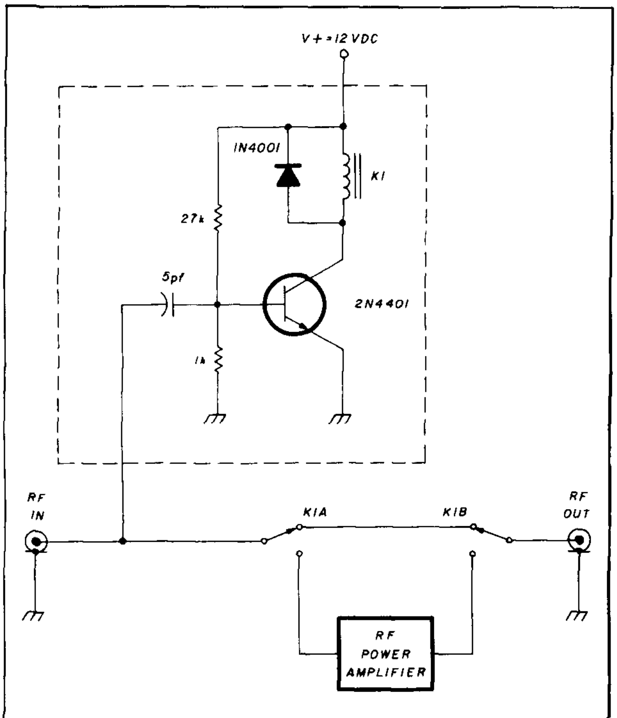

fig. 1. System diagram for 2-meter power amplifier described in Motorola Application Note EB92A. Carrier operated relay shown within dotted line.

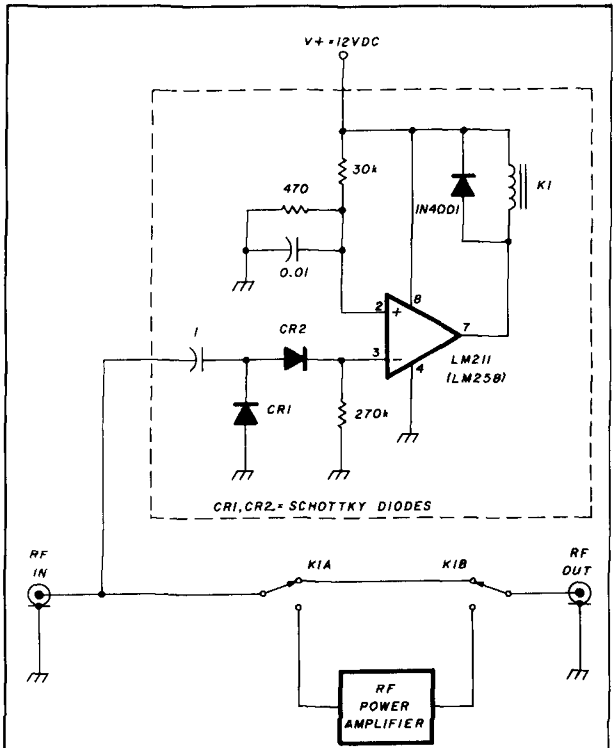

fig. 2. Modified carrier operated relay to provide less receiver loading and operation at power inputs of 100 mW to 10 watts.

circuit traces were. The LPF was also deleted. Other harmonic filters were investigated, and although power loss was not significant, they were not used for the balance of our tests.

In addition to these problems, discussions with the manufacturer provided critical information regarding the input attenuator network. We were told that deleting the network would result in damage to either the ICOM or amplifier because of the adverse interaction between the amplifier and the ICOM resulting from improper source impedance at the ICOM output (on low power only). A 1 to 1.5 dB minimum amount of attenuation is required for the network to reduce the interaction. Fortunately we found this problem by accident before these discussions, but the ICOMs were not damaged. (Our MHW-252 modules were replaced at no charge.)

The process of changing circuit boards to the microstrip type required some work and a bit of magic which we did not possess. In the absence of magic or luck we noted that the MHW-252 leads promptly detach from the module substrate after one soldering/desoldering operation. We assumed we had defective modules, but units from different sources behaved similarly. A honed soldering iron tip and a steady hand allowed us to reattach the leads to the substrate. (We might add that because our construction and design experience is extensive, we cannot wholly be blamed for the aforementioned problems.)

While waiting to overcome some of these difficulties with the circuit, we set out to improve the COR circuit. We decided that the ideal COR circuit should have the following properties:

- as little loading of the receiver signal path as possible (1 pF at 146 MHz)
- reliable operation down to 100 mW
- *simple and inexpensive*
- ultra-reliable with respect to temperature

COR description

A quick calculation shows that the peak-to-peak voltage of a 100-mW RF signal on 50-ohm line is about 6 volts. This should, in a properly designed circuit, be more than sufficient to guarantee reliable operation. The trick is to make a suitably high impedance switch so as to allow a 1-pF capacitor to provide the coupling. The capacitive reactance of a 1-pF capacitor is approximately 1100 ohms, which presents little loading to the receiver even if the switch input is highly capacitive.

No problems were experienced with the configuration shown in **fig. 2**. CR1 and CR2 are inexpensive Schottky diodes configured as a voltage doubler to extract the peak-to-peak voltage of the RF signal, aside from the voltage drops of the diodes. The bias built up by the action of the diodes keeps them operating primarily in reverse bias where their capacitance is typically 1 pF. This is entirely acceptable for VHF applications and is significantly better than that of low-cost junction diodes. Furthermore, the insignificant storage time of the hot-carrier diodes makes them much more efficient as an RF level detector.

circuit description

The circuit operates as follows. Application of RF power to the input provides a rectified signal at the diode doubler as described. The input impedance of the comparator (or op-amp) and bleed-off resistor is sufficiently high to allow a significant voltage to be developed at pin 2 compared to the bias applied at pin 3 (200 mV). The comparator changes its output state from high to low at this time. Relay K1 is actuated by the near ground potential assumed at output pin 1, provided the required relay current is within the sink capabilities of the comparator output stage. Upon release of input excitation, the voltage at pin 2 decays to a value sufficiently below the bias voltage at a rate determined by circuit capacitance and the 270-kilohm resistor. Subsequently, K1 is de-energized.

The comparator chosen was an LM211 (preferable) or an LM258 (a dual device). Either can sink the 25-30 mA required to operate the OMRON relay used in the Motorola design. In addition these ICs come packaged in a TO-5 metal can package which permitted the pins to be conveniently inserted into the appropriate holes already present in our circuit boards from the Motorola design: (Incidentally the old transistor circuit was removed and tested for 60 Hz response at 120 VAC.) Although double the component count is used compared to the suggested COR circuit of **fig. 1**, the new circuit easily fits on the circuit board and has been used in other power amplifiers.

The circuit, which met all of our requirements, has been in use for several months. If anything, it is too sensitive, as nearby mobile transmitters will sometimes cause the relay to chatter furiously. We noted this effect only at the Dayton Hamfest and consider the inconvenience of minimal consequence.

references

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a PSK telemetry demodulator for OSCAR 10

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OSCAR-10, launched by Ariane rocket on June 16, 1983, is the latest transponding satellite for Radio Amateurs. Its period is just under 12 hours and its orbit very elliptical ($\epsilon = 0.6$); this makes it appear quasi-stationary. Much of the time its range exceeds 21,750 miles (35,000 km), which means nearly a hemisphere is within its view, and for many hours on end. The satellite enables Amateurs with modest equipment to communicate at some time or other with stations almost anywhere on earth.[1]

OSCAR-10 (**fig. 1**) carries two linear transponders, at UHF and in L-band. Mode B accepts 70cm uplink signals and has a 2-meter downlink; mode L is 23cm up and 70cm down.

Associated with each mode are two alternative telemetry transmissions: from a general beacon (GB) or an engineering beacon (EB). Of these, the 145.810 MHz general beacon is used predominantly, and will be a familiar sound to most users of the 2-meter Amateur band. The other frequencies are 145.987 MHz (EB), 436.04 MHz (GB) and 436.02 MHz (EB).

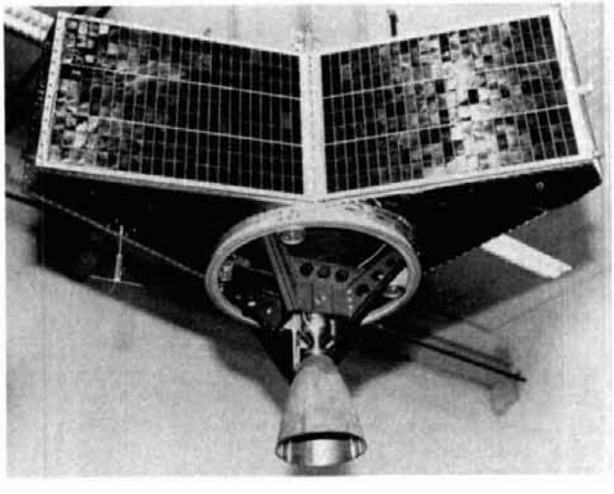


fig. 1. OSCAR 10 (*courtesy AMSAT*).

Transmissions are continuous: on the hour and half-hour UTC there is a 5-minute Morse code bulletin, followed by 20 minutes of phase shift keying (PSK) telemetry. There are two periods of 50 baud RTTY, 5 minutes each on the odd quarter hours.

Telemetry from OSCAR-10 is transmitted in 512-byte blocks, preceded by a four-byte synchronization code (hex 39 15 ED 30) and followed by a two-byte cyclic redundancy check and then a run of about 100-200 padding bytes (hex 50). A byte consists of 8 bits and is transmitted serially, most significant bit first, at a rate of 400 bit/s. So a new block is sent every 12-14 seconds and lasts for 10.3 seconds. The interval before the next block can be used for computer processing of the telemetry.

There are several different kinds of blocks; **Q**, **Y**, and text blocks **K**, **L**, **M**, **N**, are the most common. Their first two characters are always an ASCII identifier, for example M <space>. Line feed and carriage return are in general not used.

K, L, M, and N Blocks are plaintext messages, comprising eight lines of 64 ASCII characters. They are at present used for routine communications between command stations.

Y Blocks are entirely ASCII telemetry (**fig. 2**). The first line contains the time (UTC) and AMSAT day number (0 = January 1, 1978). Lines 2 and 3 are command and control status information. Lines 5-8 are 64 selected telemetry values that may be converted using the equations published in the OSCAR-10 operating manual.[2]

As an example, columns 3, 7, 11, and 15 represent temperatures, which decode as $T = (N-127)/1.82$. The first entries in columns 3 and 7 are the mode B transponder receiver and transmitter temperatures, 16°C and 32°C, respectively.

With the exception of the letter **Q**, **Q Blocks** begin like Y blocks, but lines 5 to 8 contain the full suite of 256 hexadecimal telemetry bytes.[2]

By James Miller, G3RUH, 3 Benny's Way, Coton, Cambridge, CB3 7PS, England

The schedule of transmissions interleaves ASCII blocks and Q blocks. It repeats approximately every 2¼ minutes.

modulation

The digital information stream at 400 bit/s (called the message) is first differentially encoded such that a 1 is represented by a change in the output data stream (i.e. 01 or 10) while a 0 is denoted by no change (00 or 11). This data is next exclusive-ored with a 400Hz clock, low-pass filtered to restrict its bandwidth (third order Bessel, 560 Hz) and then balance modulated on to the transmitter carrier (**fig. 3**). This modulation is called antipodal phase-shift keying (PSK). Carrier phase is either 0° or 180° according to the data. Because of the low pass filtering there is also some amplitude modulation at bit or clock transitions.

The signal spectrum for random data is shown in **fig. 4**. Note the absence (on average) of a carrier or other line components; these would waste transmitter power.

In the following sections, note the distinction between "message" and "data." The data is the stream which represents the message. Let us use the following notation:

M(n)	n^{th} bit of the original message
D(n)	n^{th} bit of the data, derived from message
S, S(t)	transmitted signal
A(t)	signal amplitude
CLK	the 400 Hz data clock
CAR	carrier
$\oplus$	means EXOR; ($A\oplus A = 0$, $0\oplus 1 = 1$ etc.
$\pm$A	means "either A or its inverse"

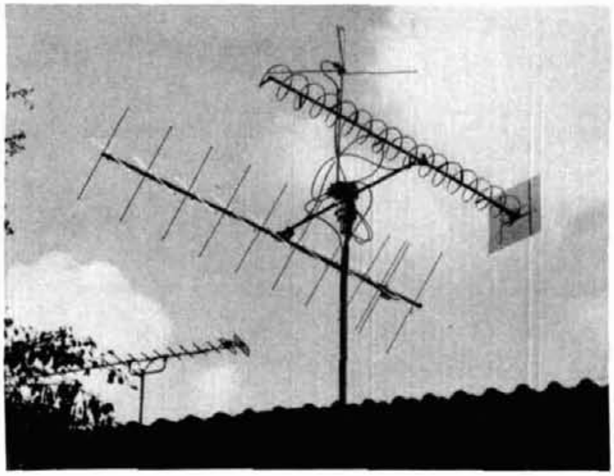
It is useful to remember that if we associate the numeric value +1 to logic 1, and the value −1 to logic 0, then the EXOR operation is equivalent to multiplication, that is: A EXOR B = A × B.

Some features of this modulation scheme that facilitate the demodulation process are the following:

- The signal can be described as $S(t) = A(t)\sin(\omega t)$
- Ignoring amplitude modulation, the signal can also be thought of as the EXOR of data, clock, and carrier: $S = D(n)\oplus CLK \oplus CAR$
- The message stream is related to the data stream by $M(n) = D(n)\oplus D(n-1)$
- Differential message encoding is used to enable a decoder to deal with the unavoidable 180° phase ambiguity in recovered carrier
- All data bits have a mid-bit transition, but not always a transition between bits

demodulation

Essentially this reverses the modulation operations.



Author's helix and cross-Yagi antennas.

The receiver will be set to CW or SSB mode so that the carrier is translated down to audio frequency for input to the decoder (**fig. 5**).

The signal carries negligible information in its amplitude variations, so it may first be limited, which has the great advantage that all subsequent processing can be digital.

First a carrier and clock (denoted by CARR and CLKR) are recovered from the signal (S) and then EXORed with the signal, giving a product (P):

$$P = S\oplus[CARR\oplus CLKR]$$

Provided the local carrier and clock are (excepting possible inversion) replicas of the originals, i.e. CARR = ±CAR and CLKR = ±CLK, this product simplifies to:

$$P = [D(n)\oplus CLK\oplus CAR]\oplus[\pm CARR\oplus \pm CLKR] = \pm D(n)$$

which is the original data. If the signal were noise-free, the data D(n) would be perfectly usable at this point. Noise, however, perforates the bits so mod-bit sampling would lead to random errors. Instead, D(n) is integrated over the bit interval and the resulting accumulation sampled at its end, a process called integrate-and-dump. The system as a whole is a matched filter.[4]

Note that in order to clock the data and time the integration properly, a means must be provided to resolve the CLKR 180° phase ambiguity. (The information to do this is implicit in the signal.)

The message M(n) is next found from ± D(n) by differential decoding. The present data bit is EXORed with the previous data bit. The possible inversion of D(n) is of no consequence, for $-D(n)\oplus -D(n-1)$ and $D(n)\oplus D(n-1)$ are both the same.

The message stream is now available for processing, by either hardware and/or software.

A truly optimum matched filter would take account of the sinusoidal nature of the signal and its amplitude modulation. The hardware required to do this involves deconvolution and is not trivial. The penalty for using a limiter and binary processing is less than 2dB, which in actual on-the-air practice is insignificant.

```
Y  HI, THIS IS AMSAT OSCAR 10                  10:53:03  2308
#034D   #0020   #019E
64  7   0   0   13  225 0

204 0   156 0   199 0   185 126 212 51  152 7   103 47  159 56
0   37  138 31  29  36  140 128 0   0   135 133 111 0   135 13
64  137 141 192 179 170 139 149 118 150 139 13  238 138 148 13
198 140 126 177 198 152 139 0   11  142 141 0   11  137 133 0


Y  HI, THIS IS AMSAT OSCAR 10                  11:08:30  2308
#034D   #0020   #019E
64  7   0   0   13  225 0

204 0   157 0   199 0   185 111 212 51  152 10  103 48  159 56
0   37  139 0   30  36  140 130 10  0   135 146 110 0   135 13
66  137 142 186 177 170 139 96  213 150 140 13  239 138 148 13
198 140 126 10  220 153 139 0   11  142 141 0   11  137 133 0
```

fig. 2. Y blocks are entirely ASCII telemetry. First line includes the day number (2308 indicates April 27, 1984); next two lines contain command and control status information; 64 telemetry values follow.

decoder operation

The receiver should be set to the SSB mode; the normal 2.4 kHz bandwidth is more than adequate, and can with advantage be reduced to around 1 kHz before decoding loss becomes apparent. Because of the decoder's limiter, unnecessarily wide bandwidths reduce performance.

If the signal is tuned in at the middle of the receiver's passband, then the carrier frequency will typically be 1500 Hz. The actual frequency is not important, but there is a lower limit caused by the onset of aliasing, when the lower sideband folds around 0 Hz and into itself. This sets in at a carrier frequency of about 500 Hz. The upper limit is set only by the performance of the logic family; if TTL were used, the carrier could well be at 455 kHz — i.e., at an intermediate frequency.

carrier recovery (fig. 6)

A phase locked loop (PLL) cannot be used to extract the carrier directly because there is no carrier line component in the spectrum of an A(t) sin(ωt) signal where A(t) = random data.

However if the signal is subjected to a nonlinear

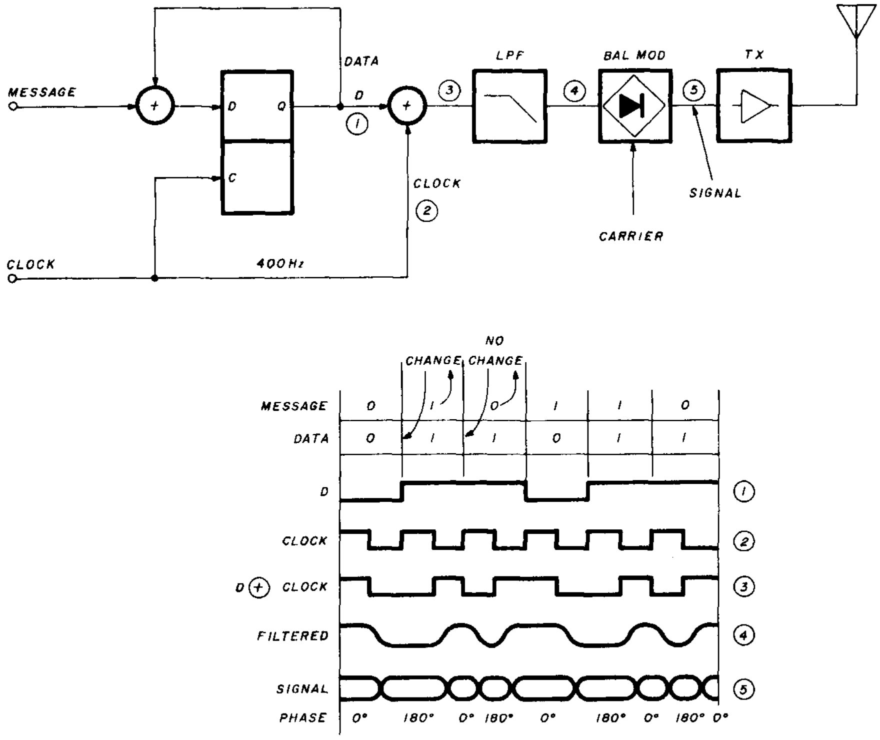


fig. 3. P.S.K. telemetry modulator. Message is first differentially encoded: 0 becomes 00 or 11 (that is, no change in data), while 1 represented by a change (01 or 10). Data is then multiplied (ex-or) with a 400 Hz clock, filtered and modulated on to the carrier.

process such as self-multiplication, the ± is eliminated, and a line component at 2ω is generated. A simple digital way of achieving this is to EXOR the signal with itself delayed by a quarter-cycle. Every zero-crossing of the carrier creates one new cycle of twice the carrier frequency, which can be regenerated with a phase locked loop and followed by a divide-by-two circuit. This division does, however, result in the 180° phase uncertainty previously noted.

The carrier PLL must accomodate receiver frequency instability, noise and changing doppler shift (−250 Hz/hour at 145 MHz, USB. If the loop bandwidth is too small the PLL is difficult to tune in initially, has only a small tracking range, and is generally fussy. If it is wide, with little noise, the loop will hold lock over a wide tuning range, but it will constantly lose lock on noisier signals. So a fixed loop bandwidth will not suit every situation. A few experiments will show what is wanted in practice: the value will lie between 10 and 100 Hz.

clock recovery

This is accomplished in exactly the same way as carrier recovery, by multiplying the data by itself delayed by a quarter-bit. This generates an 800 Hz proto-clock, which is regenerated with PLL and then divided by two. Since the clock frequency is constant, the loop bandwidth can be 1 Hz, even with cassette tape signals. As with the carrier loop, the clock at this stage also has a 180° phase ambiguity.

clock ambiguity resolution (fig. 7)

As long as there are 01's and 10's in the data, which means that the inter-bit transitions will be absent, a second proto-clock of 400 Hz can be generated by EXORing the data with itself delayed by half a bit. Although somewhat sparse (trace 3) this extra clock has the virtues of correct phase, and coherence with the ambiguous clock.

If these two are now EXORed together, the smoothed result is a net high or low. This signal can then be used to invert (or not invert) the ambiguous clock to the correct sense.

In fact the second 400 Hz proto-clock could be used to excite a PLL. However, with the signals encountered in practice, the effective loop gain and bandwidth are caused to vary constantly, which makes the loop rather fragile — though it does work.

A particular feature of the clock and carrier recovery circuits is their aperiodic, digital design, involving no tuned circuits. So their operating frequency can be modified simply by changing the VCO center frequency.

block sync detection

Hex 30, 15, ED, 30 is the pseudo-random sequence generated by the first stage of a five stage shift register having its middle and last outputs EXNORed and fed back to its input, starting off all all 0s.

Using such a feedback shift register, 100% sync detection can be effected by comparing the message stream bit serially with the output of the first stage. If there is a disagreement, the shift register is reset to zero; otherwise it is clock on. If the full sequence is successfully checked the register will reach its last state, which can be detected with a five-bit AND gate and used to set a start-of-block flag. The flag then inhibits the shift register.

byte counting

The block flag releases a byte-block counter. Every eighth count signals that a byte is available, and when the counter reaches 4096 (8 × 512), the block flag is cleared and sync code testing resumes.

outputs

Parallel data output is buffered to TTL level, and consists of an eight-bit byte, positive-going mid-bit strobe, and the block flag. These are brought out to a 20-way PCB connector. Pin-out is compatible with the BBC Acorn microcomputer 6522 user port, which will also provide a 5V supply for the output buffers.

The serializer gives an RS232-type output at 5 volts, 1200 baud, with one start, eight data and many stop bits (50 characters per second). This could be used to drive a printer directly, but the hexadecimal, non ASCII Q blocks will cause unpredictable results — weird characters and reams of waste paper! Using a VDU (video display unit) or the serial port of a computer is tidier.

circuit notes

The complete circuit diagram is shown in **fig. 8**. A printed circuit board is available.[2,5] All other parts may be obtained from Bob Wilson at Radio Kit (Box 411, Greenville, New Hampshire 03048).

A half V_{DD} bias is incorporated to 'float' the op-amps. The 12V supply is not especially critical.

There is no channel filter built into the design because the receiver provides one. The simple limiter U1A* will be effective on a few millivolts of signal. The meter circuit U1B is primarily intended to aid tuning; but with an external switch S1 it can be used to monitor other signals, in particular the state of lock of the two PLL's.

The main PLL, U3, runs at 16 times the carrier frequency. This drives a four-bit shift register, U5, to give

*Note that the author's prepared printed circuit board uses the "IC" designator for integrated circuits. All other figures follow *ham radio* style, designating IC's as U1A, U1B, etc.

— Editor

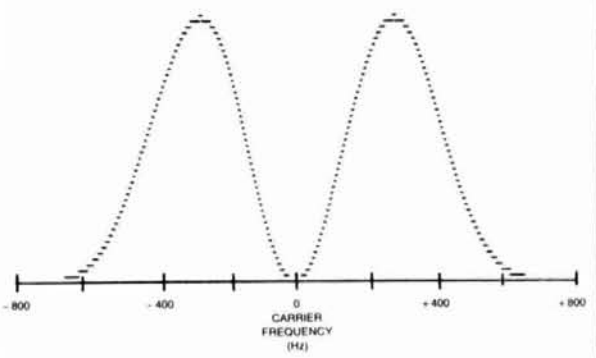
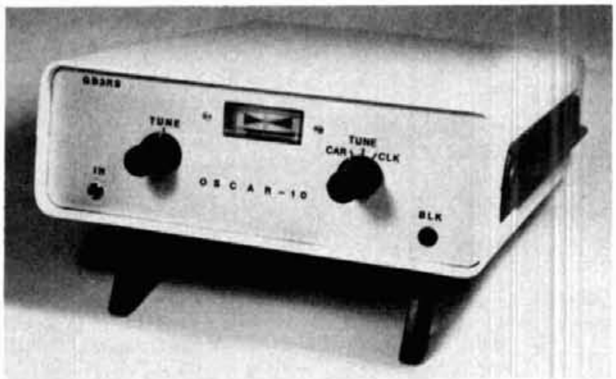


fig. 4. Telemetry power spectral density. Vertical scale is linear. Note the absence of any line components, which would waste power and also make it impossible to lock on to the carrier directly.



Front panel of decoder.

a quarter-cycle delay. The VCO, U3 is followed by a divide by 16 counter, U4, and some logic to generate the local Q:2f signal and the recovered carrier CARR which translates the signal to baseband in U2A. The loop bandwidth is 10 Hz; component values for other bandwidths are shown in the table.

Extraction of the 400 Hz clock is similar to the carrier loop; the VCO, U7, runs at 6400 Hz and has a 1 Hz loop bandwidth. Note the additional quarter-bit delay, U9B, which provides the overall half-bit delay needed for the CLKR ambiguity resolution performed by U10A, U10C, and R13, C14. The clock extraction circuits were originally devised for a UOSAT data demodulator[3] and the integrate-and-dump U1C, U11, and U12A is taken from that source too.

The block sync code detector consists of the feedback shift register code generator U14, U15C and final state (00001) tester, U16A, U16C. The shift register is released when the block bistable, U19A, is clear. Incoming data (bit D(0) is tested against the code generator in EXOR gate U15A. A high state indicates a disagreement, and a CLK pulse resets the shift register. If the shift register reaches its last state, a CLK pulse at AND gate U17D sets the block bistable, UI9A, which also lights an LED. The shift register is reset and inhibited, while the bit/byte counter, U20, is released.

Every eighth count generates a positive, mid-bit byte strobe from the shaper circuit, U18C. This pulse signals external equipment to read the byte in buffer U13, via buffers U21 and U22.

The serializer works as follows. The rest condition has the start bistable, U24B, permanently clocking out "stop" which is cascading through from the shift register DS input, U25, pin 11. When the decoder has a byte ready, byte strobe sets the control bistable U24A pin 1 to "load". This prepares the start bistable U24B to go to 'start' and the shift register U25 to load with bits D(0)-D(7) from the decoder buffer. This happens on the next high-going edge of the 1200 Hz clock, which also clears the load condition. Subsequent 1200 Hz clock pulses shift the data out of the serializer.

Outputs have the following conventions: the parallel data output byte is 1 high, 0 low. D(0) is the least-significant bit. "Block" is high true. Byte strobe is a high going 20 microsecond pulse, and begins mid-bit.

VDU, keyboard, and detector.

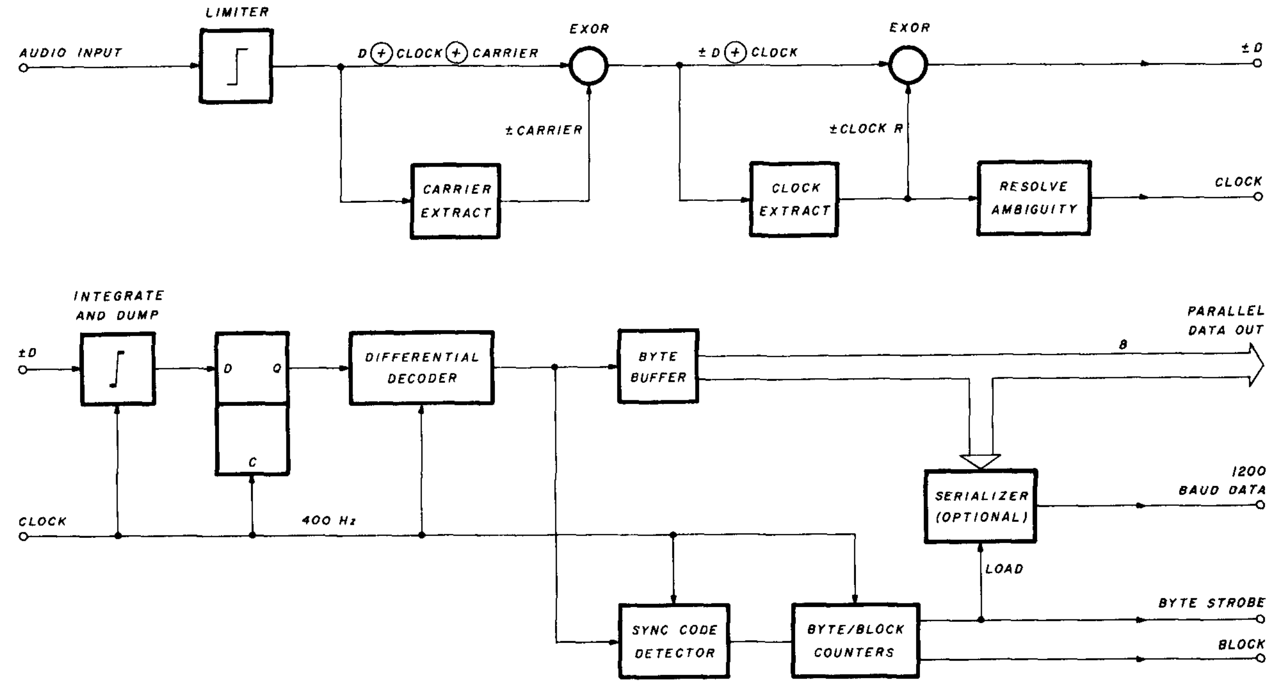


fig. 5. Audio input to the decoder comes from a 2-meter Amateur band receiver. The output is available in parallel or serial form to suit the user's computer.

The serial 1200 baud data format is Start and Data 0 high, Stop and Data 1 low, eight data bits, l.s.b. sent first. The 1200 baud square-wave clock goes low mid-bit. Note that not all data is ASCII in particular the last 256 bytes of Q blocks.

400 bit/s serial data and CLK may be selected as the serial output by changing link D and link C. Square wave CLK goes low mid-bit. The data is *not* in start/stop telegraphy format.

setting up

An audio generator, oscilloscope, and multimeter will be needed. A telemetry data test tape will be invaluable: this can be recorded off-air or obtained via AMSAT-UK[2]. The satellite itself can be used for live testing, but is not always available when you want it. The signal is also noisy, which may confuse matters.

receiver

The first job is to decide what carrier frequency will be used. Trigger the scope at 200 Hz (or 50 Hz if available), tune the receive to OSCAR 10's beacon, and display the audio. Amplitude modulation should be discernible. Experiment with tuning and bandwidth until the signal looks healthy with the mid-bit cross-over clear and sharp. Now trigger the scope with the signal estimate and note the carrier frequency, say f_c. Any frequency exceeding 1000 Hz will be satisfactory.

carrier loop

The objective is to set the loop mid-frequency to the measured carrier frequency (f_c) and achieve a total frequency swing (f_{sw}) at the output of the divide-by-16 (TP1) of $f_{sw} = 800$ Hz. This is slightly complicated by the fact that a 4:1 spread in oscillation frequency between different samples of a 4046 VCO is quite typical.

Start with the VCO swing, nominally given by $f_{sw} = \frac{1}{R7 \times C10}$. Apply V_{DD} and then 0V to pin 9 of the PLL chip, U3, measure high and low frequencies at TP, and subtract. If the difference f_d, is within 25 percent of 800 Hz, then all is well. Otherwise, change C10 for the correct swing.

Calculate and note the desired VCO upper frequency, $f_u = f_c + f_d/2$. Again, connect V_{DD} to pin 9 of the 4046. With the main tuning potentiometer, VR_3, at mid-position, adjust trimmer VR_2 to give frequency f_u. If this cannot be achieved, then change C10 and R7 (in inverse proportion to each other, so as to preserve f_d) and start again.

Now inject f_c at the audio input. Check that the loop locks on this signal. The lock meter should indicate to one side. Slightly vary the input frequency and the main tuning control VR_3 and observe the tuning meter center-zero response. The loop should stay in lock over a range of $\pm f_d/2$.

clock loop

In the same way as for the carrier loop, check that the available frequency swing at TP_2 is about 100 Hz in total. If necessary, change C12 to achieve this. Then adjust VR4 so that the mid-frequency is 400 Hz.

Temporarily ground U2A pin 1. Inject 400 Hz at the audio input. The loop should lock up correctly; this will show on the lock meter. Next inject 200 Hz, which simulates data 010101 . . . (message 11111 . . .). Verify that the loop locks again. Examine the ambiguity signal at U10D pin 12. This should be either high or low and should not vary about $V_{DD}/2$. Now examine the CLK signal at U10D pin 10. The low-going edges should coincide with the transitions of the input signal.

Enlarged view of display.

Disturb loop lock a number of times by removing the 200 Hz input signal for a few seconds. Each time this is done the ambiguity signal will assume a random state, but CLK should always resolve itself to the correct phase.

$V_{DD}/2$ supply

Remove the temporary ground from U2A pin 1. Apply receiver random noise to the system input. Connect an analog meter (on VDC) across pins 1 and 2 of the bistable U12A. Adjust the half-supply control VR_1 for zero reading.

testing

Once the system has been adjusted it may be checked out live or with a test tape[2]. The waveforms obtained should be as shown in **figs. 6, 7, and 9**. A number of features of the satellite data make this easier. The padding character hex 50 and <space> both occur in longish bursts. In addition, the sync code tester will obviously not work unless everything else is going properly, and so illumination of the "block" LED once every 14 seconds for 10 seconds provides a quick, comprehensive overall check.

decoding data

The design of the software to decode and display the data is straightforward enough, but it is outside the scope of this article to present it in full.

The computer should examine the block flag until

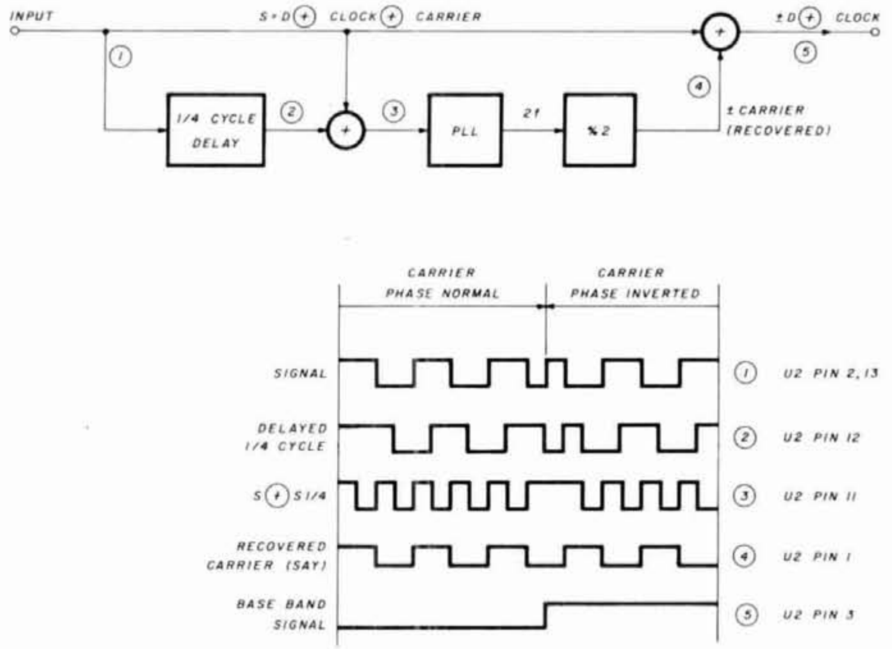


fig. 6. Carrier recovery and conversion to baseband. Clock recovery is identical: simply add 4 to each of the U numbers.

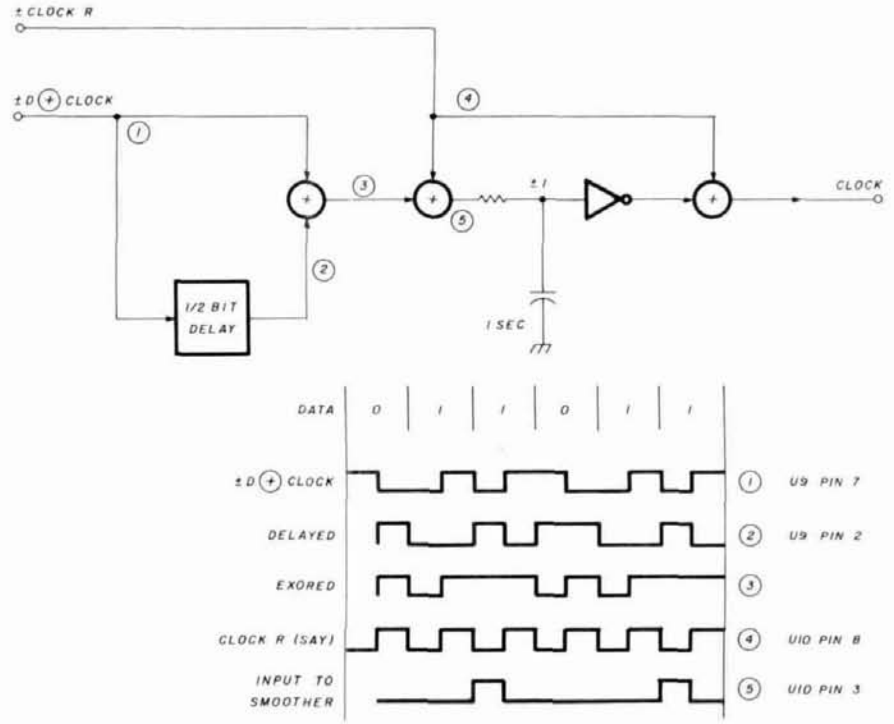
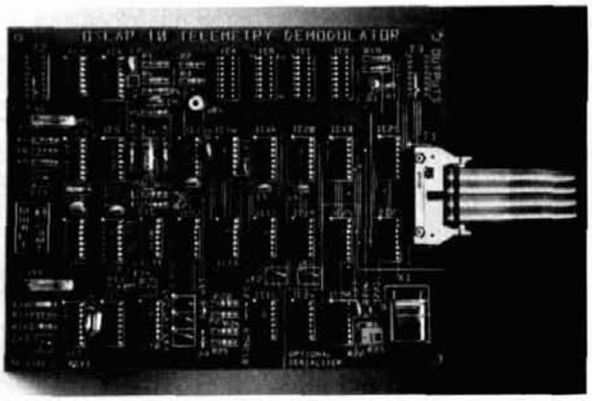


fig. 7. Resolving clock ambiguity: input to the smoother results in a net low on the capacitor, so ex-or inverts *ClkR* to the correct phase, *Clk*. If *Clk* had been correctly phased in the first place, then the capacitor would hold high and *ex-or* would not invert.

it is asserted, then wait for a byte strobe. It should then read in the byte; place it into a 512-byte buffer and await the next strobe. Alternatively, bits may read in serially and packed away.

When all 512 bytes have been read, decoding can begin. In real-time there are 4 seconds in which to do this. Check that the first two bytes are recognizable identifiers, e.g. Q <space>. Then all that remains is to pick out the items of interest such as volts, amperes and temperatures and to display them on a printer or screen in an appropriate format.

Alternatively it is possible to dump the lot, or selected bytes, to storage for later processing, perhaps to monitor specific parameters or to plot graphs.



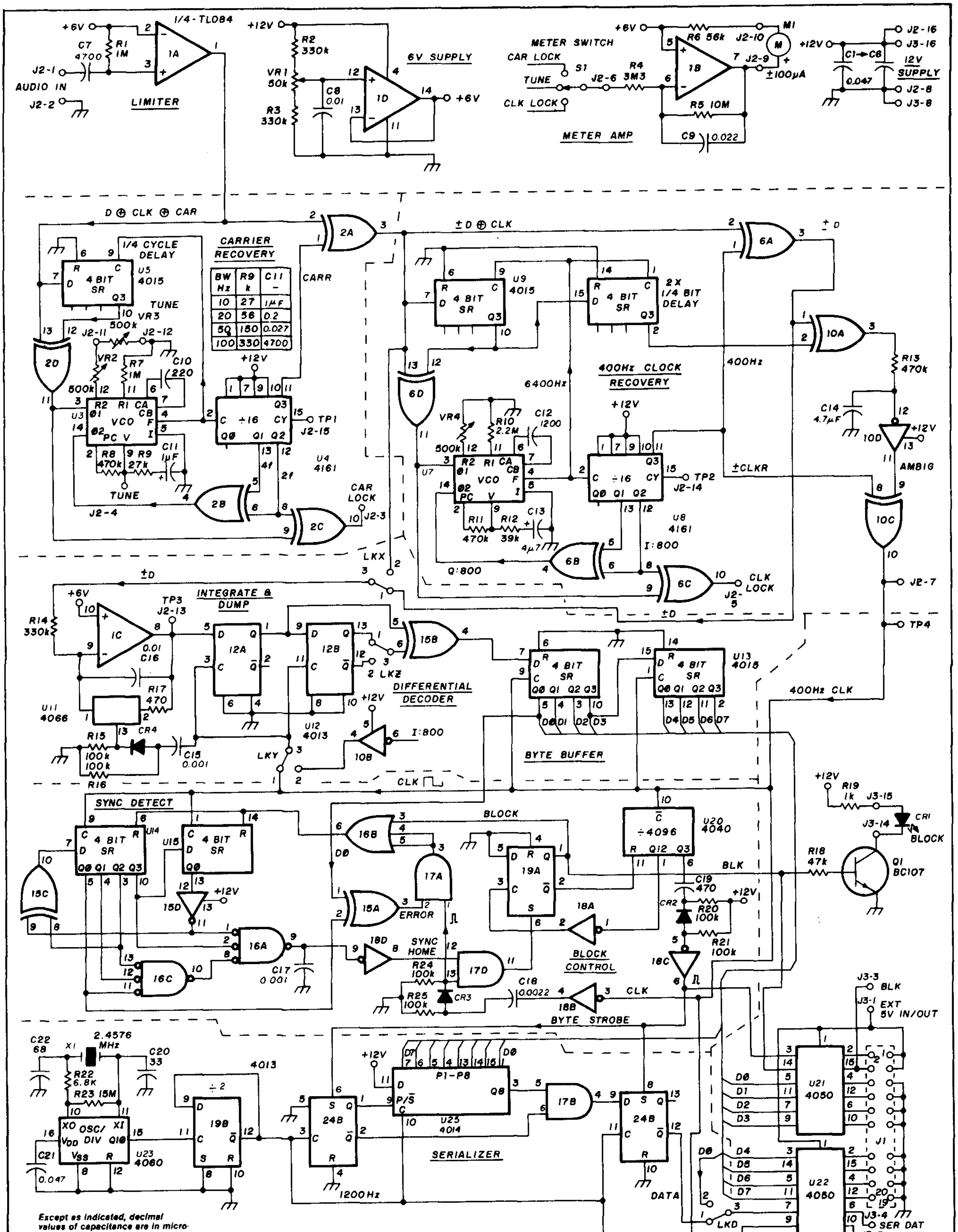
Decoder board with ribbon connector and decoder.

performance

A useful indicator of performance is given by the bit error rate. If we define a reasonable rate as less than 1 in 10,000 bits, i.e. an average of one error every other block, the theoretical channel signal-to-noise ratio S/N should be 2.4dB in 1600 Hz bandwidth. Allowing for the signal amplitude modulation and the limiter, the practical figure is actually about 6.2dB, peak signal power to noise power, or 2:1 in voltage. With care this can be verified experimentally — the signal sounds and looks pretty ragged.

An S/N of 6.2dB is represented in the lab by the surprisingly small figure of 52 nanovolts (nV) (−133dBm) at the input of a receiver having a 3dB noise figure. Now, the 2m general beacon transmits about 1W (+30dBm); the space loss over a 24,850 mile (40,000km) path is 168dB, so the received signal at a unit gain antenna is roughly −138dBm. Thus an antenna gain of 138 − 133 = +5dBi is needed, plus a margin for fading, cable losses, wider bandwidth, higher receiver noise figure and so on.

In practice this means that for satisfactory recep-



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

fig. 8. Circuit diagram of the decoder.

OSCAR-10 PSK demodulator parts list.

capacitors (all 16 volt rates)

C1-C6,C21	0.047 μF
C7	4700 pF
C8	0.01 μF
C9	0.022 μF
C10 — see text	220 pF 5 percent,*
C11 — see text	1 μF tantalum
C12 — see text	1200 pF 5 percent*
C13,C14	4.7 μF tantalum
C15,C17	0.001 μF
C16	0.01 μF**
C18	0.0022 μF
C19	470 pF
C20	33 pF
C22	68 pF

*polystyrene
**good polyester

CMOS integrated circuits

U1	TL084	quad op amp
U2,6,10,15	4070	quad EXOR
U3,7	4046	PLL
U4,8	4161	divide-by-16
U5,9,13,14	4015	quad-4 bit SR
U11	4066	quad switch
U12,19,24	4013	dual-D type
U16	4075	triple-3 OR
U17	4081	quad-2 AND
U18	4069	hex inverter
U20	4040	12-bit divider
U21,22	4050	hex buffer
U23	4060	Osc/14-bit divider
U25	4014	8-bit SR

potentiometers

VR1	50k preset, cermet. Spectrol 62
VR2,4	500k preset, cermet. Spectrol 43
VR3	500k linear, carbon

resistors (all 5 percent)

R1, R7	1M
R2,R3,R14	330k
R4	3.3M
R5	10M
R6	56k
R8,R11,R13	470k
R9 — see text	27k
R10	2.2M
R12	39k
R15,R16,R20 R21,R24,R25	100k
R17	470
R18	47k
R19	1k
R22	6.8k
R23	15M

semiconductors

CR1	LED 10 mA red
CR2,3,4	1N4148 (or equivalent)
Q1	BC107 (or equivalent)

miscellaneous

M1	$\pm$100 μA center zero meter
X1	2.4576 MHz crystal HC33/U size
J1	20-way PCB header for IDC connector
J2,3	16-pin DIL socket
S1	1P3T (1 pole, 3 position) switch
TP1-4	test points

Note: The meter, VR3, CR1, and the switch, S1, are not mounted on the board. Links LKC, D, X, Y, and Z are made from hook-up wire. PC board designators have been left in British style.

tion a modest Yagi or equivalent is needed, pointed at the satellite.

It is worth noting that it is typical of optimal demodulators that they exhibit a marked performance threshold effect. In our "6.2dB" example above, a reduction in the S/N of only 1dB results in a dramatic tenfold error rate increase. This is most apparent where there is a rapid fading (usually induced by the satellite's 40 rev/min spin): what appears to be a healthy signal actually results in bursts of errors at S/N minima. Spin fading occurs most strongly a few hours each side of apogee, when the spacecraft's antennas are not pointing directly towards Earth.

Because of the differential decoding scheme, a single bit error leaving the integrate-and-dump section results in two adjacent bit errors at the system output. This should be remembered if any software error checking is to be attempted.

a further decoding method

Finally, there is another method of decoding the signals. There is a distinctive relationship between the message bits (as opposed to data bits) and the encoded stream. Each message1 results in a $D \oplus CLK$ signal with missing inter-bit transitions, whereas a message 0 does not (**see fig. 3**).

So an alternative decoding method is to treat $D \oplus CLK$ as a stream of 800 bit/s half-bits, grouped in pairs. Two similar successive half-bits are decoded to a logic 1 output, and two differing half-bits to a 0.

This can be implemented most simply by feeding the integrator with $D \oplus CLK$, clocking the intergrate-and-dump and differential decoder with $\overline{1:800}$, and inverting the data output sense! Links X, Y and Z are provided to enable experimenters to evaluate this.

The error properties of this arrangement are interesting. Because the signal energy per dump decision has halved, the half-bits' intrinsic error rate is much higher than a whole bit's, but it is now possible for single message bits only to be corrupted.

The presence of a mid half-bit-pair transition for *zeros implies that the carrier energy per bit for a 0 is* about two-thirds of that of a 1. So message 0s are more easily corrupted than 1s. This contrasts with the

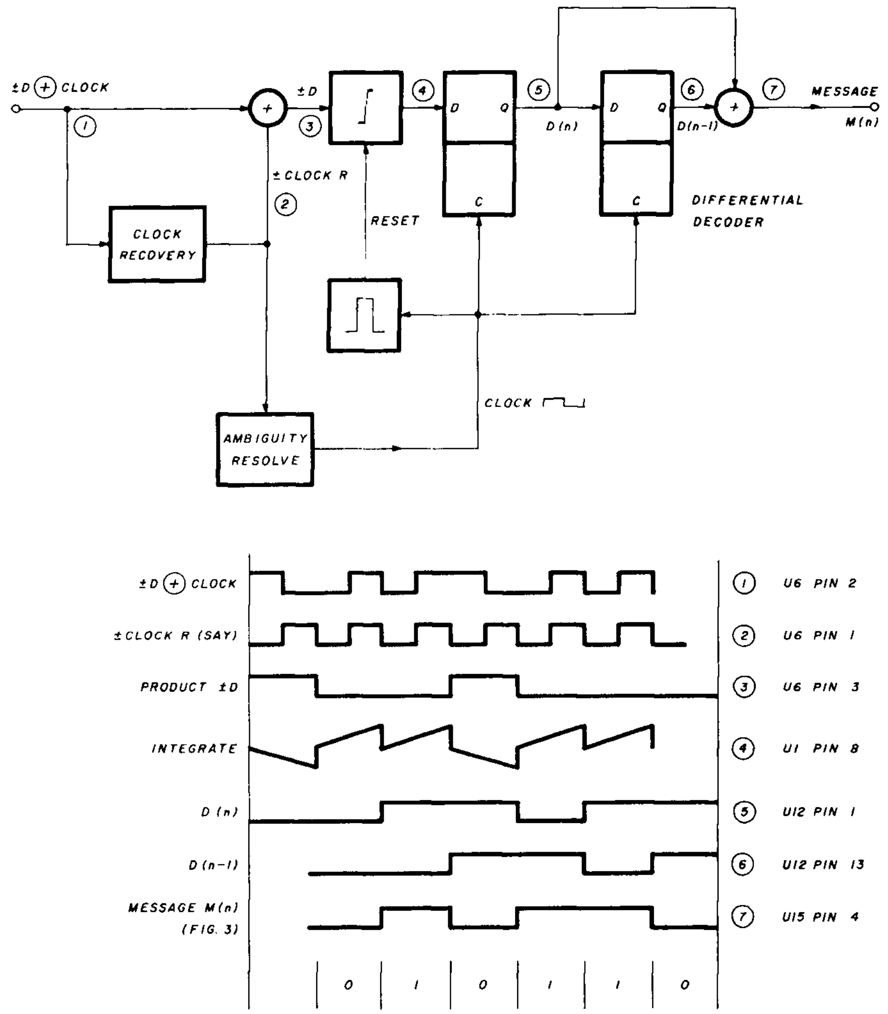


fig. 9. Message recovery: with noise present, the product $\pm$D will be perforated and the triangles in trace 4 will become ragged. The differential decoder compares present data with the previous bit. If they are the same, the message bit is 1; if they are different, the message bit is 0. Thus the polarity of data is unimportant.

whole-bit decoder, where 0 or 1 data bit errors are equally likely but *two* message bits are always corrupted together, though less frequently.

acknowledgements

This article originally appeared in the British publication, *Electronics and Wireless World*, formerly *Wireless World* (October and November, 1984). Sincere thanks are due to editor Philip Darrington for permission to adapt the original text and illustrations for U.S. publication.

Many colleagues deserve mention, especially Trevor Stockill, G4GPQ, for encouragement, PCB layout facilities, and comparative testing with other decoders; Ron Broadbent, G3AAJ, of AMSAT-UK; Janet Miller, for letting me hog our home computer; and Cambridge Consultants Limited, for the free use of facilities.

A double-sided, legended, plated-through, printed circuit board is available from the author at the address indicated on page 50. The price, which includes shipping by air is £ 20.

references

1. M. R. Davidoff, *The Satellite Experimenter's Handbook*, American Radio Relay League, 1984. (Available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, $00.00 postpaid.)
2. AMSAT-UK, London, E12 5EQ, England. Decoder alignment test tape, £ 7.50; "Oscar-10 Operating Manual", £ 5; Telemetry decoding software for BBC (Acorn) microcomputer, on cassette £ 7.50; PCB £ 20. Prices include packing and postage from UK to USA. A bank draft in sterling is requested (cash at own risk). A stamped addressed envelope must accompany ALL inquiries. AMSAT-UK depends on donations.
3. J. R. Miller, "Data Decoder for UOSAT," *Wireless World*, Vol. 89, No. 1568, May, 1983, pages 28-33.
4. A. Viterbi, *Principles of Coherent Communication*, McGraw-Hill, 1966.

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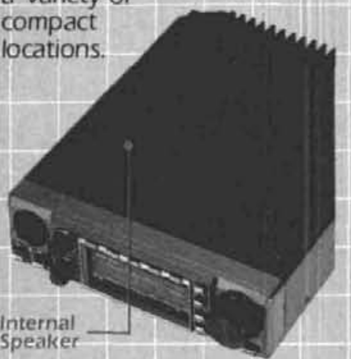

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ham radio TECHNIQUES

Bill Orr, W6SAI

electron-hole theory exposed as fraud

Have you been bemused and confused by the electron-hole therory? Do charges, valence bonds, and the 3/2-power law make you nervous?

A startling discovery by Mark Persons invalidates all of this claptrap! In a recent issue of *Radio World*,[1] Mark reveals how electronic equipment really works:

For many years, young electronic technicians have been taught the "hole" theory of electronics. This theory explains how electrons move along conductors and semiconductors. The explanation has been good enough to satisfy or keep at bay anyone who might otherwise question the theory.

However, after a number of years working in the broadcast industry, I have come to realize the "hole" explanation may not be correct.

My theory, which has been proven time and again by personal observation, is that electronics works on smoke. *Yes, that's right. I recently learned that every manufacturer encapsulates a certain amount of smoke in every piece of electronic component he builds. The smoke is what does the work.*

You have probably noticed that a component will quit working when the smoke leaks out. I've documented this many times and it conclusively proves my theory. My theory sure beats the "hole" theory. I've never seen holes in a wire, and why don't electrons pour out of the end of the wire, if the wire is broken?

I say Mark Persons is RIGHT. I've seen smoke many times, but I've never seen an electron. Hats off to this pioneer whose discovery will be celebrated each April in the years to come!

more on VCR RFI

The subject of video cassette recorder RFI seems to keep coming up. It's a tough problem, and will probably get worse, according to Bill Pasternak, WA6ITF, who writes:

VCR-RFI is becoming a major problem and unless the manufacturers return to a higher quality product as was the case with the earlier models, I am afraid that there is little that can be done to solve the problem.

While broadcast station VCRs are designed to be immune to relatively strong rf fields, this is not true with the consumer machines. For the past five years, the manufacturers have been concentrating on reducing size and bulk and thereby the cost of manufacture by eliminating as much of the internal metallic construction as is possible. In most cases, the modern VCR consists of one or two printed circuit boards on which are mounted all of the electronic components. The only shielding is that of "tin cover plates" soldered over individual circuits that must be shielded to operate. The boards are, for the most part, secured to the plastic mainframe of the VCR and grounding between boards and chassis-of-transport is done with No. 18 wire. This construction technique, combined with the operating frequencies of the unit leaves it wide-open to interference.

The home VCR package is too small and confined to properly shield it without chancing damage of components on the PC boards. Today it appears to be "build them as cheaply as you can so the local discounter can sell them for under $300."

Unlike a TV receiver which can be effectively shielded and protected, this cannot be done with the modern home VCR. Only the equipment manufacturer can solve the problem, and as long as we go to cheaper, plastic construction, the problem will worsen rather than improve.

Finally, if you are not trained on how to service a VCR, don't even open it up to see what's inside. This is one piece of consumer electronics that should only be serviced by a highly skilled technician.

I have written a book that covers the entire spectrum of video recording from the VRX-1000 to the home VHS machine. It is titled Videocassette Recorders: Buying-Using-Maintaining *and it is available from TAB Books. If any reader of your column is contemplating the solution of VCR-RFI, I urge them to read this book, or any other*

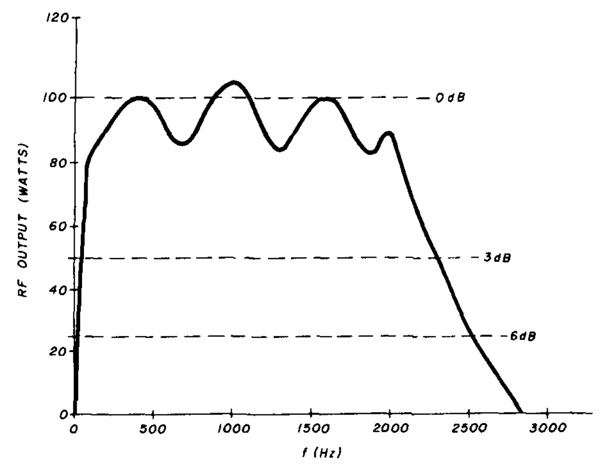


fig. 1. Representative audio response of the Yaesu FT-980 using audio generator at microphone jack and measuring RF power output. (Curve run by NE5S.)

good book on the subject, before they open up the case. Once they read it, however, they will understand why they should turn their problem over to a professional.

Hard words! But WA6ITF has been in video recording since the late 1950s, when Amperex introduced the famous VRX-1000 broadcast video recorder/reproducer. He emphasizes a serious problem that looks like it will only get worse in 1985!

transceiver frequency response

In the October,1984, issue of *ham radio* I wrote about the audio frequency response of various SSB transceivers. I received a note from Emile, N5ES, who writes:

I ran a test on my FT-980 (see **fig. 1**). *While the fluctuations don't seem too bad, it is my opinion that the whole response curve seems 300 to 400 Hz too low. Of course, there are other factors to consider, such as microphone response. I use a SHURE 444 and have gotten consistently good quality reports. Wouldn't it be nice if we had some standards in this respect!* (See "Microphone Calibration," by Daniel Peters, NY6U, *ham radio*, June, 1984, page 73.)

In this regard, Steve, K6FS, says: *The point that concerns me is the apparent confusion among Amateurs between speech "quality" and communication "intelligibility." By no stretch of the imagination can "quality" or "fidelity" be applied to the band-limited, relatively low signal-to-noise, high distortion conditions prevailing in Amateur SSB service. On the other hand, a 2800 to 3000 Hz wide passband will yield adequate intelligibility under typical Amateur conditions, if properly placed in the speech spectrum.*

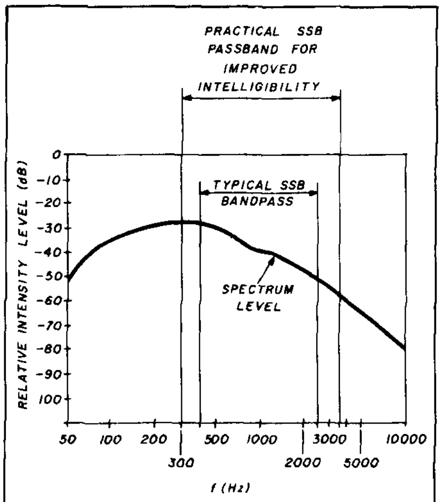


fig. 2. Typical (400-2600 Hz) and improved (300-3500 Hz) SSB passband (long-term average speech spectrum). Figure is adapted from *The Speech Chain* by Denes and Pinson, Bell Telephone Laboratories (1960). (*Data courtesy of K6FS.*)

Results of a good amount of research indicate that practical passband limits are 300 and 3000 Hz (possibly as high as 3300 Hz) with an in-band ripple of plus or minus one decibel. (See "Defining the Decibel" by Michael Gruchalla, *ham radio*, February, 1985, page 51, and "Better Sounding SSB" by Richard L. Measures, AG6K, *ham radio*, February, 1984, page 58. — Ed.)

If it is necessary to narrow the transmitted passband, then intelligibility can best be protected by raising the lower limit to as high as 1000 Hz rather than shrinking the passband symmetrically. Voice frequencies below 1000 Hz are not as important to intelligibility as are those above.

One other point of confusion seems to exist: the terms for expressing filter passband limits. Usually measured in voltage terms, filter characteristics are often expressed as Hz between the "6 dB-down" points. This is equal to the half-power points "3 dB-down," in power terms. (A dB is a dB is a dB. — Ed.)

Measurements based upon power, such as shown in your article in the October, 1984, issue of ham radio *(see pages 109-111), were evidently based on output power measurements. If this is so, the distance between half-power points (shown on right-hand ordinates) delineate much narrower passbands than the "6 dB bandwidth" shown for each filter. Thus, by conventional terms, the IC-730 has a 2000 Hz passband (400-2400 Hz), the KWM-1 about 1600 Hz (550-2150 Hz) and the modified TS-830, 1800 Hz (400-2200 Hz). My suspicion is that all three would sound "muddy," cutting off the critical higher frequencies, as they appear to do.*

I note that my TS-130 service manual directs that carrier insertion be adjusted so as to set the −6 dB points at 400 and 2600 Hz. That's 2200 Hz bandwidth — tolerable but hardly optimal.

I am enclosing a copy of the basic spectrum level curve by Denes and Pinson, published in 1960 by Bell Telephone Laboratories. I have drawn in some passbands showing typical and suggested SSB filter character-

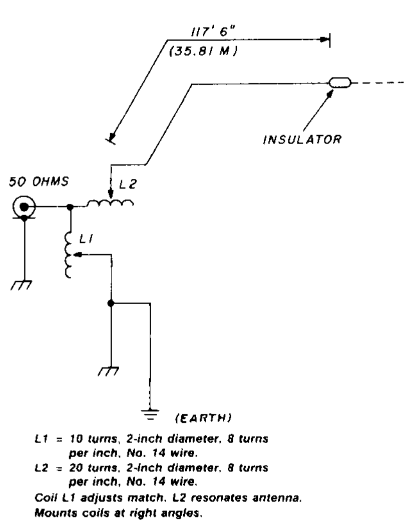


fig. 3. **160-meter Marconi antenna.**

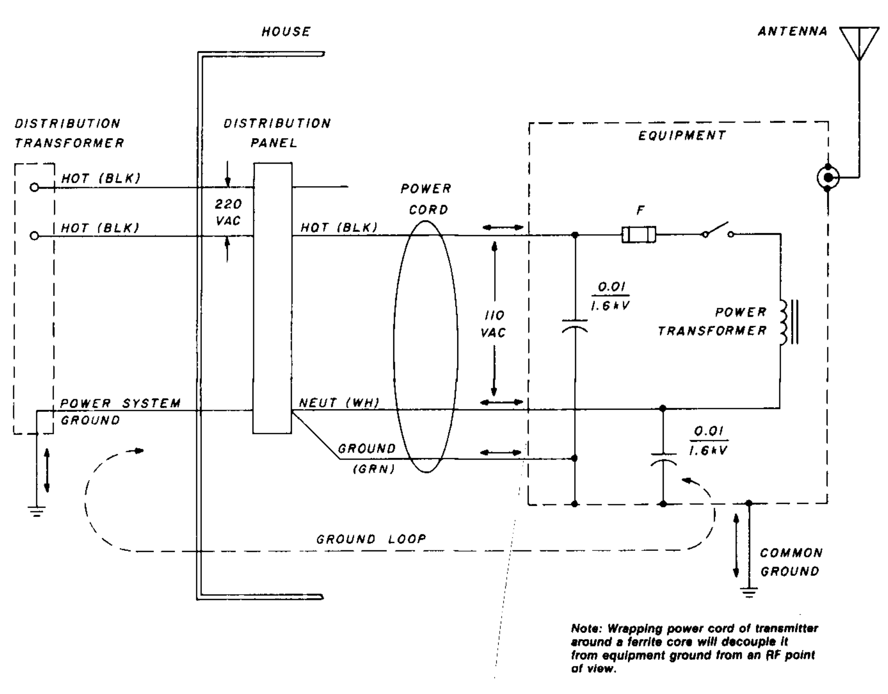


fig. 4. **RF ground currents reach earth ground via power line as well as by radio ground.**

istics to illustrate the relationships discussed (*see* **fig. 2**). *Note how much energy the human voice produces below 500 Hz. It's not essential to intelligible communication.*

Well, it looks to me as if the passband filter in most Amateur SSB equipment cuts off too soon in the HF-voice region. Passing voice frequencies out to about 3500 Hz can improve intelligibility and not widen the spectrum of the signal appreciably. Most Amateur signals I've heard are wider than their voice passband anyway, mainly because of flat-topping in a linear amplifier stage (the "all-knobs-to-the-right" syndrome).

the 160-meter Marconi antenna revisited

My remarks on a 160-meter antenna in my July, 1984, column included a discussion of a practical Marconi antenna and matching network for 160 meters. After six months of use, I've come up with a better, simpler and even cheaper unit. The new design is shown in **fig. 3**.

The Marconi is matched to the 50-ohm antenna part of the transmitter through an L-network, which consists of a shunt inductor and a series capacitor. The capacitor consists of a shorter than resonance antenna — thus it costs *nothing*. The inductor (L1) is quite small and can take the form of a tapped coil. This arrangement eliminates the expensive high-capacitance variable capacitor required for the popular L-network that most Amateurs use.

The antenna is cut for the high frequency end of the band (2 MHz) and has a passband of about 75 kHz between the 2:1 SWR points. A small series inductor (L2) is added to the antenna to operate it lower in frequency.

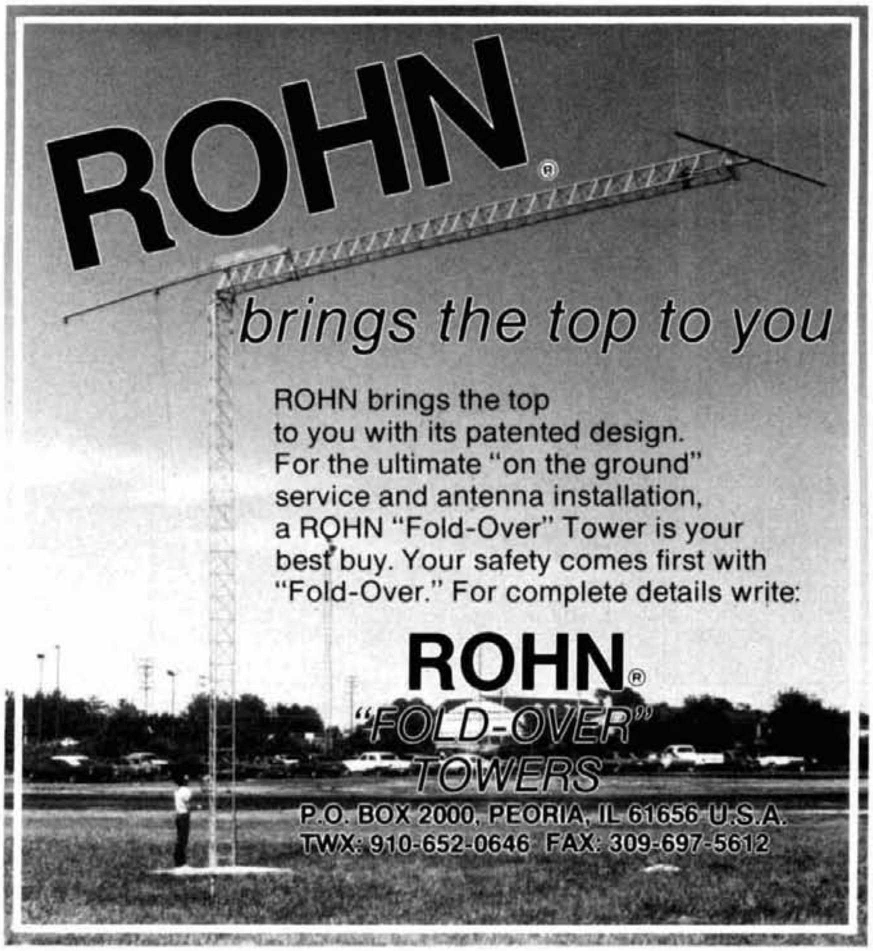
So for the price of two inexpensive inductors, it's possible to construct a Marconi antenna that will work at any point in the 160-meter band.

a few words on ground current

The "mirror image" in the ground makes up the missing portion of the Marconi antenna, and power lost in ground resistance is subtracted from the total power. One problem with 160-meter operation is that the antenna is large with respect to the residence and the electric wiring therein, and it's easy to get unwanted coupling into the power lines that shows up as TVI and RFI in nearby entertainment equipment. (When I first went on 160 meters a few years ago, after a 40-year absence from the band, I was chagrined to find that the ceiling light in the living room lit up every time I transmitted.)

It is not easy to keep ground currents where they belong, since the transmitting equipment is connected directly to the power system ground by means of the power cord (**fig.4**). If a radio ground is added to the transmitter, two ground return circuits exist and a ground loop is formed in which high levels of current can flow. This circulating current can show up as mysterious manifestations in nearby radio and TV receivers as they, in turn, may be coupled back into the power system ground.

Since utility companies demand power system grounding, there's not much that can be done about it. The correct approach, therefore, is to isolate the transmitting equipment (from an RF point of view) from the utility ground, which is often located wavelengths away at the main distribution transformer.



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The solution is to wrap the power cord of the transmitter around a ferrite core, forming an RF choke that isolates the equipment from the utility ground and makes the RF flow to ground via the radio ground attached to the equipment.

While this technique is a *must* for a Marconi antenna, it can also be useful with other antenna types. Regardless of the antenna type you have, if you have RFI problems, try wrapping the power cord of your transmitter (and amplifier, if you use one) around a ferrite rod. Wrap the power cord of the entertainment device around another ferrite rod, too. You might be surprised at how it helps clean up interference.

In my case, I taped two rods together so that the winding form was large enough for the bulky power cord. I got nine turns of line cord around the rods and then tied the power line into position at each end of the core. A recommended rod for the job is the Amidon R-33-075-1200, 12 inches long (30.48 cm) and 3/4-inch (1.90 cm) in diameter. For bands higher than 160 meters, the Amidon R-33-075-750, 7.5 inches (19 cm) long and 1/2-inch (1.27 cm) in diameter, will suffice. (The rods have a permeability of 800.) If a toroidal core is desired, the Amidon FT 240-43 can be used. It is 2.4 inches (6.1 cm) outer diameter and has a permeability of 850.

new list of EME operators

An up-to-date list of all 2-meter EME (moonbounce) operators has been compiled by Lance, WA1JXN. If you would like to have a copy, send a business-size SASE to me at Varian EIMAC, 301 Industrial Way, San Carlos, California 94070, and ask for the "EME List." Please enclose five first-class stamps, or five IRCs, for copying and postage. The list provides calls, addresses, and equipment used at active 2-meter EME stations throughout the world.

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a state-of-the-art electromagnetic jargon generator

The active field discriminator circuit presented in this article operates on the principle of balanced product isolation. Signals from the electromagnetic vector multiplier and one parasitic signal coupler are combined with the output of an external harmonic amplitude detector. The resulting waveform is routed through the isotropic polarization generator for processing, before being applied at the output to drive an orthogonal distortion filter. (See block diagram, **fig. 3***). Possible applications include circular wave oscillator adjustment, as well as optimized linear frequency amplification.*

Impressive, isn't it? The above paragraph from one of my previous articles generated considerable excitement in the technical community, inspired two doctoral dissertations, and ultimately led to the Nobel Prize in Linguistic Obfuscation. But now the secret is revealed: the text above, along with all the rest of my previous technical articles, was generated by a computer. And here, for the first time in print, I reveal the secret of my literary success.

The technique upon which the state-of-the-art electromagnetic jargon generator is based was pioneered by social scientists, perfected by government employees, and has long been the mainstay of the legal profession. It involves no more than generating lists of appropriate buzzwords and catch phrases and combining them in a more or less random manner to produce a desired effect. Frequently three separate columns of words are supplied; thus, creating a ponderous technical term becomes no more complex than ordering dinner in a Chinese restaurant. Simply choose an adjective from Column A; a noun from Column B; and a noun from Column C. Add a fortune cookie ("You will meet an attractive stranger and be disappointed . . .") and a cup of hot tea, and you're ready to go. The result is the generation of phrases that sound important but mean absolutely nothing!

table 1. "Starter list" of technical terms selected by author for optimal obfuscation potential.

column A	column B	column C
linear	wave	amplifier
circular	frequency	oscillator
elliptical	phase	mixer
orthogonal	distortion	filter
isotropic	polarization	detector
harmonic	amplitude	coupler
parasitic	signal	generator
electromagnetic	vector	multiplier
balanced	product	isolator
active	field	discriminator

origin of the specious

Jim Buss, formerly KØQWI, provided the inspiration for this article. As a technical manager at the NASA Johnson Space Center in Houston, Jim generates reams of paperwork daily, including such classic phrases as: Integrated Management Options (IMO), Total Organizational Flexibility (TOF), and Systematized Policy Projection (SPP). Why not, he suggested, apply his literary technique to the fields of microwave and electronic communications?

Why not, indeed? **Table 1** contains a three-column "starter list" of words judiciously selected to meet your technical jargon requirements. Mix and match at will. By changing the suffix of the words in Column C (such as "genera*tor*" to "genera*tion*"), you can create grammatically correct terms guaranteed to fit practically anywhere in a sentence. To automate this process, I have provided, in **table 2**, a BASIC program listing designed to generate up to 1000 unique terms. How's that for Parasitic Distortion Generation?

Remember, Electromagnetic Wave Isolation requires the use of active phase detectors in combination with at least one elliptical polarization coupler to result in a harmonic vector discriminator of unparalleled quality. Now, reread all of my previous *ham radio* articles[1-19] and see how many of these terms you recognize!

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

table 2. Microsoft™ BASIC program facilities generation of up to 1000 incomprehensible technical terms.*

```
10 '---------> JARGON.BAS <---------------------
20 '           Rev. A, 13 Aug '84
30 '           by N6TX
40 '  COPYRIGHT (C) 1984 MICROCOMM
50 '
60 '  Generates totally meaningless combinations
70 '  of Microwave/Electronics buzzwords!
80 '
90 '------------------------------------------
100 CLR$ = CHR$(26) ' Defines Clear-Screen String
110 PRINT CLR$
120 PRINT "DO YOU WISH OUTPUT ROUTED TO:"
130 PRINT
140 PRINT "                    PRINTER (P)"
150 INPUT "          or      SCREEN  (S)";PR$
160 IF PR$="P" OR PR$="p" OR PR$="S" OR PR$="s" GOTO 200
170 PRINT CLR$
180 PRINT "YOU MUST RESPOND WITH 'P' OR 'S'" : PRINT
190 GOTO 120
200 '------------------------------------------
210 PRINT CLR$
220 '             Random Number Seed entered here
230 PRINT "'JARGON.BAS' generates random combinations of"
240 PRINT "Microwave/Electronics buzzwords, for inclusion"
250 PRINT "in technical manuscripts.
260 PRINT
270 PRINT "To start the randomization process, it will be"
280 PRINT "necessary to enter a Seed Number."
290 PRINT
300 INPUT "ENTER ANY NUMBER HERE:  ",S
310 RANDOMIZE (S)
400 '----------------------------------------------------
410 PRINT CLR$
420 INPUT "How many technical terms do you wish to generate";N
430 IF N>0 GOTO 460
440 PRINT : PRINT "number entered must be greater than 1."
450 GOTO 420
460 IF N = INT(N) GOTO 490
470 PRINT : PRINT "number entered must be an integer."
480 GOTO 420
490 PRINT CLR$
500 '----------------------------------------------------
510 '      ARRAY LISTED HERE
520 DIM A$(10,3)
530 A$(0,0) = "LINEAR": A$(0,1) = "WAVE": A$(0,2) = "AMPLIFIER"
540 A$(1,0) = "CIRCULAR": A$(1,1) = "FREQUENCY": A$(1,2) = "OSCILLATOR"
550 A$(2,0) = "ELLIPTICAL": A$(2,1) = "PHASE": A$(2,2) = "MIXER"
560 A$(3,0) = "ORTHOGINAL": A$(3,1) = "DISTORTION": A$(3,2) = "FILTER"
570 A$(4,0) = "ISOTROPIC": A$(4,1) = "POLARIZATION": A$(4,2) = "DETECTOR"
580 A$(5,0) = "HARMONIC": A$(5,1) = "AMPLITUDE": A$(5,2) = "COUPLER"
590 A$(6,0) = "PARASITIC": A$(6,1) = "SIGNAL": A$(6,2) = "GENERATOR"
600 A$(7,0) = "ELECTROMAGNETIC": A$(7,1) = "VECTOR": A$(7,2) = "MULTIPLIER"
610 A$(8,0) = "BALANCED": A$(8,1) = "PRODUCT": A$(8,2) = "ISOLATOR"
620 A$(9,0) = "ACTIVE": A$(9,1) = "FIELD": A$(9,2) = "DISCRIMINATOR"
630 '------------------------------------------------------------------
640 '           PRINT HEADER
650 IF PR$ = "S" OR PR$ = "s" THEN 700
660 LPRINT "     ELECTROMAGNETIC JARGON BY MICROCOMM"
670 LPRINT "     --------------- ------ -- ---------"
680 LPRINT
700 '------------------------------------------------------------------
710 '           START LOOP HERE
715 PRINT CLR$
720 FOR I = 1 TO N
730 '           GENERATE RANDOM 3-DIGIT NUMBER
740 X = INT (RND * 1000)
750 A = INT (X / 100)
760 B = INT (X / 10) - (10 * A)
770 C = X - (100 * A) - (10 * B)
780 PRINT A$(A,0);TAB(17);A$(B,1);TAB(34);A$(C,2)
1000 '------------------------------------------
1010 IF PR$ = "S" OR PR$ = "s" THEN 1030
1020 LPRINT A$(A,0);TAB(20);A$(B,1);TAB(40);A$(C,2)
1030 '------------------------------------------
1040 NEXT I
1050 PRINT : PRINT
1060 IF PR$ = "S" OR PR$ = "s" THEN 1100
1070 LPRINT : LPRINT
1080 LPRINT : LPRINT
1090 '------------------------------------------------
1100 INPUT "TYPE <return> TO CONTINUE, 'Q' TO QUIT ",D$
1110 IF D$ = "Q" OR D$ = "q" THEN GOTO 1130
1120 GOTO 630
1130 END
```

*This program is also available for the Apple IIE. Send SASE.

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

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More Details? CHECK—OFF Page 160

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the ham notebook

improved carrier suppression for the MC1496

The MC1496 has been around for many years and has enjoyed widespread popularity in many balanced modulator design applications. Recently, while attempting to build a 50-MHz DSB generator using this device, I came across several small circuit improvements that will increase the carrier suppression levels, especially at the higher frequencies.

The first improvement involves the use of a bifilar wound toroidal tank circuit that takes advantage of the inherent self-balance of the bifilar windings, yielding a noticeable improvement in carrier feedthrough. In addition, by feeding the signal into the center tap of the bifilar winding through a series choke arrangement, the windings are isolated from unbalanced ground effects.

Normally carrier balance is set through a DC biasing adjustment using a trim pot. The addition of two small-value trimmer capacitors from each side of the MC1496 output pins to ground substantially improves carrier suppression by allowing further balance of the RF tank circuit. If interaction occurs, some readjustments may be necessary for optimum results.

Finally, while the recommended carrier injection level is 60 mV RMS, I found that by varying the LO drive slightly above or below this level often offered improved carrier suppression levels. Any variation in drive level, at 50 MHz, will upset the circuit balance and require further carrier balance readjustments.

While these changes apply to the MC1496 at 50 MHz, one may wish to try similar modifications, at lower frequencies, to improve the expected level of performance from this device. It is likely that these techniques could be applied to the SN76514 and SL6440 IC mixer devices.

Peter Bertini, K1ZJH

bulkhead connector

The type 83 bulkhead connector (83-1F) is useful as a panel-mounted coax feedthrough; however, these fittings are expensive and are available to most hams only through mail order. The PL258 (83-1J) double-female connector is inexpensive and is carried by Radio Shack stores, but it has no provision for panel mounting and attempts to solder a flange to it invariably result in melted polystyrene dielectric. A weekend project can be thwarted by the lack of a suitable fitting.

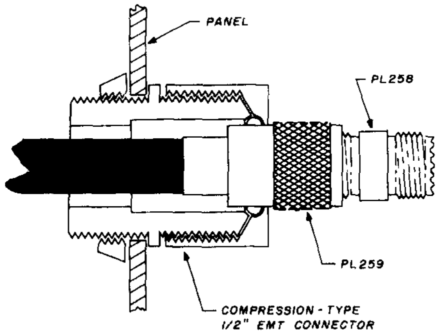

fig. 1. Cross-section of homemade bulkhead connector.

A satisfactory, if not aesthetically perfect substitute for the 83-1F will provide mechanical stability and can be assembled from locally-available parts. The outside diameter of the coupling ring of a PL259 (83-1SP) is nearly the same as that of 1/2″ EMT electrical conduit. Half-inch EMT fittings are available at most hardware stores. A compression-type 1/2-inch EMT junction box connector will grip a PL259/PL258 pair to form a sturdy panel-mount coax connector, as shown in cross-section in **fig. 1**. It is only necessary to remember to slide the EMT connector onto the cable *before* installing the PL259.

Gary Myers, K9CZB

Midway Amateur Radio Club assumes management of North American TRN

The Midway Amateur Radio Club of Kearney, Nebraska, now sponsors the North American Teleconference Radio Net (TRN). TRN links together over 150 gateway stations (mostly VHF/UHF repeaters) across the US and Canada to present high quality technical and informational programs of interest to Radio Amateurs. Past speakers on TRN have included Vic Clark, W4KFC, and Senator Barry Goldwater, K7UGA.

The idea for TRN began with Ed Piller, W2KPQ, and Charlie Kosman, WB2NQV. In the early 1980's Ed and Charlie began linking repeaters by telephone to provide technical presentations as a joint project of the Long Island Mobile Amateur Radio Corps (LIMARC) and the Long Island Chapter of IEEE. However, with the telephone bridging equipment available to them, it was difficult to provide high quality audio to and from all participating repeaters. In late 1982 Rick Whiting, WØTN, a telecommunications engineer, became net manager. Rick made arrangements with Lou Appel, KØIUQ, of Darome, Inc., to use Darome's sophisticated multipoint teleconference bridges to provide the "land line" links for repeaters. The result was superb audio quality and a rapid growth in the number and distribution of gateway stations in the net. Lou will continue to be the bridge engineer in TRNs under the new net manager.

Requests for TRN information should be sent to TRN Manager, c/o Midway Amateur Radio Club, P.O. Box 1231, Kearney, Nebraska 68847-1231. (SASE please, Canada excepted.)

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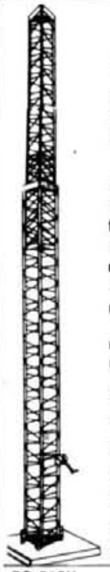

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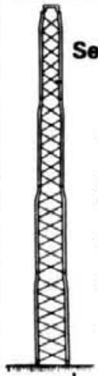

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HDBX48	48 ft	18 sq ft	363	$489

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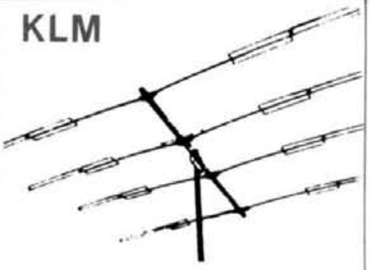

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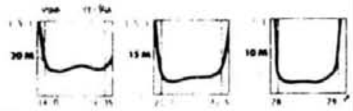


BUTTERNUT ELECTRONICS CO.

HF6V $129 Delivered (Cont. USA)

- Full Legal Power 80/10 Meters
- Optional Stub Tuned Radial Kit Model STR II $29
- Optional Roof Mounting Kit Model RMK II $49 (includes STR II)
- Optional 160 Meter Resonator Kit Model TBR 160HD $49

New 80/40 Meter Vertical Antenna HF2V $129 Delivered (Continental USA)

- Optional 160 Meter Resonator Kit Model TBR 160S $49

TRI-EX TOWERS

W36-36 ft. Crank Up (9 sq. ft. ant.) $579
WT51-51 ft. Crank Up (9 sq. ft. ant.) $899
LM470D-70 ft. Crank Up (motorized) (16 sq. ft. ant.) $3199

RG-213U

$.29/ft $279/1000 ft Up to 600 ft via UPS

- RG-213/U—95% Bare Copper Shield
- Mil-Spec Non-contaminating Jacket for longer life than RG8 cables.
- Our RG-213/U uses virgin materials.
- Guaranteed Highest Quality!

RG-8X

$.19/ft $179/1000 ft

- RG8X—95% Bare Copper Shield
- Low Loss Foam Dielectric
- Non-contaminating Vinyl Jacket

Coaxial Cable Loss Characteristics (DB/100 ft)

Cable Type	Imped.	10MHz	30MHz	150MHz	450MHz
RG-213/U	50	.6	.9	2.3	5.2
RG8X	52	.8	1.2	3.5	6.8
RG-58/U	52	1.4	1.9	6.0	12.5
½" Alum	50	.3	.5	1.2	2.2
½" Heliax	50	.2	.4	.9	1.6
⅞" Heliax	50	.1	.2	.5	.9

HARDLINE/HELIAX™

Lowest Loss for VHF/UHF!

½" Alum. w/poly Jacket ..... $.79/ft
½" LDF4-50 Andrew Heliax ™ $1.69/ft
⅞" LDF5-50 Andrew Heliax ™ $3.99/ft
select connectors below.

HARDLINE & HELIAX ™ CONNECTORS

Cable Type	UHF FML	UHF MALE	N FML	N MALE
½" Alum	$19	$19	$19	$25
½" Heliax ™	$22	$22	$22	$22
⅞" Heliax ™	$49	$49	$49	$49

AMPHENOL CONNECTORS

Silver PL259 ..... $1.25 UG23D N Female . $2.95
UG21B N Male .. $2.95

ANTENNA WIRE & ACCESSORIES

14 Ga. Stranded Copperweld ..... $.10/ft
450 Ohm H.D. Line ..... $.16/ft
18 Ga. Copper coated steel wire ¼ mile long ... $30
H.D. End Insulators $2/ea
Van Gorden 1:1 Balun $11
Van Gorden Center Insulator $6

HUSTLER

6BTV 80-10 mtr Vert. $129
4BTV 40-10 mtr Vert $89 5BTV 80-10 mtr Vert. $109
G6-144B 2-mtr Base $89 G7-144 2-mtr Base ... $119

Mobile Resonators	10m	15m	20m	40m	75m
400W Standard	$16	$17	$19	$22	$26
2KW Super	$20	$22	$25	$29	$39

Bumper Mounts - Springs - Folding Masts in Stock!

CUSHCRAFT

MULTI-BAND HF ANTENNAS
A3 3-el Tribander ..... $219 A4 4-el Tribander $289
R3 20/15/10mtr Vert $279 A743/A744 40mtr Kit $75

HF MONO-BAND ANTENNAS

10-3CD	$ 95	10-4CD	$109
15-3CD	$119	15-4CD	$129
20-3CD	$199	20-4CD	$279
40-2CD	$289	D40	$149

VHF/UHF BEAMS

A50-5	$ 79	617B	$199
214B	$ 79	3219	$ 95
220B	$ 95	424B	$ 79

OSCAR/TWIST ANTENNAS

A144-10T	$ 52	A144-20T	$ 75
A147-20T	$ 63	416TB	$ 59
A14TMB	$ 29	PS4	$ 69

VHF/UHF FM ANTENNAS

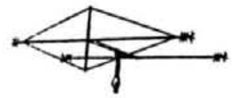
A147-4	$ 29	A147-11	$ 49
214FB	$ 79	228FB	$219
A449-6	$ 29	ARX2B	$ 39

HY-GAIN

Discoverer 2-el 40-mtr Beam $319
Discoverer 3-el Conversion Kit $199
Explorer-14 ..... $309
QK710 30/40 mtr. Add-On-Kit $79
V2S 2-mtr Base Vertical $49
TH5MK26 Broad Band 5-el Triband Beam $389
TH7DXS 7-el Triband Beam $439
TH3JRS 3-el Triband Beam $189
TH2MK3S 2-el Triband Beam $179
205BAS 5-el 20-mtr Beam $349
155BAS 5-el 15-mtr Beam $199
105BAS 5-el 10-mtr Beam $129
204BAS 4-el 20-mtr Beam $259
64BS 4-el 6-mtr Beam $69
66BS 6-el 6-mtr Beam $135
18HTS 80-10 mtr Hy-Tower Vertical $439
LC-160 160-mtr Coil Kit for 18HTS $45
214BS 14-el 2-mtr Beam $49
2BDQ 80/40 mtr Trap Dipole $69
5BDQ 80-10 mtr Trap Dipole $129
BN86 80-10 mtr KW Balun W/Coax Seal $22

MOSLEY

Pro37 7-el Triband Beam $469
CL-33 e-el Triband Beam $279
TA-33 3-el Triband Beam $249
TA-33JR 3-el Triband Beam $189
TA40KR 40 mtr Kit for TA33 ..... $119



MINI-PRODUCTS HQ-1

LIST $182.50 SALE $159

- Wing Span - 11 ft
- Boom - 54 in. long
- Wind Area - 1.5 sq ft
- 1200W P.E.P. Input

ALPHA DELTA COMMUNICATIONS

Transi-Trap™ Surge Protectors—In Stock Now!

Model LT 200W UHF Type ... $19
Model HT 2KW UHF Type ... $29
RT 200W Deluxe UHF Type .. $29
RT/N 200W Deluxe N Type .. $32
HV 2KW Deluxe UHF Type ... $32
HV/N 2KW Deluxe N Type ... $35

KLM

KT34A 4-el Broad Band Triband Beam $349
KT34XA 6-el Broad Band Triband Beam $499
80m-1 80-mtr Rotatable Dipole $595
40m-1 40-mtr Rotatable Dipole $179
40m-2 2-el 40-mtr Beam $309
40m-3 3-el 40-mtr Beam $459
40m-4 4-el 40-mtr Beam $649
2m-13LBA 13-el 2-mtr Beam $79
2m-14C 14-el 2-mtr Satellite Antenna $89
2m-16LBX NEW-16-el 2-mtr Beam $99
2m-22C NEW-22-el 2-mtr Satellite Antenna $119
432-30LBX NEW-30-el-432 MHz Antenna $99
435-18C 435 MHz Satellite Antenna W/CS-2 ..... $119
432-16LB 16-el 432 MHz Beam $69

ROTORS & CABLES

Alliance HD73 (10.7 sq ft rating) $99
Alliance U110 (3 sq ft rating) $49
Telex HAM 4 (15 sq ft rating) $219
Telex Tailtwister (20 sq ft rating) $269
Telex HDR300 Heavy Duty (25 sq ft rating) $519
Kenpro KR-500 Heavy duty elevation rotor $189
KLM EL-3000 Moon Tracker Elevation Rotator $369

Standard 8 cond cable $.19/ft (vinyl jacket 2-#18 & 6-#22 ga)
Heavy Duty 8 Cond cable $.36/ft (vinyl jacket 2-#16 & 6-#18 ga)

SOUTH RIVER ROOF TRIPODS

HDT-3 3 ft Tripod .... $19 HDT-5 5 ft Tripod $29
HDT-10 10 ft Tripod .. $49 HDT-15 15 ft Tripod .. $69
Heavy Duty Tripods include mtg hdw-UPS Shippable

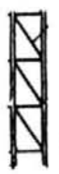
ROHN GUYED TOWERS

10 ft Stack Sections

20G $39.50 25G $49.50
45G $112.50 55G $134.50

All 20G, 25G, 45G and 55G Accessories In Stock at Discount Prices - CALL!

Foldover Towers



Model	Height	Ant Load*	Price
FK2548	48 ft	15.4 sq ft	$899
FK2558	58 ft	13.3 sq ft	$949
FK2568	68 ft	11.7 sq ft	$999
FK4544	44 ft	34.8 sq ft	$1199
FK4554	54 ft	29.1 sq ft	$1299
FK4564	64 ft	28.4 sq ft	$1399

25G Foldover Double Guy Kit $219
45G Foldover Double Guy Kit $249
*Above antenna loads for 70 MPH winds and Guys at Hinge & Apex.

TOWER/GUY HARDWARE

3/16" EHS Guywire (3990 lb rating) ..... $.15/ft
1/4" EHS Guywire (6000 lb rating) ..... $.18/ft
5/32" 7 × 7 Aircraft Cable (2700 lb rating) ..... $.15/ft
3/16" CCM Cable Clamp (3/16" or 5/32" Cable) ..... $.45
1/4" CCM Cable Clamp (1/4" Cable) ..... $.55
1/4" TH Thimble (fits all sizes) ..... $.45
3/8EE (3/8" Eye & Eye Turnbuckle) $6.95
3/8"EJ (3/8" Eye & Jaw Turnbuckle) $7.95
1/2"EE (1/2" Eye & Eye Turnbuckle) $9.95
1/2"EJ (1/2" Eye & Jaw Turnbuckle) $10.95
3/16" Preformed Guy Grip $2.49
1/4" Preformed Guy Grip $2.99
6" Diam - 4 ft Long Earth Screw Anchor $14.95
500D Guy Insulator (5/32" or 3/16" Cable) $1.69
502 Guy Insulator (1/4" Cable) $2.99
5/8" Diam - 8 ft Copper Clad Ground Rod $12.95

PHILLYSTRAN GUY CABLE

HPTG2100 Guy Cable (2100 lb rating) ..... $.29/ft
HPTG4000 Guy Cable (4000 lb rating) ..... $.49/ft
HPTG6700 Guy Cable (6700 lb rating) ..... $.69/ft
9901LD Cable End (for 2100/4000 cable) $7.95
9902LD Cable End (for 6700 cable) $8.95
Socketfast Potting Compound $14.95

GALVANIZED STEEL MASTS

Heavy Duty Steel Masts 2 in OD - Galvanized Finish

Length	5 FT	10 FT	15 FT	20 FT
.12 in Wall	$25	$49	$59	$79
.18 in Wall	$39	$69	$99	$129
.25 in Wall	$69	$129	$189	$249

TEXAS TOWERS

Telephone
(214) 422-7306

VISA MasterCard

Div. of Texas RF Distributors Inc.
1108 Summit Ave., Suite 4 • Plano, Texas 75074

Store Hours: Mon-Fri: 9am - 5pm
Sat: 9am - 1pm

(Prices & Availability Subject To Change Without Notice)

ASTRON CORPORATION

2852 Walnut Ave., Unit E
Tustin, CA 92680
(714) 832-7770

Canadian Distributor
Eastcom Industries, Ltd.
4511 Chesswood Dr.
Downsview, Ontario, Canada M3J 2V6
(416) 638-7995

INSIDE VIEW – RS-12A

ASTRON POWER SUPPLIES

• HEAVY DUTY • HIGH QUALITY • RUGGED • RELIABLE •

RS and VS SERIES

SPECIAL FEATURES

- SOLID STATE ELECTRONICALLY REGULATED
- FOLD-BACK CURRENT LIMITING Protects Power Supply from excessive current & continuous shorted output.
- CROWBAR OVER VOLTAGE PROTECTION on all Models except RS-4A.
- MAINTAIN REGULATION & LOW RIPPLE at low line input Voltage.
- HEAVY DUTY HEAT SINK • CHASSIS MOUNT FUSE
- THREE CONDUCTOR POWER CORD
- ONE YEAR WARRANTY • MADE IN U.S.A.

PERFORMANCE SPECIFICATIONS

- INPUT VOLTAGE: 105 - 125 VAC
- OUTPUT VOLTAGE: 13.8 VDC ± 0.05 volts (Internally Adjustable: 11-15 VDC)
- RIPPLE: Less than 5mv peak to peak (full load & low line)

MODEL RS-50A

MODEL RS-50M

MODEL VS-50M

RM-A Series

MODEL RM-35A

19" X 5¼ RACK MOUNT POWER SUPPLIES

Model	Continuous Duty (AMPS)	ICS* (AMPS)	Size (IN) HXWXD	Shipping Wt. (lbs.)
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50

RS-A SERIES

MODEL RS-7A

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

RS-M SERIES

MODEL RS-35M

• Switchable volt and Amp meter

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H x W x D	Shipping Wt (lbs)
RS-12M	9	12	4½ x 8 x 9	13
RS-20M	16	20	5 x 9 x 10½	18
RS-35M	25	35	5 x 11 x 11	27
RS-50M	37	50	6 x 13¾ x 11	46

VS-M SERIES

MODEL VS-20M

- Seperate 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 x W x D	Shipping Wt (lbs)
VS-20M	16	9	4	20	5 x 9 x 10½	20
VS-35M	25	15	7	35	5 x 11 x 11	29
VS-50M	37	22	10	50	6 x 13¾ x 11	46

RS-S SERIES

MODEL RS-12S

• Built in speaker

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

More Details? CHECK—OFF Page 160

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DX FORECASTER

Garth Stonehocker, KØRYW

sunspot cycle views

In the October, 1984, *DX Forecaster* we discussed the present 10.7-year sunspot cycle. How accurate were our six-month predictions?

The solar flux dropped even lower than the minimum forecast for August as a result of an almost non-existent 27-day cycle variation (i.e., a nearly spotless sun) from mid-September to mid-November. October had the year's lowest recorded monthly solar flux — 74. The minimum daily value of flux so far in sunspot cycle 21 was 69 on September 29. Since October, both flux and activity have increased, approaching the February-March annual maximum of approximately 90 flux units (36 SSN).

Expect the flux to decrease toward an annual minimum during the months of July, August, and September. Expect the *daily* flux to drop down near the previously recorded low of 67 (August 25, 1954) during the summer 1986 or 1987. The sunspot cycle decline has definitely changed during 1984, from the steep decline experienced during 1982 and 1983 at a rate of 4.5 flux units per month to a leisurely rate of about 5 per *year*.

What does this mean in terms of working DX over the next few months? As the solar flux decreases, the MUF can also be expected to decrease. Although this might suggest a pessimistic view of summertime DX, the opposite is often the case: F_2 layer propagation is poorer when the F_2 layer MUFs is low; however, two compensating propagation effects also occur. The first is due to the greater number of hours of daylight in the summer, which means MUF rises earlier in the day. Also, the MUF remains higher until sunset than it does in winter. This effect is mainly felt on paths in east-west and northern directions in our hemisphere. On southern bearings, which are usually transequatorial (TE) one-long-hop in winter, the MUFs in the evening are usually lower in the summertime — i.e., not much TE propagation is available; the high electron density areas ±20 degrees from the magnetic equator just don't build up in the summertime as they do in the winter.

The other compensating propagation factor is sporadic E, which provides short-skip conditions out to 1200 miles (2000 km), with multiple hops possible. This propagation follows the sun across the sky with maximum effect at local noon for higher band DXing and near sunrise and sunset for the lower frequency bands. More detailed information on using E_s propagation will appear in next month's column.

last-minute forecast

The higher HF bands, 10-30 MHz, are expected to be very good during the first two weeks of April, with the 27-day solar flux maximum the main determining factor. Transequatorial openings should be good the second and third weeks of the month, corresponding to disturbed geomagnetic field conditions, and during the equinoctial period. The lower frequency bands, 2-10 MHz, are expected to be best during the third and fourth weeks, at least between weather storm fronts moving by your location. Look for unusual DX on east-west paths that touch the auroral latitudes, (60 to 70 degrees north latitude) during disturbed periods in the middle of the month.

The perigee of the moon's orbit (for moonbounce DX) is on the 5th, with the moon showing full phase on the 5th. There will be a short meteor shower, the Lyrid, on April 20-22, with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on May 5, and ends in mid-May. Its rate is 10 to 30 per hour.

band-by-band summary

Ten meters will be open to the south and southeast for a short period before local noon, to the south at noon, and to the southwest in the afternoon. The openings will be longer when the solar flux is at its 27-day cycle maximum. Even better transequatorial one-long-hop conditions will occur during disturbed periods. Listen to WWV at 18 minutes after the hour and note the geomagnetic field status announcement (A and K indices).

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty should stay open on long southern paths into the night, while 15 will drop out in the late afternoon. Operate 15 first and move down to 20 meters. Contacts out to 5000 to 7000 miles (8000 to 11,200 km) are possible on these bands and one-long-hop transequatorial propagation may also occur, as it does on 10 meters.

Thirty and forty meters are both day and night bands. Intermediate distances 1000 to 1500 miles (1500 to 2200 km) in any direction, considered daytime DX, are better now than in SSN maximum years. Nighttime DX on these bands may be expected to offer greater distance paths than on 80 meters and, like 80, follow the darkness path across the sky. Reduced midday signal strengths and distances may occur on days of high solar-flux values, with 30-meter openings disappearing in the pre-dawn hours on the morning after the high radio-flux values occur.

Eighty and one-sixty meters will exhibit short skip conditions during the daylight hours and lengthen at dusk. These bands follow the darkness paths, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn. The 160-meter band opens later and ends earlier.

Coastal stations and those with good low-angle radiating systems will usually have the edge for working rare DX. QRN will be as low on some nights as that experienced during the wintertime DX season.

APRIL

WESTERN USA

GMT	PST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	6:00	20	20	20	10	15	10	10	20
0300	7:00	20	20	20	10	20	10	10	20
0400	8:00	20	20	20	10	20	15	10	20
0500	9:00	20	20	20	10	20	15	10	20
0600	10:00	20	20	20	15	20	15	15	20
0700	11:00	20	30	20	15	20	20	15	20
0800	12:00	20	30	20	15	30	20	15	20
0900	1:00	*20*	30	20	15	*30*	*20*	*20*	*30*
1000	2:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1200	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1300	5:00	*30*	20	15	*20*	*30*	*20*	*20*	*30*
1400	6:00	*30*	20	15	*20*	*20*	*20*	*20*	*30*
1500	7:00	*30*	20	15	*20*	*20*	20	*20*	*30*
1600	8:00	*30*	20	15	20	20	15	20	*30*
1700	9:00	30	20	15	20	20	15	20	20
1800	10:00	30	20	10	20	20	15	15	20
1900	11:00	30	20	10	15	20	15	15	20
2000	12:00	30	20	10	15	20	15	15	20
2100	1:00	20	20	10	15	15	10	15	20
2200	2:00	20	20	15	15	15	10	15	20
2300	3:00	20	20	15	15	15	10	10	20

MID USA

GMT	MST	N Asia Far East	NE Europe	E S. Africa	SE S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan	CST
0000	5:00	20	20	15	15	20*	10	10	20	6:00
0100	6:00	30	20	15	15	20*	10	10	20	7:00
0200	7:00	30	20	20	20	20	10	10	20	8:00
0300	8:00	*30*	20	20	20	20	15	10	20	9:00
0400	9:00	*30*	20	20	20	20	15	15	20	10:00
0500	10:00	*30*	20	20	20	20	15	15	20	11:00
0600	11:00	*30*	20	20	20	20	15	15	20	12:00
0700	12:00	*30*	30	20	20	20	20	15	20	1:00
0800	1:00	*30*	30	20	*20*	*30*	20	20	20	2:00
0900	2:00	*20*	30	*20*	*20*	*30*	20	*20*	*30*	3:00
1000	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	4:00
1100	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	5:00
1200	5:00	20	*30*	15	15	*30*	*20*	*20*	*30*	6:00
1300	6:00	20	20	15	15	*30*	*20*	*20*	*30*	7:00
1400	7:00	20	20	15	15	*30*	*20*	*20*	*30*	8:00
1500	8:00	20	20	15	15	20	*20*	*20*	*30*	9:00
1600	9:00	20	20	15	10	20	15	20	*30*	10:00
1700	10:00	20	20	10	10	20	15	20	20	11:00
1800	11:00	20	20	10	10	20	15	15	20	12:00
1900	12:00	20	20	10	10	20	15	15	20	1:00
2000	1:00	20	20	10	10	20	10	15	20	2:00
2100	2:00	20	20	10	10	20	10	15	20	3:00
2200	3:00	20	20	15	15	20*	10	15	20	4:00
2300	4:00	20	20	15	15	20*	10	10	20	5:00

EASTERN USA

GMT	EST	N Asia Far East	NE Europe	E S. Africa	SE Caribbean S. America	S Antarctica	SW New Zealand	W Oceania Australia	NW Japan
0000	7:00	20	20	15	15	20*	10	10	20
0100	8:00	30	20	15	15	20	10	10	20
0200	9:00	30	20	20	20	20	10	10	20
0300	10:00	*30*	20	20	20	20	15	10	20
0400	11:00	*30*	20	20	20	20	15	15	20
0500	12:00	*30*	30	20	20	20	15	15	20
0600	1:00	*30*	30	20	20	20	20	15	20
0700	2:00	*30*	30	20	20	*30*	20	15	20
0800	3:00	*20*	*30*	20	*20*	*30*	20	20	20
0900	4:00	*20*	*30*	*20*	*20*	*30*	20	20	*30*
1000	5:00	20	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	6:00	20	*30*	*20*	15	*30*	*20*	*20*	*30*
1200	7:00	20	*30*	15	15	*30*	*20*	*20*	*30*
1300	8:00	20	20	15	15	*30*	*20*	*20*	*30*
1400	9:00	20	20	15	15	*20*	*20*	*20*	*30*
1500	10:00	20	20	15	15	20	*20*	*20*	*30*
1600	11:00	20	20	15	10	20	15	20	*30*
1700	12:00	20	20	10	10	20	15	20	20
1800	1:00	20	20	10	10	20	15	15	20
1900	2:00	20	20	10	10	20	15	15	20
2000	3:00	20	20	10	10	20	10	15	20
2100	4:00	20	20	10	10	20	10	15	20
2200	5:00	20	20	15	15	20*	10	15	20
2300	6:00	20	20	15	15	20*	10	10	20

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

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

✓ 145

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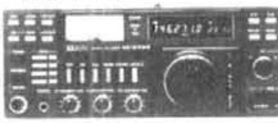

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a DC dummy load

Anyone who works with heavy-duty batteries or with low-voltage DC power supplies develops a keen appreciation for any handy way to test them under load. The instrument described in this article can be built inexpensively in a single evening. It can put the heaviest duty Amateur power supply to the test and reveal a great deal about it.

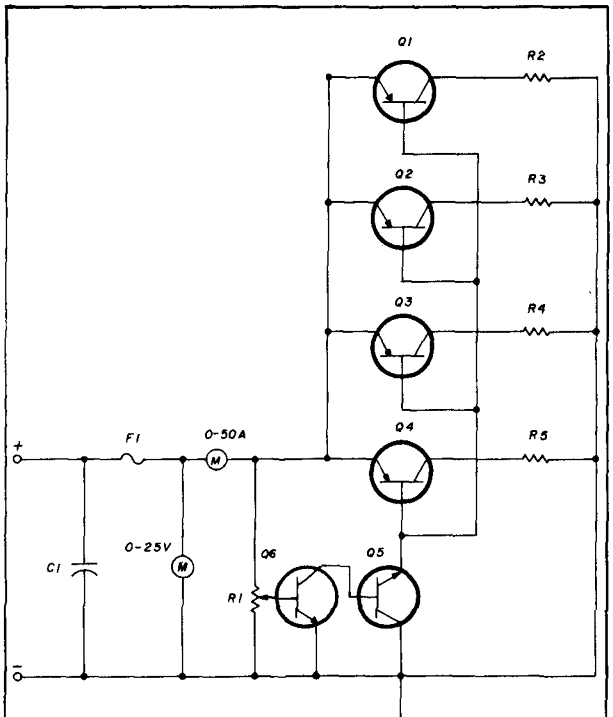
This instrument monitors load current and voltage while allowing smooth adjustment of the load from zero to maximum. With the addition of optional features, it can test transformers for current-handling capability, test the strength of automotive or stationary batteries, and check voltage regulation and overall performance of most low-voltage power systems. Better yet, you can build it from junk box parts or from Radio Shack parts costing less than $25. You'll need some power transistors to absorb energy, an ammeter, a voltmeter, a control potentiometer, a couple of heat sinks, and other assorted hardware.

design requirements are flexible

The circuit for the DC load tester I built is shown in **fig. 1**. *The idea for the unit was not original with* me; I dimly remembered seeing something like it in an old magazine,[1] but I couldn't find the article right away, so I started from scratch.

Because I have some heavy-duty power sources around, including a 50-ampere, 14-volt power supply and a big bank of lead-acid batteries, I wanted a load that could take a lot of current and push these sources hard. With an early breadboard model, I was mystified when a pair of 2N3055 transistors failed well before they reached their rated maximum collector current of 15 amperes each. Then it dawned on me: these transistors have a maximum dissipation rating of 115 watts each, for a total of 230 watts. Their efficiency, of course, is zero, so all that power is dissipated as heat. I was asking those two transistors to dissipate 15 volts at 20 amperes — a total of 300 watts.

By George L. Thurston III, W4MLE, 2116 Gibbs Drive, Tallahassee, Florida 32303



item	description
C1	12,000 µF, 50 volt
F	fuse — value depends on design
Q1,Q2,Q3,Q4	Radio Shack 2043 (PNP)
also suitable:	ECG 219, ECG 180, ECG 61, ECG 184, ECG 68, ECG 34, ECG 93
Q5	RS 2019
Q6	2N2222
R1	1,000 ohms, 1 watt
R2,R3,R4,R5	0.05 ohm, 10 watt

fig. 1. AC/DC load tester.

With a little figuring, I decided I could live with a unit that would dissipate 600 watts, so I chose somewhat higher power PNP transistors, Radio Shack 2043s, rated to handle 15 amperes and 150 watts dissipation for $2.19 each.

I mounted four 2043s on large, black anodized heat sinks. The collector resistors, which serve to keep the current evenly divided among the four transistors, had to be made up from 5-watt, 0.1-ohm units bought from a local supply house at about 45 cents each. Suitable resistors can often be found at flea markets for as little as ten cents or so. The actual value of the equalizing resistors is not critical, as long as you keep them below about 0.15 ohm, but they should be the same for each of your paralleled transistors.

Any number of different types of transistors would work equally well as long as all transistors in any one project are alike and combined power dissipating ability is sufficiently large. To economize on enclosure size and on the number of heat sinks and equalizing resistors, use the fewest high-wattage, high-current transistors that will do the job. NPN or PNP units are equally suitable, though the bias control circuit must be different for each type and, of course, the emitter-collector connections must be reversed (see **fig. 2**).

In the breadboard models a 25-watt, 50-ohm wire-wound rheostat and some fixed resistors were used to control bias on the transistor and consequently the amount of current they drew from the source. However, the wire-wound resistor did not provide smooth control, and a better method was needed. A smaller transistor could be used to regulate the base current and could itself be adjusted with an ordinary potentiometer. But what kind of base-current control transistor would work? Obviously, the control transistor would have to handle the total base current of the combined load transistors, which would be asked to deliver a maximum of 60 amperes.

The Radio Shack 2043s I used had a current-gain ratio (h_{FE} or Beta) of about 20. (The Betas of individual transistors differ quite widely sometimes, even within the same production batch.) This meant that if I wanted 60 amperes from the load transistors, I would need 60/20, or 3 amperes of base drive current.

My junk box yielded a Radio Shack No. 2019, a transistor rated to handle 10 amperes of collector current and with an h_{FE} of about 20. To deliver 3 amperes, it must have about 0.15 amperes (150 mA) of base drive. This is easily obtained from an wire-wound potentiometer hung across the input source. Even less base current would be required by a transistor with a higher Beta (or h_{FE}). The TIP 120, which can deliver 5 amperes of collector current, has a Beta of about 1,000.*

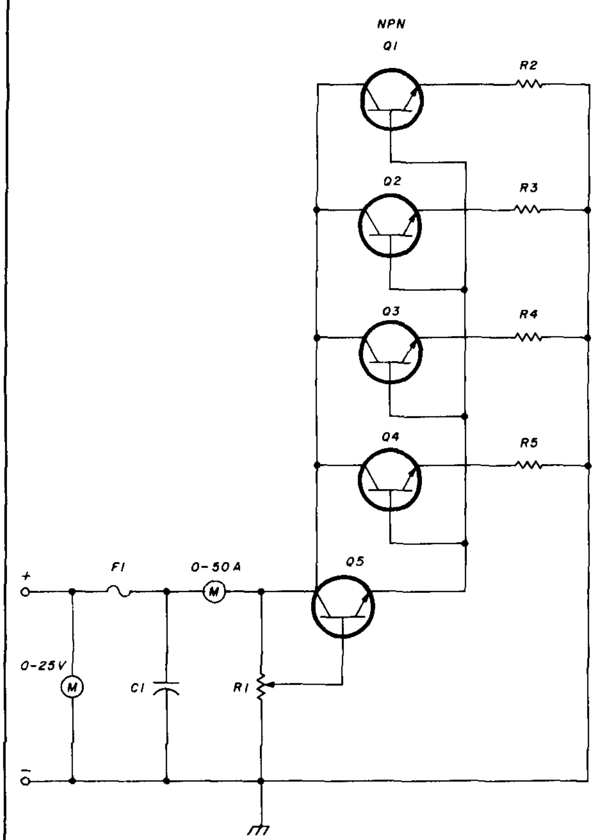


fig. 2. NPN transistors require the emitter-collector connections to be reversed.

metering

I used a commercial 30-ampere meter with an external shunt. (**Figure 3** shows how to use a sensitive micro- or milliamp meter to measure current in several ranges.) The meter measures the voltage drop developed across one of the equalizing resistors. By proper selection of values for R6 and R7, you can choose any convenient current range. One range should go slightly above the maximum current for which the unit was designed.

If you plan a maximum of, say, 30 amperes, the high-range meter should read about 50 amperes full-scale. Assuming two load transistors, half the total current will flow through each resistor. If each resistor is 0.1 ohm, the drop will be 1.5 volts.

If you use a meter with 100 microampere full-scale sensitivity, the calibration potentiometer, R6, will need to be at least 1.5 volts divided by 0.0001 amperes, or 15,000 ohms. Even though the meter's internal resistance will be approximately 1000 ohms, the potentiometer will compensate for it. Use any convenient value between about 15,000 and 25,000. A little circuit board type potentiometer will work fine.

*The final model used an ordinary 1/2-watt carbon pot in the base of a 2N2222, which drives the base of the RS 2019 driver transistor.

To calibrate, use an ammeter of known accuracy in the input lead to the load tester and adjust the calibration potentiometer until the meter readings agree. If your meter has a 0-5 scale, you can read it as 0-50 amperes. A reading of 1 would indicate 10 amperes.

Another way to calibrate, if you don't have a high-range ammeter for reference, is to put a VTVM or FET-VOM across one of the equalizing resistors and increase the load current until you reach some predetermined voltage. Then, knowing the value of the resistor, you can calculate the current flowing through it and set your calibration potentiometer accordingly. In the example above, a reading of 1.5 volts would correspond to a total current of 30 amperes.

For convenience in reading light loads, a second potentiometer can be used to provide a 0-5 ampere range and a panel switch can select the desired range.

testing

Aside from meter calibration and wiring errors, there isn't much to test for except oscillations. Transistors, while capable of amplifying, will sometimes oscillate independently. To prevent this, select transistors with low Beta and low maximum frequency ratings.

If oscillations occur, they can be detected with a scope or with an RF probe on a high-impedance voltmeter. The solution is usually a matter of bypassing something; adding a 0.1 to 0.47μF metal film, solid tantalum or ceramic capacitor, from the transistor base to ground with the shortest possible leads, will generally solve the problem. In multiple-transistor circuits such as our AC/DC load tester, it may be necessary to bypass several transistor bases. If oscillations persist, try bypassing emitters, collectors — anything above ground potential. A last resort might be to insert very small value resistors in series with one or more base leads. In the circuit used, however, the transistor Betas are quite low and oscillations were no problem.

If you use bypass capacitors, be sure their voltage ratings are high enough to withstand the highest voltages you're likely to apply to the input terminals. And remember that when you're testing an AC source through the bridge rectifier, you will encounter voltages about 1.4 times higher than the AC RMS voltage.

applications

The most obvious use of this device is to see which goes up in smoke first — the load or the power supply. But it is capable of much more sophistication than that.

Monitor both the output voltage and current. Then plot a graph of the relationship to get a good record of the quality of regulation of your supply.

Check the fold-back current limiter that's built into many supplies. At what value of load current does it go into action? It may shut down too soon, depriving you of some output capability of the supply. Or it may not shut down soon enough, subjecting your supply to unnecessary stress. Set the load to draw whatever amount of current you think is safe for the supply, then adjust the shut-down threshold to the point at which it turns off the supply.

Monitor the output of the supply with a scope while you slowly increase the load. How much current can you draw before ripple appears in the output? That may tell you something about the design of the filter and the regulation of the rectifier output.

The scope will also tell you if the supply will tend to oscillate at certain load settings — a possible cause of poor regulation and regulator burnout, not to mention TVI or birdies in nearby broadcast radios.

As any auto mechanic knows, the health of a lead-acid battery is best tested under load. A fully-charged battery whose voltage sags to 10 or 11 volts under a load of 10 to 15 amperes is sick! Voltmeter readings across individual cells or hydrometer readings on each cell will spot the defective one and confirm your diagnosis.

Using a bridge rectifier at the input, you can check the performance of a transformer before you build it into a piece of gear. The unregulated output of any transformer, when rectified and filtered, will drop under load. Plot measurements of voltage and load current on graph paper to check performance of the filter. The voltage will fall gradually as load increases until the transformer is delivering all it is designed to provide. After that, the voltage will begin to fall faster with each additional ampere of load current because of copper losses, core saturation, hysteresis, and other problems that crop up when the transformer is overloaded. That point should appear on your graph as a slight "knee" in your curve.

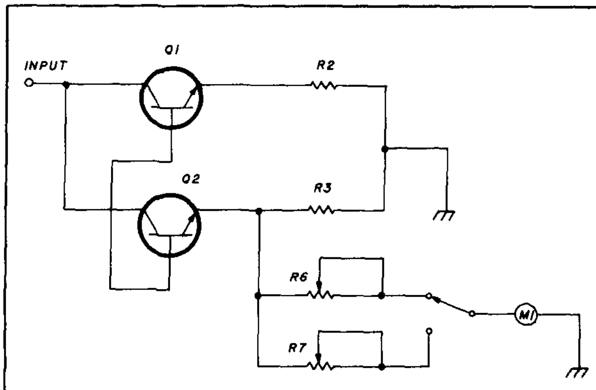


fig. 3. External resistors enable sensitive meter to be used on several ranges. (See text for component values.)

rectifier bridge

To add a bridge rectifier to your dummy load, all that's necessary is to provide two additional terminals for the AC input and to hook them to a bridge (see **fig. 4**). The positive output of the bridge is connected directly to the positive input terminal of the dummy load. The negative bridge terminal goes to the negative load terminal. The bridge output can be left connected to the DC input terminals. It won't conduct with positive voltage applied to the positive load terminal.

Radio Shack offers a 25-ampere, 50-volt rectifier bridge for less than $3 that will work well in this application, provided you never demand more than 25 amperes or apply more than 50 volts to it. For really heavy-duty applications, you may want to build a bridge from discrete diodes, each rated at 30 amperes or more at 10 volts or more.

With AC applied to the bridge, it will handle twice the maximum rated current of the individual diodes, since two diodes are working at once on alternate halves of the cycle. Thus, a bridge with 35 ampere diodes would safely deliver 70 amperes of current — more than adequate for testing most Amateur-service transformers.

The bridge input could be hooked directly to the DC input of the dummy load and would provide automatic

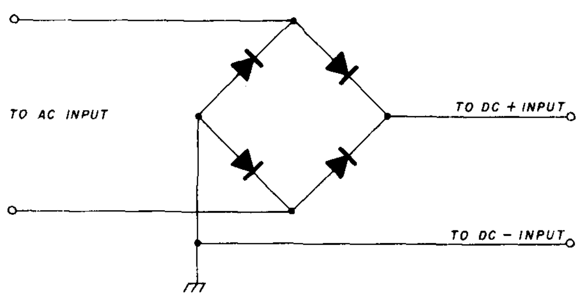


fig. 4. Addition of bridge rectifier adds to versatility of load tester.

polarity correction. No matter how the DC input is connected to the AC bridge terminals, the output from the bridge will always be the same. But there's a good reason for not hooking up the bridge this way. When you put DC on the AC terminals of the bridge, two diodes in series work simultaneously, just as with AC. But they work *all* the time — not alternately, as in a 50 percent duty cycle. That means that DC input to the bridge must be limited to the current each diode will handle, or about half the AC rating. Thus, the Radio Shack bridge would be good for only 12.5 amperes when DC is applied to its input terminals.

another refinement

You will probably find that by advancing the load control potentiometer on your dummy load you can increase the load current enough to disintegrate either the load, the power supply, or both. You can protect against this in several ways:

- Install a fold-back current limiter with an adjustable threshold.
- Use a fuse low enough to blow if you exceed a safe current.
- Limit current by putting a maximum-load resistor in series with the transistors.
- Limit current by padding the control potentiometer with a fixed resistor.

The fold-back limiter may be unnecessarily fussy and complex. The fuse may not blow until the transistor junctions have gone to glory.

Although W7RXV uses fuses as equalizing resistors in his dummy load design, a hazard is involved besides their slowness compared to the junction.[1] Fuses are seldom exactly the same. If one blows before the other, much of the load will be shunted to the other transistors, overloading them and blowing their junctions before the fuses go.

A maximum-load limiting resistor is feasible. A 0.5-ohm resistor inserted between the transistors and ground will prevent the load impedance from going below that value. Thus, with 20 volts applied, the load current would be limited to 40 amperes, even if the transistors shorted or were turned fully on. The same resistor would limit current to 30 amperes at 15 volts and to 20 amperes at 10 volts. This would considerably reduce the range of the load or the effectiveness of the resistor. Additionally, the resistor would have to be rated at 800 watts to handle 40 amperes since $P = I^2R$. More realistically, it would have to be rated at about 450 watts to handle currents up to about 30 amperes. Such resistors are bulky and difficult to find.

The most satisfactory choice could be a resistor in series with the control potentiometer. For NPN load transistors, it would be inserted between the positive input terminal and the potentiometer. For PNP transistors, it would be inserted between the potentiometer and ground. Even this is not foolproof, because it will still be possible to exceed the dissipation rating or the current rating of the load transistors under some conditions.

reference

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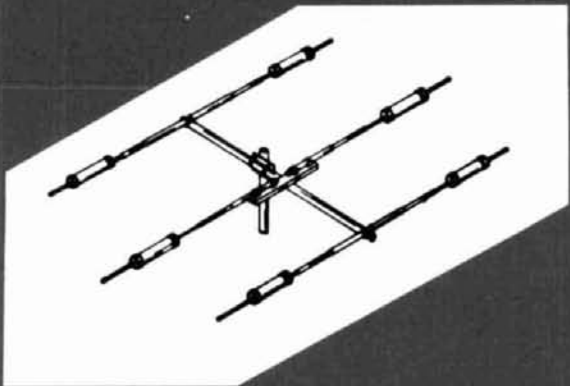

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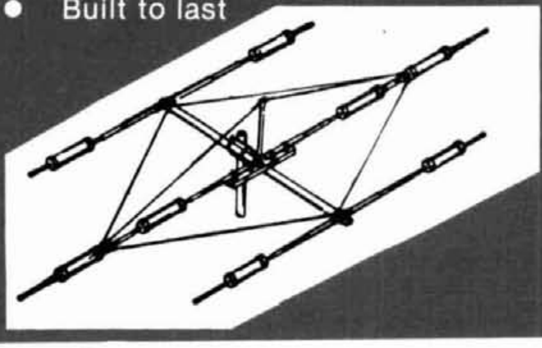

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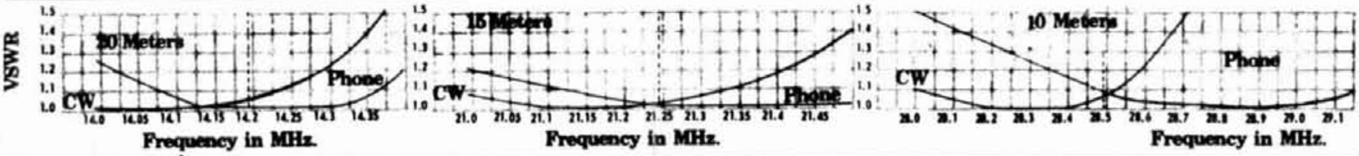

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105	117	129	141	153	165	177	189	201	213	225	237	249	261	273	285	297	309	321	333	345
106	118	130	142	154	166	178	190	202	214	226	238	250	262	274	286	298	310	322	334	346
107	119	131	143	155	167	179	191	203	215	227	239	251	263	275	287	299	311	323	335	347
108	120	132	144	156	168	180	192	204	216	228	240	252	264	276	288	300	312	324	336	348
109	121	133	145	157	169	181	193	205	217	229	241	253	265	277	289	301	313	325	337	349
110	122	134	146	158	170	182	194	206	218	230	242	254	266	278	290	302	314	326	338	350
111	123	135	147	159	171	183	195	207	219	231	243	255	267	279	291	303	315	327	339	
112	124	136	148	160	172	184	196	208	220	232	244	256	268	280	292	304	316	328	340	

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controlled vertical radiation rhombics

part 2: antenna erection and performance

Four towers, careful siting and plenty of wire yield topnotch results

Very few projects can be completed without compromise, and erecting a rhombic antenna is no exception. In this project, the compromise lay between *tower height and low-frequency operation.* An antenna height of about 65 feet was the maximum desired; this meant that 7 MHz would be the lowest design frequency. The desire to have no more than eight wavelengths on the 28-MHz band established the leg length. As pointed out in part 1 of this article, an eight-wavelength array produced an 8-degree beamwidth, and anything less was not desired. Thus two wavelengths on 7 MHz equals 277 feet. The antenna performs well on the 7, 10, 14, 18, 21, 24, and 28 MHz bands and to a lesser extent on the 1.8- and 4-MHz bands.

Based on the height and leg length, the tilt angle for the various bands was determined, with special emphasis on the lowest and highest bands — 7 and 28 MHz, respectively. These bands would determine the greatest width between the side towers and the greatest length between the end towers. **Figure 1** shows the horizontal layout of the designed rhombic, emphasizing the 7 and 28 MHz configurations. Note that a 2 A angle (see **part 1, fig. 3**) of 50 and 20 degrees are required for the 7 and 28 MHz configurations, respectively. This requires the let-out of the rhombic from the side towers of about 72 feet to go from 7 to 28 MHz (see **table 1**). The resulting takeup at the end tower is about 45 feet.

The determination of these parameters is a matter of trigonometry and is not detailed here.

geographical bearing and antenna layout

The determination of the location of the four towers is the crucial point of the design; once the towers are set in concrete, they can't be moved. Two points to consider in this regard are the desired bearing of the rhombic from true north and the allowance for a gap between pulleys and down lead. The downlead supports a counterweight, which must clear the tower as the counterweight is raised and lowered. An additional three feet between pulley and tower is advisable.

Unlike a Yagi, whose beamwidth may be anywhere from 40 to 60 degrees, the eight-wavelength rhombic will have a beamwidth of only 8 degrees. Therefore, accuracy in determining both the true bearing to the desired reception area and the physical positioning of the four towers supporting this antenna is especially important. Check all measurements and calculations carefully. (See references 1 and 2 for how to determine bearings.)

To determine the bearing of my antenna, and to set the ground posts properly, I borrowed a transit and stood where the fixed-antenna tower was to be erected. I sighted Polaris — the North Star — while KA4ECM, my wife Millie, stood about 400 feet away, shining a flashlight toward me. I lowered the transit, to a point parallel to the ground still keeping it pointed squarely in the direction of Polaris, while KA4ECM walked slowly in an east-west direction. As soon as I spotted her light in the transit viewfinder, I signaled for her to stop. She then planted a ground post at that point.

The following day we reset the transit to align on that ground post, and swung it 46.8 degrees from North, inserting a second post at the distant point. (Accuracy of ± 1 degree is recommended.)

By Henry G. Elwell, Jr., N4UH, Route 2, Box 20G, Cleveland, North Carolina 27013

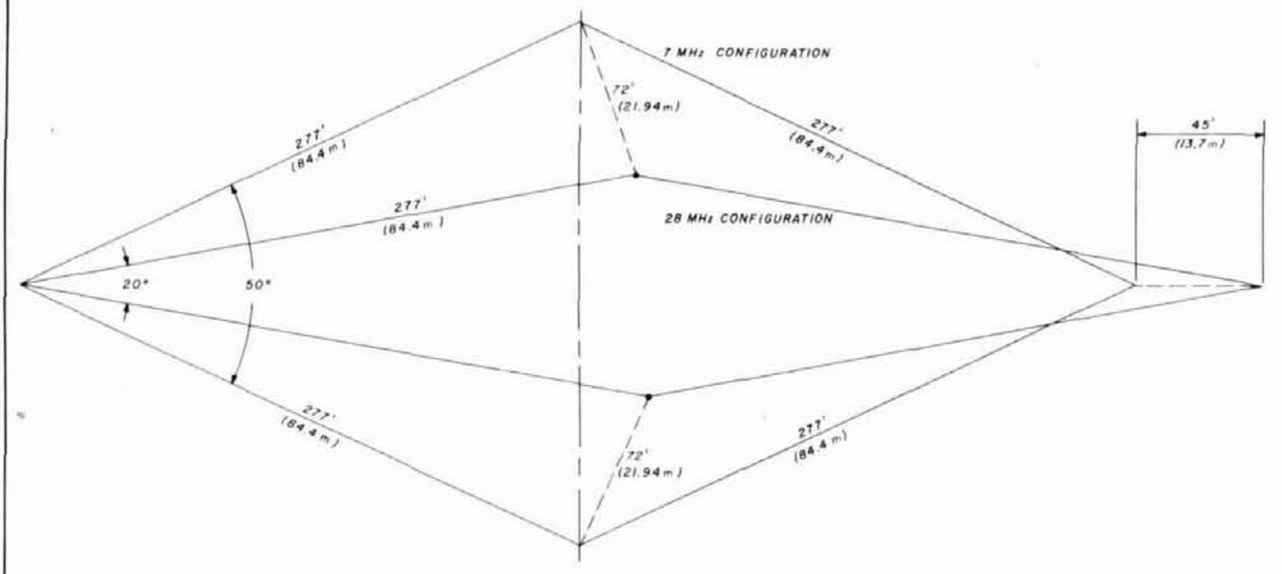
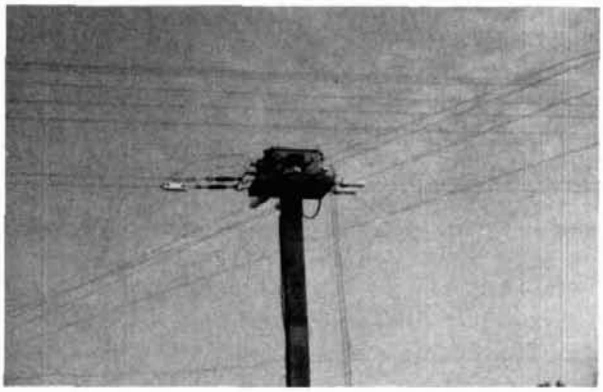


fig. 1. Horizontal layout of controlled vertical radiation rhombic at the lowest frequency (7 MHz) and the highest design frequency (28 MHz band).



Reversing switch at center of rhombic field.

table 1. Correlation between amount of "let-out" and antenna height.

side tower let-out, feet	antenna height feet
0	65.00
10	61.75
20	59.00
30	57.00
40	55.00
50	53.75
60	53.50
70	52.25

We stretched a length of nylon string between the two posts, each representing an end tower, and then took accurate measurements (using a steel tape) to (a) the point representing the center line of the side towers, and (b) the point representing the end point of the 28-MHz rhombic, corrected for the increased distance needed for pulley-tower separation. (Plan on installing a pulley on the fixed end so that the whole array can be dropped to the ground when necessary.)

We then placed the transit on the nylon string at the point representing the side tower's centerline. Ninety-degree right and left bearings were made and posts temporarily set at distances of approximately 120 feet. We strung nylon cord between these posts, took accurate measurements along the cord, and then drove stakes representing the two side towers into the ground.

Even though Polaris is easy to spot with the naked eye, it may be difficult to locate with the transit or telescope. Because of its great distance from Earth, its light reaches the telescope in parallel rays, making magnification difficult. Taking your bearings on a clear, windless — and not too cold — night minimizes the discomfort of an already difficult task.

The terrain over which my rhombic had to be erected was generally level but fell off quite sharply to the east. Various tower heights were necessary to produce an antenna that would be parallel to sea level. Two were 70 feet; a side tower had to be 80 feet; and the far end tower was 100 feet. The necessary tower heights were determined from topographical maps.

If the rhombic is to be erected on ground with a uniform slope extending for at least 1000 yards in front table 2. Control of the vertical angle of radiation is by paying out the side tower cables.

let-out distance (feet)	angle 2A (degrees)	vertical angle of radiation — degrees 7.2 MHz	10.1 MHz	14.2 MHz	21.3 MHz	28.6 MHz
0	50.0	26	16.0	5	beam splits	
10	45.8	28	20.0	11	beam splits	
20	41.6	30	22.5	16	beam splits	
30	37.6	31	24.0	18	6.0	beam splits
40	33.4	32	26.0	21	10.0	7
50	29.2	33	27.0	22	14.0	11
60	25.0	34	27.5	23	16.0	13
70	21.0	35	28.0	24	17.5	15

of the antenna, the practice is to make the front and rear poles approximately equal in height in order to bring the major axis of the antenna parallel to the average ground slope.

The erection of towers has been well covered in the literature. Substantial towers, well guyed in three directions according to the manufacturer's specifications, are necessary. One point to remember is that one guy set should be directly behind the vector force of the antenna at each tower.

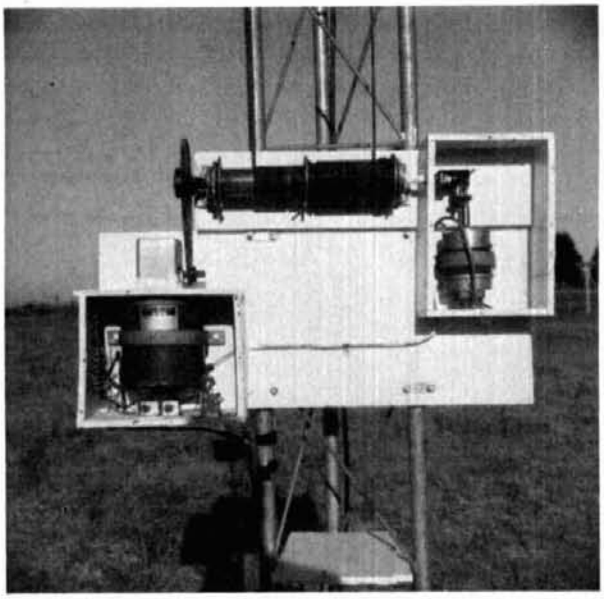
In private communications, Marshall Etter, W2ER, who was chief engineer at the old RCA overseas receiving station in Riverhead, Long Island, stated that tests showed it was not necessary to interrupt the guy wires with insulators for rhombic arrays. However, if the towers are also to be used for mounting Yagis, the use of guy insulators should be considered.

As to choice of antenna wire, the military specifies high strength, 40 percent conductance, using three strands of No. 12 AWG copperweld wire, with a rated breaking point of 2433 pounds. Other wire, such as No. 6 AWG (0.162 inch), 40 percent conductance copperweld, may also be used. W2ER stated that seven-strand No. 16 AWG bronze wire was used in the RCA Rocky Point installations. He advises against the use of solid wire, which tends to vibrate in long spans. I used a special stranded steel core, wrapped with copper, used for aircraft trailing-wire antennas.*

For maximum strength, treat the rhombic as four long-wire antennas. Terminate each leg at an insulator. At the side tower, terminate both the left and right legs on the same insulator holes, connecting the two legs with a flexible jumper soldered to each leg for a good electrical circuit (see **fig. 2**). The flexible jumper provides slack when changing the tilt angle for different bands or propagation conditions.

The end tower antenna legs are terminated in insulators, and two insulators connected to a pear ring

*Marshall Etter, W2ER, has a limited supply of wire, insulators, and other rhombic antenna construction materials. Inquiries (enclose SASE) should be addressed to W2ER at 16 Fairline Drive, East Quogue, New York 11942.



Side tower drive system. Motor drive is at left; take-up spindle, center; synchro position transmitter, right. Note the chain drive from motor to spindle to provide increased torque.

(see **fig. 3**). The pear ring provides a strong, easy way of connecting the three forces: two antenna legs, and the opposing force of the restraining cable.

standing wave ratio

Once the complete system was operational, SWR was measured. **Figure 4** shows the average overall SWR from 160 meters through 10 meters. The increase in SWR in the 160-meter band may be expected because the exponential lines were designed for a minimum frequency of 3.5 MHz and the 4:1 balun cannot be expected to operate properly over such a wide frequency range. Although the big rise in SWR in the 21-MHz band is not understood, it is believed to be associated in some way with the balun. Operation in the 15-meter band is excellent, and with all open wire lines, losses due to a 3:1 SWR are minimal.

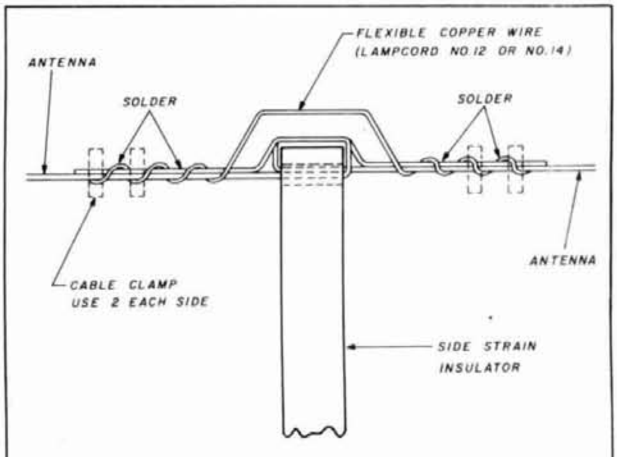
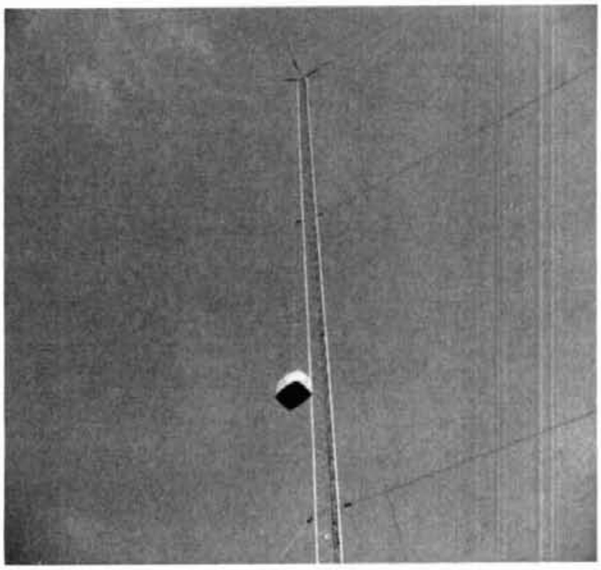


fig. 2. Connection of antenna leg at side towers to obtain antenna strength and good electrical connection between antenna legs.



End tower; 220-pound weight is shown in 40-meter position.

I have long believed in using tubes, which are very forgiving on SWR excursions, in the output amplifier stages. Use of solid-state outputs would require investigation and correction for SWRs greater than 2:1 or suffer the reduction in output levels which such transceivers automatically provide.

report on results

It will take many more months of operation and evaluation to explore fully all the capabilities of the CVR rhombic. After almost a year's operation during its development period, results can be reported in qualitative terms. Additional testing with data on the effects of changing the configuration for operation on a specific band as well as other details may be the subject of another article.

The rhombic has been used on all current Amateur bands from 160 through the 10-meter band. Foster graphs[3] determined the expected results to be as shown in **table 2**. A report on its actual performance, band by band, follows.

• **160 meters.** The rhombic has a constant 3:1 SWR over the entire band. Using an FT102, phone contacts were made from this QTH to the Virgin Islands, to Canada, and west to Texas. Definite front-to-back response was noticed when reversing direction of fire. Comparisons were made against an 80-foot W2LL vertical with apparent advantages going to the vertical for longer distances, and to the rhombic for shorter distances.

• **80 meters.** The rhombic, 1 wavelength on a leg, has an SWR less than 2.5:1 over the entire band. It exhibits definite gain over a W2LL vertical (60-foot tower with TH7DX Tribander on top) for DX. Friends in the New York City area said the rhombic's signal was stronger than that of any other antenna I've ever used. One of the first contacts was with VK6HD, long path at 2157Z. There is about a 15-dB front-to-back difference when reversing antenna fire direction. The beamwidth is noticeable, ZL4PO/C, within the beamwidth, gave me a 58 when all other East Coast stations were getting 56 to 57. However, Australian stations, which are outside the beamwidth by 23 degrees, give me reports comparable to other East Coast stations.

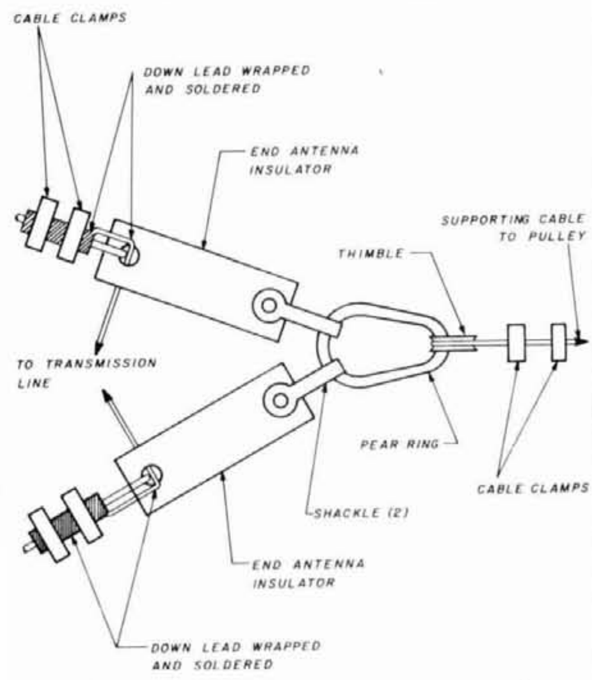


fig. 3. Pear ring used for terminating antenna and supporting cables.

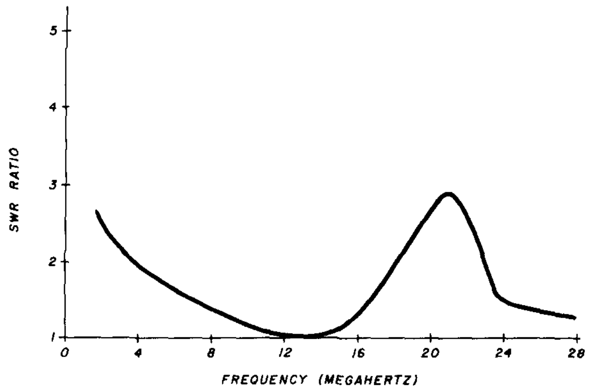


fig. 4. Graph illustrates *smoothed* SWR ratio versus frequency.

• **40 meters.** The rhombic exhibits a sharp change in SWR over this band: 3:1 from 7.0 MHz, dropping to 1.5:1 at 7.2 MHz and remaining there over the rest of the band. These SWR changes are peculiar to my particular setup. Being terminated, the SWR should be reasonably flat over the entire band. The rhombic is 2 wavelengths on a leg on this band. Reversing the direction of fire results in a 30 dB change in signal strength.

I've used the rhombic on this band mainly in contests, with phone results being the most informative. Because contest signal reports are meaningless, antenna effectiveness has been judged by the fact that, except for a single exception, my station received the first reply when a European stood by for Stateside calls on a stated frequency. No other 40-meter antenna was available during this evaluation period to supply comparative reports. However, the "feel" is that the rhombic, in its favored directions, is equal to or better than 3- or 4-element Yagis.

• **30 meters.** The rhombic, 3 wavelengths on a leg, has an SWR of approximately 1.5:1 across the band, and a front-to-back ratio of about 30 dB. Its signal really shines on this band because most of the antennas in competition with the rhombic are relatively simple. In the 40-meter configuration (on the 30-meter band) the rhombic's vertical angle of radiation (VAR) is at its lowest angle — 16 degrees — so ground reinforcement on this band (30-meter) has not been of major significance.

• **20 meters.** The rhombic, 4 wavelengths on a leg, has an SWR practically flat over the entire band; that is, 1.2:1 or less. Operation is with the side insulators 30 feet from the side towers for open-band operation into Europe where its vertical angle of radiation is about 18 degrees. Experience has shown that for band opening, long-path operation, and band closing operation, a 40-meter configuration with a VAR of 5 degrees on 20 meters provides very impressive signals. My first CQ on this band with the FT102 during band-opened conditions was a sufficient reward for all the effort put into siting and construction of the rhombic.

The comparison antenna on this band is a Hy-Gain TH7DX at 60 feet; this is a tribander beam on a 24-foot boom with two driven elements, a director, and a reflector. It is an excellent antenna that puts me among the top callers in most pileups.

The results of comparisons with European stations show a 1-1/2 to 2 S-unit advantage of the rhombic over the TH7. Front-to-back ratio is about 25 dB. The narrow beamwidth of the rhombic is noticeable on this band; thus, signals into Australia are superior on the TH7, as expected, because Australia is off the 3-dB edge of the main lobe by almost 30 degrees.

• **16 meters.** No operation has been accomplished on this band, although a quick SWR check indicated an SWR of 1.7:1.

• **15 meters.** The rhombic is 6 wavelengths on a leg on this band and for some unknown reason, not investigated, the antenna system has an SWR ranging from 3:1 to 2.2:1 over the band.

Preliminary controlled vertical radiation operation was accomplished over the entire configuration change of the rhombic, — that is, with side insulators at zero feet to 70 feet. Because it takes about 30 seconds for a 10-foot increment change of the side insulators, I thought it best to take qualitative measurements on a broadcast station. Radio Berlin at 21.6 MHz was used for the test. A 25-dB change in signal strength was noted between zero feet and 40-50 feet, with a buildup of 5 dB from that point to 60 feet and 70 feet. Much more work has to be done in this area.

• **12 meters.** Again, no operation is permitted on this band, although a quick SWR check indicated a 1.2:1 SWR ratio.

• **10 meters.** On this band the rhombic is 8 wavelengths on a leg and its operation is truly awesome to someone like me, whose biggest 10-meter antenna was a 5-element Yagi at 55 feet. I first operated on this band in February, 1983, with the rhombic in a 40-meter configuration. At that time I had not yet fully explored the beamwidth of the rhombic. But I foolishly asked Roger, N4ZC, with his large antenna farm to work VKs and ZLs with me to see how our signals compared. In both countries his 6-element Yagi at 90 feet was 2 S-units better than my rhombic.

Since that time I've realized that even ZL is too far beyond the 8-degree beamwidth of the rhombic on 10 meters, and the 40-meter configuration simply does not operate properly on 10 meters.

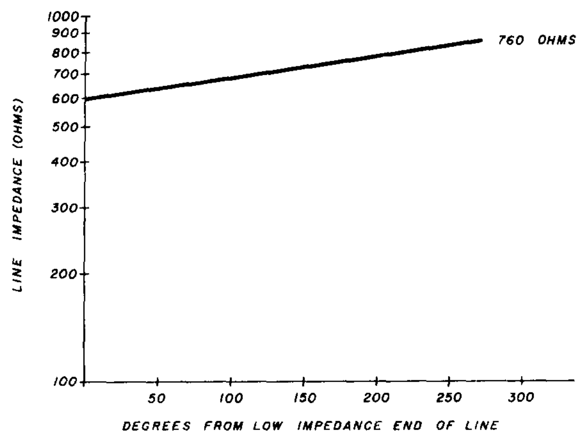


fig. A1. An exponential transmission line is used to transform impedances.

Ten meters opened again in the Fall after I had gained more experience with the rhombic and adjusted it to operate on this band. In the RSGB 10/15 meter contest in October, although not noted for my speedy contest operation, I worked 61 UK stations in 25 minutes because of the strength of the signal I was putting into Europe; "First W heard this morning," and "Loudest signal on the band," were pleasant to hear.

The general opinion appeared to be that the rhombic was 3-4 S-units stronger than the TH7. However, the narrow beamwidth is very noticeable on this band. With 45 ARRL countries within the beamwidth of the rhombic, the population density of Europe was being saturated, which was the original objective of the project.

acknowledgements

A project of this size cannot be accomplished without help and I wish to thank W2ER, WD4KJZ, WD4FFX, W2IRC, W2KXD, Alan Sielke, K1AA, W2LL, and my wife Millie, KA4ECM, whose contributions made it all possible.

references

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2. Chester H. Brent, WB4GVE "Aim Your Beam Right," *73*, June, 1976, page 122.

3. Donald Foster, "Radiation of FM Rhombic Antennas," *Proceedings of the I.R.E.*, Volume 25, October, 1937, page 1327.

appendix

determination of an exponential line

Exponential transmission lines are useful in transforming impedances. The values at any point along the line can be determined using graphical or mathematical approach.[A1,A2] Since these references may not be available to everyone, a description of their method follows.

graphical method

This method uses a minimum transmission-line dimension of one-half wavelength drawn on semi-log paper.

- Mark the vertical scale with the desired impedance range. All cases will probably be from 100 to 1000 ohms.
- Mark the horizontal scale in wavelengths or electrical degrees.
- Mark a point at the 0 degree location with the desired input impedance.
- Mark a point corresponding to the desired output impedance and degrees point. The separation between these two points should be at least one-half wavelength (180 degrees).
- Draw a line between the two impedance points.
- The required characteristic impedance can now be read from the graph at all intermediate points along the line.
- Determine the necessary line configuration from each impedance point from the formula:

$$a = P\ 10^{Z/276} \text{ 2-wire line} \qquad \textbf{(A1)}$$

$$a = \frac{P}{\sqrt{2}}\ 10^{Z/138} \text{ 4-wire line side-connected} \qquad \textbf{(A2)}$$

$$a = \sqrt{2}\ P\ 10^{Z/138} \text{ 4-wire line, cross-connected} \qquad \textbf{(A3)}$$

where a = distance between wires of transmission line, in inches
P = radius of wire, in inches
Z = desired line impedance

Table A1 lists wire radius for various size wires. **Table A2** shows impedance variation along an exponential transmission line as a function of location and line spacing.

Example: Design an exponential 2-wire line to go from 760 ohms to 600 ohms using No. 14 bare copper wire, 3/4 wavelength (270 degrees). **Figure A1** is first drawn and then a table set up to show distance from the 600-ohm point, the impedance representing that distance, and (from the above equations) the line spacing required; see **table A2**. For greater accuracy, use the second method.

mathematical method

The impedance, Z, at any point on an exponential transmission line can be mathematically described as:

$$Z = Z_S\ e^{2s\theta} \qquad \textbf{(A4)}$$

where Z_S = the input or sending end impedance in ohms
s = line length in wavelengths (1 wavelength = 360 degrees)
θ = is a transformation function

The transformation function can be determined from the desired characteristics of the transmission line; i.e., input impedance, Z_S, and desired output impedance, Z, at the end of the line of s wavelengths. Its equation is:

$$\theta = \frac{1}{2s}\ ln\ \frac{Z}{Z_S} \qquad \textbf{(A5)}$$

To solve, determine:

- input and output impedance; Z_S and Z.
- line length; (minimum of 1/2-wavelength at lowest operating frequency); s.
- solve for θ.
- solve for Z for each selected value of s; let s be no greater than 20 degrees; 10 degrees preferably.
- determine the necessary line configuration for each value of Z using line spacing formulas shown in **graphical method**.

Example: Design an exponential 2-wire line to go from 600 ohms to 760 ohms using No. 14 bare copper wire, 3/4-wavelength long (270 degrees).

table A1. Radius of common wire gauges.

wire size	radius (inches)
8	0.0642
10	0.0509
12	0.0406
14	0.0320
16	0.0254
18	0.0200

table A2. Impedance variations along an exponential transmission as a function of location and line spacing.

distance from 600-ohm point (degrees)	characteristic impedance (ohms)	line spacing (inches)
0	600	4.78
20	611	5.23
40	620	5.64
—	—	—
250	750	16.69
270	760	18.14

table A3. Mathematical and graphical method results compared.

distance from 600-ohm point (degrees)	characteristic impedance (ohms)	line spacing inches
0	600.0	4.78
20	610.6	5.22
40	621.4	5.71
—	—	—
250	747.2	16.31
270	760.0	18.14

Steps 1 and **2** have already been determined. It is then necessary to solve for θ, which will become a constant for this example; Z_s is also a constant.

$$\theta = \frac{1}{2 \cdot 0.75} \ln \frac{760}{600} = 0.158$$

Then solve for Z at 10 or 20-degree intervals and tabulate. Twenty-degree intervals are shown in **table A3**.

$$Z = 600\, e^{(2 \cdot \frac{20}{360} \cdot 0.158)} = 610.6$$

Line spacing is then determined:

$$a = 0.032\ (10^{600/276}) = 4.78 \text{ inches}$$

It will be seen that a slight discrepancy exists between the values of Z obtained from the graphical method and from the mathematical method because of difficulty in reading the curve.

For a given spacing, a four-wire line will give a much lower impedance than a two-wire line, and a cross-connected four-wire line will give an even lower impedance (than the same dimensioned side-connected four-wire line). Also the larger the wire-diameter wire, the lower the impedance for a given line separation. These factors may be used to arrive at your design of an open wire exponential line.

references

1. Edmund Laporte, *Radio Antenna Engineering*, Chapter 3, figure 3.81 and Chapter 4, page 422.
2. John D. Ryder, *Networks and Fields*, Prentice-Hall Inc., Chapter 6, Section 6-13.

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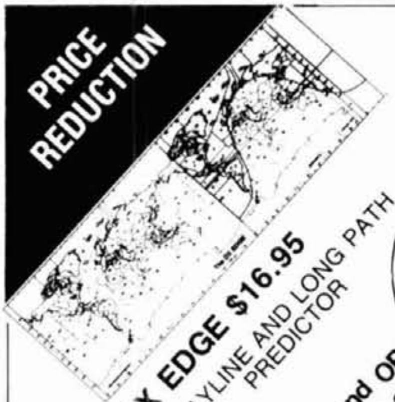

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4.916 Bd. Rate	1.25
5.000000	1.50
5.0688	2.50
5.616	1.59
6.000	2.49
6.176	1.49
7.164112	1.00
7.3728	1.49
8.000	1.49
9.000	1.49
9.90000	1.25
10.69425	2.50
10.8864	1.49
10.920	1.49
11.088	1.59
12.000	2.75
13.440	1.00
14.31818	1.00
15.2	1.10
16.00000	1.50
16.5888	1.49
17.430	2.49
18.000	1.49
18.2259	1.00
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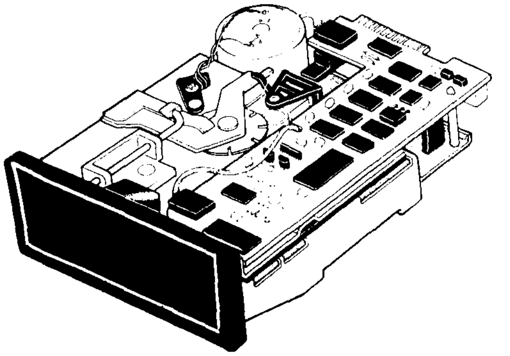

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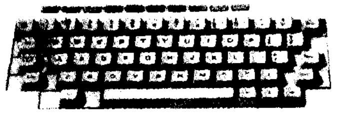

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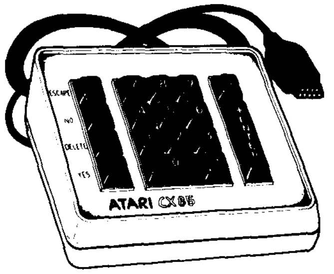

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basic requirements

A terminal unit (TU) decodes the Mark and Space tones from a receiver. An input/output (I/O) port enables the decoded signals to communicate with the computer and software tells the hardware when and how to perform.

A simple serial I/O port and a TU are described. Assembler routines are included for initializing, reading from, and writing to the port.*

RTTY is transmitted in Baudot code. Characters consist of a start bit, five data bits, and stop bits. ASCII characters contain seven data bits and an eighth bit sometimes used as a parity check bit. Each bit in each character must be checked or generated by the computer.

Communication to and from the Timex/Sinclair is most easily accomplished via the cassette port, which represents only one bit of a parallel I/O port. The computer is synchronized with the data by writing precisely timed delays, called software timing loops, into the program.

UART aids data handling

The low price of the Timex/Sinclair computer has no doubt contributed significantly to its mass appeal. One cost-cutting measure in its design was the elimination of the video controller chip. The video display is generated by the microprocessor. When operating in the continuous display mode, the microprocessor devotes most of its time to creating the display with execution of the program carried out only *between* video frames. Consequently there is no time for the microprocessor to use software timing loops to process data at reasonable baud rates.

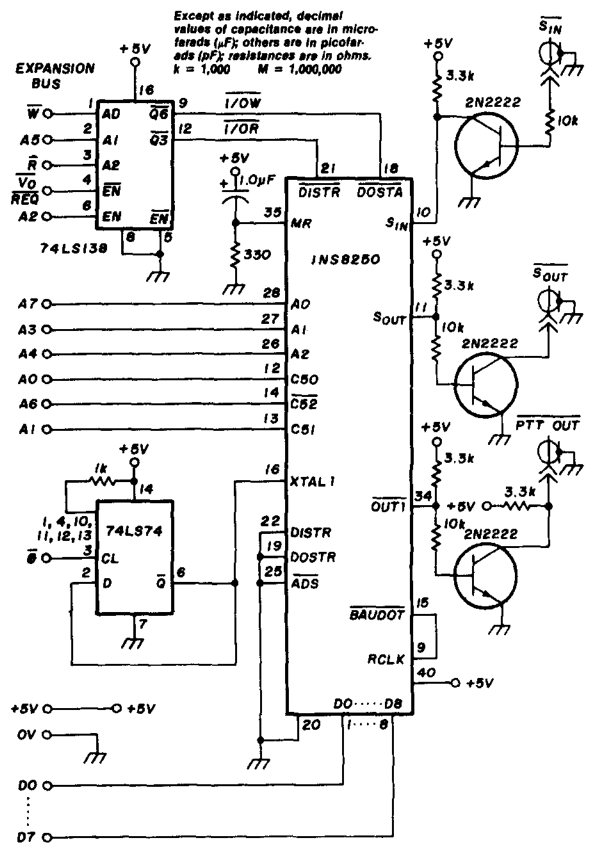
This problem can be solved by the addition of a serial I/O port. In a serial port, individual bit timing is handled by a Universal Asynchronous Receiver Transmitter (UART). The workload on the computer microprocessor is greatly reduced as data bits are assembled into complete characters during receive and characters are converted to serial bits during transmit by the UART.

I first considered using the popular AY series UART, but these devices must be furnished with a baud rate clock. A different clock frequency is required for each desired baud rate. Generation of a stable baud rate clock for the various baud rates used for RTTY and ASCII created yet another problem.

One suggested solution involved using an INS 8250 chip, a combination UART and baud rate generator.

By Cliff Nunnery, NU4V, 313 Vaughn Street, Fort Walton Beach, Florida 32548

*Circuit boards for the Terminal Unit and the I/O port, as well as a full-featured RTTY/ASCII transceive program for Timex/Sinclair 1000 and 1500 computers are available from the author.



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4	3.3 kilohm		
2	10 kilohm		

fig. 1. I/O port schematic.

When supplied to the INS 8250, a clock frequency of up to 3.1 MHz can be divided in frequency by any two-byte word from software to provide the baud rate clock. To implement the port it's necessary only to provide address decoding and to divide the 3.25 MHz computer clock by two.

Although the cost of the INS 8250 is about twice that of a simple UART, this is more than justified by its overall circuit simplicity: a serial port with a crystal controlled, keyboard adjustable baud rate can be built from a design using only three chips.

configuration of INS 8250

The INS 8250 is enabled by bringing pins CS0 and CS1 high and pin CS2 low. Eight internal registers are addressed by pins A0, A1, and A3. Data is strobed into or out of these registers by pins DOSTR and DISTR.

table 1. Initialization and Read/Write routines. (Routines written in Z80 mnemonics.)

```
;               ROUTINE TO READ A CHARACTER FROM PORT
;
;    ROUTINE RETURNS WITH NEW CHARACTER IN 'A' REGISTER IF CHARACTER
;    READY, ELSE RETURNS WITH 'A' REGISTER = FFH
;
;
RXCHR:  IN      A,(LNSTAT)      ;READ LINE STATUS REG
        BIT     0,A             ;IS A CHAR READY?  NZ = YES
        JR      Z,NORDY         ;GO IF NOT CHAR READY
        IN      A,(DBUFF)       ;READ CHAR FROM REC BUFFER
        RET                     ;RETURN WITH CHAR IN A REG
NORDY:  LD      A,(0FFH)        ;PUT FFH IN A REG
        RET                     ;RET WITH A = FFH IF NOT CHAR READY
;
;
;
;
;               ROUTINE TO SET PORT FOR TRANSMIT
;
XMIT:   LD      A,RTTYT         ;GET MODE CONTROL WORD IN A REG
        OUT     (LNCON),A       ;ENTER WORD IN LINE CONTROL REG
        LD      A,4             ;PREPARE TO SET BIT 2 IN MODEM CONT REG
        OUT     (MCONR),A       ;ACTIVATE PRESS TO TALK OUTPUT
        RET                     ;DONE: EXIT
;
;
;
;             ROUTINE TO WRITE A CHARACTER TO THE PORT
;
;   ENTER ROUTINE WITH HL REGISTER POINTING TO CHAR TO BE TRANSMITTED
;   ROUTINE CHECKS TO SEE IF THE INS8250 IS READY TO SEND ANOTHER
;   CHARACTER AND RETURNS WITHOUT TAKING ANY ACTION IF PORT NOT READY
;   CHARACTER IS SENT IF PORT READY, AND BIT 7 OF CHARACTER IN BUFFER
;   IS SET INDICATING TO CALLING PROGRAM THAT CHARACTER HAS BEEN SENT
;
;
TXCH:   IN      A,(LNSTAT)      ;READ THE LINE STATUS REGISTER
        BIT     5,A             ;IS THE PORT READY FOR A NEW CHAR NZ = YES
        RET     Z               ;RETURN IF PORT NOT READY
        LD      A,(HL)          ;GET NEW CHAR IN A REG
        OUT     (DBUFF),A       ;WRITE CHAR TO THE PORT
        SET     7,(HL)          ;IN TIMEX/SINCLAIR THIS SETS INVERSE VIDEO
        RET                     ;DONE: EXIT
;
;
```

The Line Control register sets the number of data bits, stop bits, and various error checking options. This register also allows access to the Divisor latches for setting the baud clock, which runs at 16 times the actual baud rate.

The Line Status register is read during transmit to determine when the chip is ready for the next data byte to be sent to the Transmit register. It is read during receive to determine when a character is ready to be read from the Receive buffer. Various errors are also shown in this register.

The Modem Control register determines the output of four Modem Output pins. These could be used for such functions as Press-To-Talk, Keying, CW ID, or CW keying. (My program used only the Press-To-Talk function.) The status of four Modem Input pins can be read from the Modem Status register. One of these could be used to detect a decoded CW Mark (key down) signal.

active filter improves operation

The second piece of hardware required is the Terminal Unit. My first circuit used an XR 2211 phase locked loop demodulator. Receiver output was fed directly to the demodulator. It was immediately apparent that filtering of the signal before the demodulator would be necessary for satisfactory operation under normal QRM and QRN conditions. An

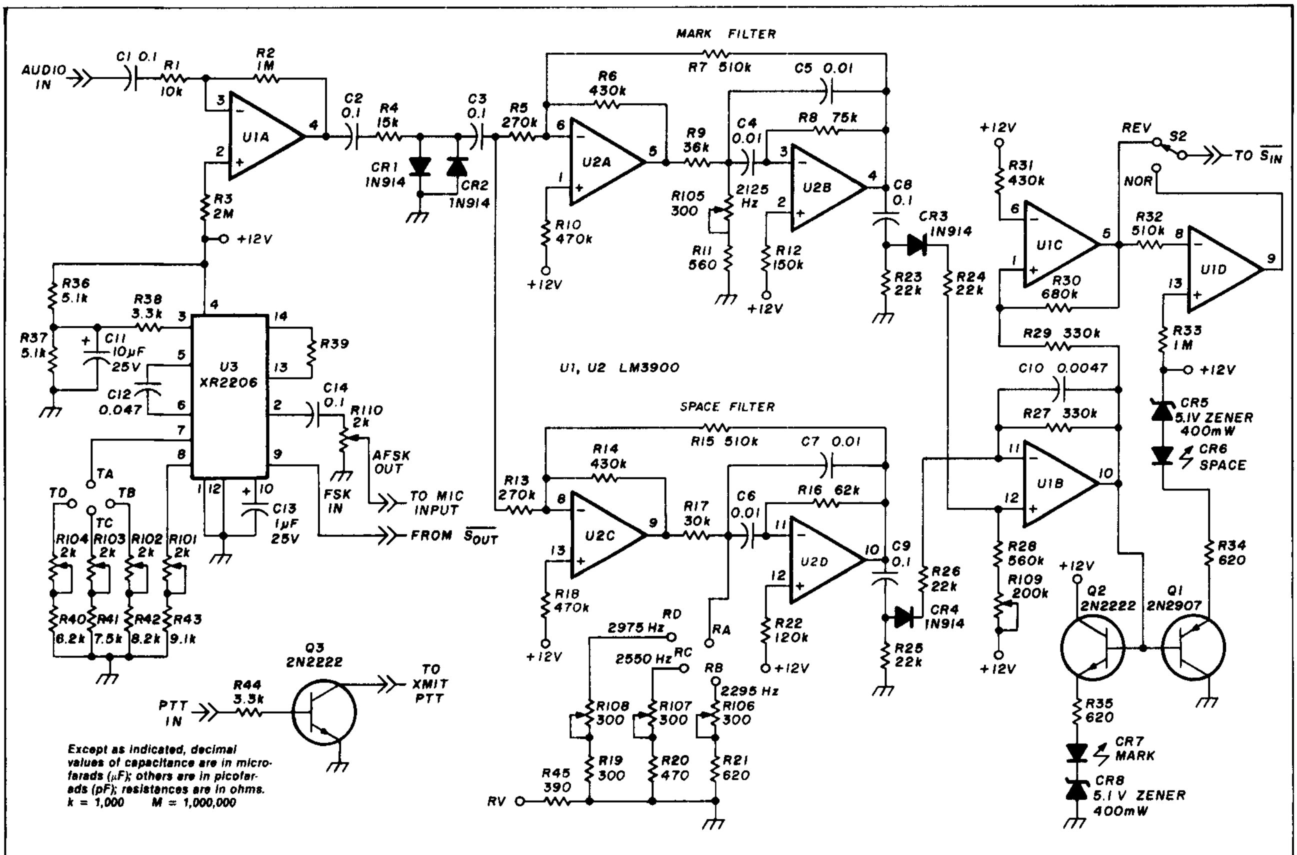

fig. 2. Terminal unit schematic.

active filter circuit for 170 Hz shift was no problem, but the circuitry became complex when provisions were added for copying signals at 425 Hz and 850 Hz shifts as well.

The design shown uses an active filter-type demodulator. It represents a reasonable compromise between performance and circuit complexity. A familiar XR 2206 circuit is used to generate the transmit AFSK tones.

operation at the I/O port

The schematic for the I/O port is shown in **fig. 1**. Power for the unit and signals from the computer are taken from the computer expansion bus. Charging of the 1.0 μF capacitor connected to pin 35 of the INS 8250 provides a power-on reset function. The I/O port would normally occupy eight consecutive port addresses, but this is not possible because of the Timex/Sinclair port addressing scheme. Port addresses for the various INS 8250 registers are given in **table 1**.

Read, Write, I/O Request signals, and Address lines A2 and A5 are decoded by the 74LS138 to make the I/O R and I/O W signals. INS 8250 select signals are furnished by Address lines A0, A1, and A6. The remaining Address lines A3, A4, and A7 control the register addressing. One half of the 74LS74 is used to divide the 3.25 MHz computer clock by two, to make the 1.625 MHz clock for the INS 8250. The three 2N2222 transistors should provide adequate protection for this expensive chip.

construction of the I/O port

This unit can be built on an etched circuit board or on perfboard with point-to-point wiring. The finished unit is installed between the computer and the 16K memory pack. An edge connector with wire-wrap pins should be used to attach the I/O port to the computer expansion bus.

The edge connector is installed on the *foil* side of the board. Allow the pins to extend through the circuit board approximately 3/16 inch (4.76 mm). Bend the pins together slightly and solder them to a small extender board; this extender board duplicates the computer expansion bus and provides connections to the 16K memory pack.

Although not shown on the schematic, at least two capacitors of about 0.047 μF should be placed on the circuit board and connected from the +5 volt supply to ground for the purpose of filtering out switching transients. A 3.3 kilohm pull-up resistor may be required on the SOUT lead with some Terminal Units.

part of fig. 2

item	description
C1,C2,C3	
C8,C9,C14	0.1
C4,C5,C6,C7	0.01*
C10	0.0047
C11	10, 25V electrolytic
C12	0.047*
C13	1.0, 25V electrolytic
C15	1000, 35V electrolytic
CR1-CR4	1N914
CR5,CR8	5.1V zener
CR6,CR7	red LED
CR9	green LED
CR10-CR13	1N4002
P1	500 ohm panel-mount potentiometer
Q1	2N2907
Q2,Q3	2N2222
R1	10 kilohm
R2,R33	1.0 megohm
R3	2.0 megohm
R4	15 kilohm
R5,R13	270 kilohm
R6,R14,R31	430 kilohm
R7,R15,R32	520 kilohm
R8	75 kilohm
R9	36 kilohm
R10,R18	470 kilohm
R11	560
R12	150 kilohm
R16	62 kilohm
R17	30 kilohm
R19	300
R20	470
R21,R34,R35	620
R22	120 kilohm
R23,R24, R25,R26	22 kilohm
R27,R29	330 kilohm
R28	560 kilohm
R30	680 kilohm
R36,R37	5.1 kilohm
R38,R44	3.3 kilohm
R39	220 kilohm
R40	6.2 kilohm
R41	7.5 kilohm
R42	8.2 kilohm
R43	9.1 kilohm
R45	390
R46	1.5 kilohm
R101-R104,R110	2 kilohm
R105-R108	300 kilohm
R109	200 kilohm
S1	SPST switch
S2,S4	2PDT switch
S3	2P3POS wafer switch
T1	12.6V, 300 mA Radio Shack 273-1385
U1,U2	LM3900
U3	XR2206
U4	7812 volt reg (TO-220 case)
case	Radio Shack 270-252 or -272

*Denotes Mylar. Recommended these be 5 percent "V" series.

Digi-key part No. P4513 for the 0.01 μF
P4521 for the 0.047 μF

The layout on the PC board is for the pin spacing of 5/16 inch preset variable resistors:

Digi-key part No. K4A32 for 300 ohm
K4A23 for 2 kilohm
K4A25 for 200 kilohm

All other resistors 1/4 watt, 5 percent tolerance
Miscellaneous plugs, jacks and other hardware is necessary.
All capacitance is in microfarads.

(Values below are for low tone operation (see text).

R8	120 kilohm	R21	1.2 kilohm
R9	62 kilohm	R22	180 kilohm
R11	1.1 kilohm	R40	9.1 kilohm
R12	240 kilohm	R41	11 kilohm
R16	91 kilohm	R42	13 kilohm
R17	47 kilohm	R43	16 kilohm
R19	470	RV	510
R20	820	P1	1 kilohm

operation of the terminal unit

The schematic for the Terminal Unit is shown in **fig. 2**. **Figure 3** illustrates the circuit for the TU power supply. Wiring for the associated control and switching components is shown in **fig. 4**. UC 2A and 2B and their associated components make up the Mark filter which is tuned by R105. U2C and 2D and associated components comprise the Space filter. R106, R107, and R108 are switched to provide tuning for the 170 Hz, 425 Hz, and 850 Hz fixed shifts. A panel mounted control is switched into the circuit for the variable shift tuning.

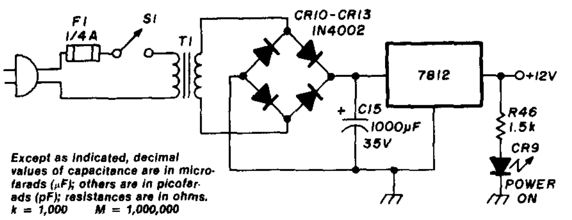

fig. 3. TU power supply schematic.

Audio from the receiver speaker is fed into amplifier U1A and limited by diodes CR1 and CR2 to minimize effects of a building or fading signal. Good limiting action occurs with about 30 mV applied to the Audio In jack.

The limiter output is coupled to both active filters by C3. The outputs of the filters are rectified and applied to differential amplifier U1B. The signal from the Mark filter is applied to the noninverting input and the signal from the Space filter is applied to the inverting input. The output of U1B will, therefore, go high for a Mark and low for a Space. Capacitor C10 filters the individual tone frequencies from U1B's output.

Positive feedback applied to U1C "squares up" the output from the differential amplifier. U1D is an inverter for copying Normal and reverse signals. Driver transistors Q1 and Q2 provide power for the Space and Mark indicator LEDs CR6 and CR7. Zener diodes CR5 and CR8 furnish bias to keep the Mark indicator turned off when the output of U1B is less than approximately 6 volts and to keep the Space indicator off when the U1B ouput is above the 6 volt midpoint.

Function generator U3 provides the Mark and Space tones for transmit. Timing capacitor C12, together with the resistance connected to pin 7 or pin 8 determine the tone frequency. When pin 9 is low, frequency is controlled by the resistance connected to pin 8. When pin 9 is high, frequency is controlled at pin 7.

Variable resistor R101 tunes the Mark frequency. Variable resistors R102, R103, and R104 are switched to set the Space frequency for 170, 425, or 850 Hz. With the value shown for R38, the output level at pin 2 is about 0.2 volts. The output level can be increased by increasing the value of R38. A value of 50 kilohm will give an output of about one volt.

Frequencies of 2550 and 2975 are seriously attenuated in modern HF transceivers. If your operation is primarily on the HF bands you may prefer a Mark tone of 1275 and Space tones of 1445, 1700, and 2125 Hz. Component value changes for low tone operation are noted on the parts list.

constructing the terminal unit

While prototype was constructed on perfboard,

using the etched circuit board will save time and result in a neater unit. Install components, beginning with those that lie flat on the board and finishing with the power transformer. Refer to **fig. 4** for connecting the components that are not mounted on the circuit board.

I brought the 110 volt power line in from the back of the unit and mounted the fuse and jacks on the back of the case. The switches, LEDs, and the variable shift tuning control were installed on the front panel. Because it is necessary to watch both the Mark and Space LEDs when tuning in a signal, they should be mounted reasonably close together — not more than 2 inches apart.

Some older model transmitters use a tube to key the Press-To-Talk relay. In this case use a small reed relay (Radio Shack 275-233) connected between the collector of Q3 and +12 volts to key the transmitter Press-To-Talk circuit.

tuning and aligning the terminal unit

Be sure to do the following before installing the integrated circuits or applying power:

- Check for shorts or any abnormally low resistance reading across C15.
- Recheck the polarity of all diodes, electrolytic capacitors, and proper orientation of transistors.
- Turn on the power.
- Confirm that the Power On LED lights up.
- Check for 12 volts at the proper pins of each socket.

Then turn off the power and install the ICs, observing proper pin 1 orientation. Turn on the power and check the ICs, electrolytic capacitors, and the voltage regulator for heat. Although the voltage regulator may be slightly warm, nothing should feel hot.

Ground the Audio In jack and observe the Mark and Space indicators. It should be possible to turn on either LED by adjustment of R109; adjust R109 to turn them off. In darkness, it should be possible to see a faint glow in both LEDs; use R109 to balance this indication.

Remove the ground from the Audio in jack and place it on the FSK In jack to simulate a Mark. Connect a frequency counter to the AFSK Out jack. Adjust output level control R110, if necessary, for a stable reading on the counter. Adjust R101 for a frequency reading of 2125 Hz.

Remove the ground from the FSK In jack, set S3 to N (narrow shift), and adjust R102 for a reading of 2295 Hz. Set S3 to M and adjust R103 for a reading of 2550 Hz. With S3 in W, adjust R104 for a frequency reading of 2975 Hz.

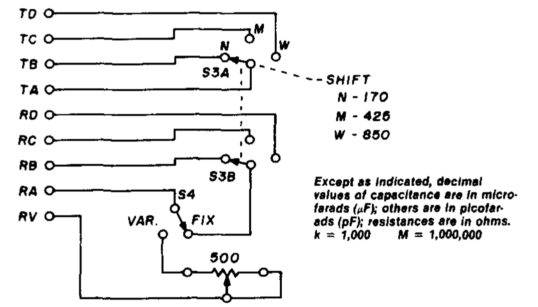

fig. 4. TU controls schematic.

Ground the FSK In jack again, and connect the AFSK Out jack to the Audio In jack. Tune the Mark filter to resonance by adjusting R105 for maximum brilliance of the Mark LED. If necessary, reduce the AFSK output level with R110 to achieve a sharp tuning peak.

Set switch S4 to FIX. Remove the ground from the FSK In jack, set S3 to N, and adjust R106 for maximum brilliance of the Space LED. With S3 set to M, adjust R107 for maximum brilliance again of the Space LED. Repeat this adjustment with S3 set to W using R108. Place S4 to VAR and S3 to M.

Adjust the variable shift tuning control for maximum brilliance of the Space LED. Repeat this check in N and in W positions. If the knob on the variable tuning control is set at the 12 o'clock position for 425 Hz shift, the 170 Hz tuning will be at about the 8:30 position and the 850 Hz tuning will be at about the 3:30 position.

Ground the FSK In jack once again. Set R110 at about its midpoint and place S2 to M. Measure the output of the Mark filter at the anode of CR3. Remove the ground from the FSK In jack and measure the output of the Space filter at the anode of CR4.

If the output of either filter is less than 8 volts peak-to-peak or 2.8 volts RMS at resonance, decrease R5 in the case of the Mark filter or R13 in the case of the Space filter, to a value of 240 kilohms. It should not be necessary to go below 240 kilohms if 5 percent or better tolerance parts have been used.

software notes

Table 1 shows example routines for initializing the port, reading data from the port, and sending data to the port. The formula for determining the Divisor is:

$$Divisor = \frac{1.625 \times 10^6}{baud\ rate \times 16}$$

For 60 WPM RTTY (45.45 Baud):

$$\frac{1.625 \times 10^6}{45.45 \times 16} \cong 2235$$

The Divisor Most Significant Byte is (in BASIC):

LET DMSB = INT (DIVISOR/256)

The Divisor Least Significant Byte is:

*LET DLSB = DIVISOR − DMSB * 256*

For 45.45 Baud:

Divisor Least Significant Byte = 187

Divisor Most Significant Byte = 8

The port address for the Line Control register is AF(hex). When bit 7 of this register is set, the Divisor Latches are enabled and port address 27(hex) allows the Divisor Least Significant Byte to be entered. Address A7(hex) allows the Divisor Most Significant Byte to be entered. When bit 7 of the Line Control register is reset, port 27 (hex) addresses the Transmit and Receive buffers. Line Control register bits 0 and 1 control the number of data bits per character. Bit 2 determines the number of stop bits.

initializing the I/O port

1. Set bit 7 of Line Control register to gain access to the Divisor Latches.

2. Enter Divisor bytes.

3. Reset bit 7 of Line Control register and set up for desired mode.

4. Disable interrupts or set up for desired interrupt mode.

5. Set any desired output conditions with the Modem Control register.

reading I/O port

1. Read the Line Status register. Bit 0 will be set if a character is ready in the Receive buffer.

2. If a character is ready, read it from the Receive buffer. Bit 0 in the Line Status register will automatically be reset when the character is read.

writing to the I/O port

1. Read the Line Status register. Bit 5 will be set if the port is ready for a new character.

2. If the port is ready, write the character to the Transmit buffer.

The INS 8250 I/O port provides a simple solution for obtaining stable, software-selectable baud rates. I think the Terminal Unit design represents a reasonable compromise between circuit complexity and performance. Notes have been given on software design to help those interested gain a better understanding of Input/Output programming.

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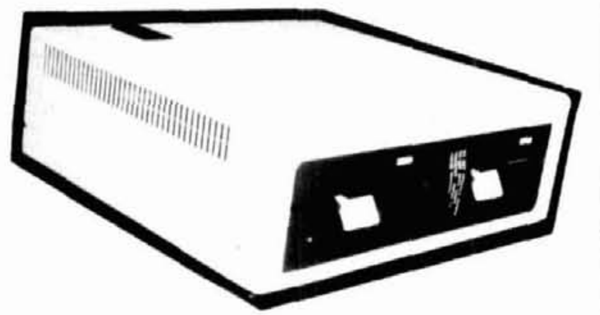

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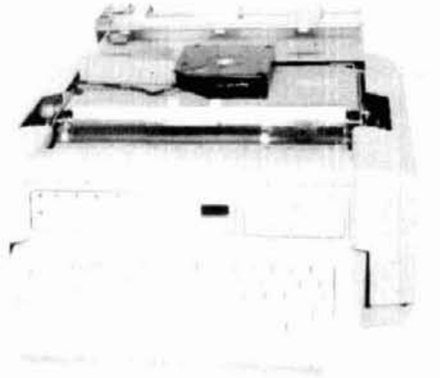

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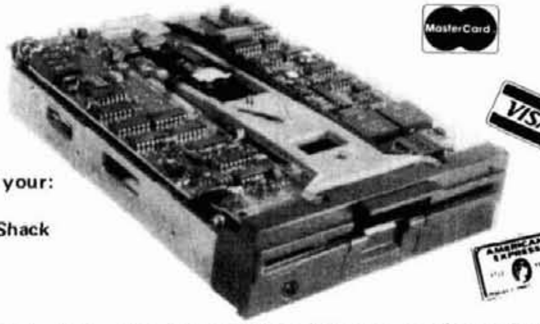

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2102L-4	1024 x 1 (450ns) (LP)	.89
2102L-2	1024 x 1 (250ns) (LP)	1.29
2111	256 x 4 (450ns)	2.29
2112	256 x 4 (450ns)	2.29
2114	1024 x 4 (450ns)	.99
2114-25	1024 x 4 (250ns)	1.10
2114L-4	1024 x 4 (450ns) (LP)	1.20
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2114L-2	1024 x 4 (200ns) (LP)	1.40
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TMS4044-3	4096 x 1 (300ns)	3.45
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TMM2016-20	2048 x 8 (200ns)	3.95
TMM2016-15	2048 x 8 (150ns)	4.45
TMM2016-10	2048 x 8 (100ns)	5.95
HM6116-4	2048 x 8 (200ns) (cmos)	4.40
HM6116-3	2048 x 8 (150ns) (cmos)	4.90
HM6116-2	2048 x 8 (120ns) (cmos)	5.90
HM6116LP-4	2048 x 8 (200ns) (cmos)(LP)	4.90
HM6116LP-3	2048 x 8 (150ns) (cmos)(LP)	5.90
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HM6264P-15	8192 x 8 (150ns) (cmos)	19.95
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HM6264LP-12	8192 x 8 (120ns) (cmos)(LP)	24.95

LP = Low Power   Qstat = Quasi-Static

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4164-20	65536 x 1 (200ns) (5v)	2.22
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5V = Single 5 Volt Supply

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2732	4096 x 8 (450ns) (5v)	3.95
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2732 A-35	4096 x 8 (350ns) (21v)	3.95
2732 A	4096 x 8 (250ns) (21v)	5.95
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MCM68766	8192 x 8 (350ns) (5v) (24-pin)	19.95
27128-45	16384 x 8 (250ns) (5v)	14.95
27128-30	16384 x 8 (300ns) (5v)	16.95
27128-25	16384 x 8 (250ns) (5v)	18.95
27256-25	32768 x 8 (250ns) (14v)	79.95

5v = Single 5 Volt Supply

74LS00

74LS00	.23	74LS125	.48	74LS260	.58
74LS01	.24	74LS126	.48	74LS266	.54
74LS02	.24	74LS132	.58	74LS273	1.45
74LS03	.24	74LS133	.58	74LS275	3.30
74LS04	.23	74LS136	.38	74LS279	.48
74LS05	.24	74LS137	.98	74LS280	1.95
74LS08	.27	74LS138	.54	74LS283	.68
74LS09	.28	74LS139	.54	74LS290	.88
74LS10	.24	74LS145	1.15	74LS293	.88
74LS11	.34	74LS147	2.45	74LS295	.98
74LS12	.34	74LS148	1.30	74LS298	.88
74LS13	.44	74LS151	.54	74LS299	1.70
74LS14	.58	74LS153	.54	74LS323	3.45
74LS15	.34	74LS154	1.85	74LS324	1.70
74LS20	.24	74LS155	.68	74LS352	1.25
74LS21	.28	74LS156	.68	74LS353	1.25
74LS22	.24	74LS157	.64	74LS363	1.30
74LS26	.28	74LS158	.58	74LS364	1.90
74LS27	.28	74LS160	.68	74LS365	.48
74LS28	.34	74LS161	.64	74LS366	.48
74LS30	.24	74LS162	.68	74LS367	.44
74LS32	.28	74LS163	.64	74LS368	.44
74LS33	.54	74LS164	.68	74LS373	1.35
74LS37	.34	74LS165	.94	74LS374	1.35
74LS38	.34	74LS166	1.90	74LS377	1.35
74LS40	.24	74LS168	1.70	74LS378	1.13
74LS42	.48	74LS169	1.70	74LS378	1.30
74LS47	.74	74LS170	1.45	74LS385	1.85
74LS48	.74	74LS173	.68	74LS386	.44
74LS49	.74	74LS174	.54	74LS390	1.15
74LS51	.24	74LS175	.54	74LS393	1.15
74LS54	.28	74LS181	2.10	74LS395	1.15
74LS55	.28	74LS189	8.90	74LS399	1.45
74LS63	1.20	74LS190	.88	74LS424	2.90
74LS73	.38	74LS191	.88	74LS447	.38
74LS74	.34	74LS192	.78	74LS490	1.90
74LS75	.38	74LS193	.78	74LS624	3.95
74LS76	.38	74LS194	.68	74LS640	2.15
74LS78	.48	74LS195	.68	74LS645	2.15
74LS83	.59	74LS196	.78	74LS668	1.65
74LS85	.68	74LS197	.78	74LS669	1.85
74LS86	.38	74LS221	.88	74LS670	1.45
74LS90	.54	74LS240	.94	74LS674	9.60
74LS91	.88	74LS241	.98	74LS682	3.15
74LS92	.54	74LS242	.98	74LS683	3.15
74LS93	.54	74LS243	.98	74LS684	3.15
74LS95	.74	74LS244	1.25	74LS685	3.15
74LS96	.88	74LS245	1.45	74LS688	2.35
74LS107	.38	74LS247	.74	74LS689	3.15
74LS109	.38	74LS248	.98	74LS783	23.95
74LS112	.38	74LS249	.98	81LS95	1.45
74LS113	.38	74LS251	.58	81LS96	1.45
74LS114	.38	74LS253	.58	81LS97	1.45
74LS122	.44	74LS257	.58	81LS98	1.45
74LS123	.78	74LS258	.58	25LS2521	2.75
74LS124	2.85	74LS259	2.70	25LS2569	4.20

CRT CONTROLLERS

6845	11.95	8275	28.95
68B45	18.95	7220	38.95
HD46505sp	11.95	CRT5027	18.95
6847	10.95	CRT5037	28.95
MC1372	6.90	TMS9918A	36.95
68047	23.95	DP8350	48.95

DISC CONTROLLERS

1771	14.95	2797	54.95
1791	21.95	6843	33.95
1793	25.95	8272	18.95
1795	21.95	UPD765	18.95
1797	21.95	MB8876	23.95
2791	49.95	MB8877	25.95
2793	49.95	1691	6.95
2795	54.95	2143	6.95

UV ERASERS

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ECONOMY Model

- Erases 15 EPROMS in 20 minutes
- Plastic Enclosure

6500

1 MHZ

65022	4.90
6504	6.90
6505	8.90
6507	9.90
6520	4.30
6522	4.90
6532	9.90
6545	9.90
6551	9.90

2 MHz

6502A	5.90
6520A	5.90
6522A	9.90
6532A	10.90
6545A	12.90
6551A	10.90

3 MHz

6502B	7.90

6800

1 MHz

6800	2.90
6802	7.90
6803	17.90
6808	12.90
6809E	8.90
6809	8.90
6810	2.90
6820	4.30
6821	2.90
6828	13.90
6840	11.90
6843	33.90
6844	24.90
6845	11.90
6847	10.90
6850	2.90
6852	5.90
6860	7.90
6862	10.90
6875	6.90
6880	1.90
6883	21.90

68B00 2 MHz

68B00	9.90
68B02	11.90
68B09	11.90
68B09E	11.90
68B10	5.90
68B21	5.90
68B40	18.90
68B45	18.90
68B50	5.90

68000

68000-8	34.90
68047	23.90
68488	18.90
68652	14.90
68661	8.90
68764	17.95
68766	19.95

8000

8031	14.90	8253	6.90
8035	5.90	8253-5	7.90
8039	5.90	8255	4.45
INS-8060	16.90	8255-5	4.90
INS-8073	29.90	8257	7.90
8080A	3.90	8257-5	8.90
8085	4.90	8259	5.90
8085A-2	11.90	8259-5	6.90
8086	24.90	8271	69.90
8087-3 (5 MHz)	159.90	8272	19.90
8087-2 (8MHz)	279.90	8274	28.90
8088	19.90	8275	28.90
8089	59.90	8279	6.90
		8279-5	7.90
		8282	6.45
		8283	6.45
		8284	4.90
		8286	6.45
		8287	6.45
		8288	12.90
		8289	44.90
		8292	12.90

8100

8131	2.90
8155	6.90
8155-2	7.90
8156	6.90
8185	28.90
8185-2	38.90

8200

8202	23.90
8203	38.90
8205	2.90
8212	1.75
8214	3.75
8216	1.75
8224	2.20
8226	1.75
8228	3.45
8237	12.90
8237-5	14.90
8238	4.45
8243	4.45
8250	9.90
8251	3.90
8251A	4.45

8300

8303	2.90
8304	1.90
8307	2.90
8308	2.90
8310	3.90
8311	3.90

8700

8741	28.90
8748	19.90
8749	28.90
8755	23.90

80000

80186-6	99.90
80188	89.90

Z-80

2 MHz

Z80-CPU	1.95
Z80-CTC	1.95
Z80-DART	6.95
Z80-DMA	7.95
Z80-PIO	1.95
Z80-SIO/0	8.95
Z80-SIO/1	8.95
Z80-SIO/2	8.95
Z80-SIO/9	8.95

Z80A-SIO/0	9.95
Z80A-SIO/1	9.95
Z80A-SIO/2	9.95
Z80A-SIO/9	9.95

4 MHz

Z80A-CPU	2.45
Z80A-CTC	2.45
Z80A-DART	7.95
Z80A-DMA	8.95
Z80A-PIO	2.45

6 MHz

Z80B-CPU	7.95
Z80B-CTC	8.95
Z80B-PIO	8.95
Z80B-DART	18.95
Z80B SIO/0	28.95
Z80 SIO/2	28.95

ZILOG

Z6132	29.95
Z8671	38.95

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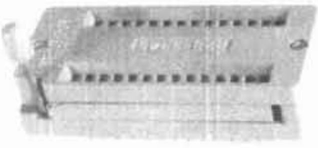
4164 200ns
9 for $19.98

DIP SWITCHES

4 POSITION	.79
5 POSITION	.85
6 POSITION	.85
7 POSITION	.89
8 POSITION	.89

ZIF SOCKETS



14 pin ZIF	4.89
16 pin ZIF	4.89
24 pin ZIF	5.89
28 pin ZIF	6.89
40 pin ZIF	9.89

ZIF = TEXTOOL (Zero Insertion Force)

IC SOCKETS
(1 to 99)

8 pin ST.	.12	8 pin WW	.58
14 pin ST.	.14	14 pin WW	.68
16 pin ST.	.16	16 pin WW	.68
18 pin ST.	.19	18 pin WW	.98
20 pin ST.	.28	20 pin WW	1.04
22 pin ST.	.29	22 pin WW	1.34
24 pin ST.	.29	24 pin WW	1.44
28 pin ST.	.39	28 pin WW	1.64
40 pin ST.	.48	40 pin WW	1.94

ST = Soldertail   WW = Wirewrap

CRYSTAL CLOCK OSCILLATORS

PART NO.	FREQUENCY	PRICE
1.000	1.0000 MHz	6.99
1.843	1.8432 MHz	6.99
2.000	2.0000 MHz	6.99
4.000	4.0000 MHz	6.99
8.000	8.0000 MHz	6.99
10.000	10.0000 MHz	6.99
16.000	16.0000 MHz	6.99
18.432	18.4320 MHz	6.99
19.660	19.6608 MHz	6.99
20.000	20.0000 MHz	6.99
32.000	32.0000 MHz	6.99

CRYSTALS

1.0000 MHz	2.69	8.0000 MHz	1.95
1.8432 MHz	2.69	10.0000 MHz	1.95
2.0000 MHz	1.95	10.7386 MHz	1.95
2.0972 MHz	1.95	12.0000 MHz	1.95
2.4576 MHz	1.95	14.3182 MHz	1.95
3.2768 MHz	1.95	15.0000 MHz	1.95
3.5795 MHz	1.95	16.0000 MHz	1.95
4.0000 MHz	1.95	17.4300 MHz	1.95
4.1943 MHz	1.95	18.0000 MHz	1.95
4.9160 MHz	1.95	18.4320 MHz	1.95
5.0000 MHz	1.95	19.6608 MHz	1.95
5.0688 MHz	1.95	20.0000 MHz	1.95
5.1850 MHz	1.95	22.1184 MHz	1.95
5.2429 MHz	1.95	32.0000 MHz	1.95
5.7143 MHz	1.95	36.0000 MHz	1.95
6.0000 MHz	1.95	48.0000 MHz	1.95
6.1440 MHz	1.95	49.4350 MHz	1.95
6.4000 MHz	1.95	49.8900 MHz	1.95
6.5536 MHz	1.95		

32.768 KHz .... .99

VOLTAGE REGULATORS

7805T	.74	7905T	.84
78M05C	.34	7908T	.84
7808T	.74	7912T	.84
7812T	.74	7915T	.84
7815T	.74	7924T	.84
7824T	.74	7905K	1.44
7805K	1.34	7912K	1.44
7812K	1.34	7915K	1.44
7815K	1.34	7924K	1.44
7824K	1.34	79L05	.78
78L05	.68	79L12	.78
78L12	.68	79L15	.78
78L15	.68	LM323K	4.90
78H05K	9.90	UA78S40	1.90
78H12K	9.90		

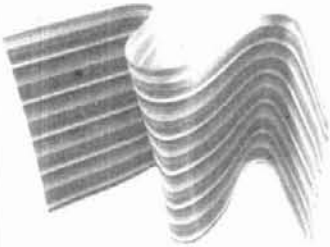
C,T = TO-220   K = TO-3
L = TO-92

ROBOT KITS!
PIPER-MOUSE

Controlled by sound sensor and 1-channel electronic circuit. Use the whistle in this kit and Piper-Mouse follows your commands, turning left or right, stopping and starting. Uses 2 AA and 1 9V battery (not included).

MV-915 $44.95

RIBBON CABLE



	SINGLE COLOR		COLOR CODED	
CONTACTS	1'	10'	1'	10''
10	.45	4.30	.78	7.20
16	.50	4.70	.95	8.70
20	.60	5.60	1.15	10.90
25	.70	6.50	1.22	11.50
26	.70	6.50	1.27	11.50
34	.93	8.50	1.55	14.40
40	1.27	11.50	1.82	16.70
50	1.28	12.00	2.40	21.90

RESISTORS

¼ WATT 5% CARBON FILM
ALL STANDARD VALUES
FROM 1 OHM – 10 MEG OHM

50 PCS	1.25
100 PCS	2.00
1,000 PCS	15.00

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MULTIVIEW 80/160 249.00

80-160 columns with any monitor!

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Cooling Fan	38.95
Power Supply	74.95
Joystick	29.95
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Disk Drive	169.95
Controller Card	59.95
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VIEWMAX-80 149.95

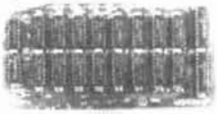
80-Column card for Apple II series

- Video Soft Switch
- Inverse Video
- VIDEX's Videoterm compatible

VIEWMAX-80e 119.95

80-Column extended video card for Apple IIe

- 64K RAM, expandable to 128K
- Double High-resolution circuit
- Compatible with Pascal & CP/M

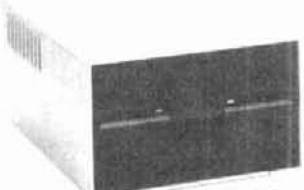


PRINTMAX 59.95

Parallel printer card, Apple II series

- Centronics compatible
- Variable print widths
- Up to 5000 characters/second

APPLE & IBM Compatible DISK DRIVES



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- Shugart mechanism, made in U.S.A.
- Directly replaces Apple Disk II
- Fully compatible with Apple Controller or other Apple compatible controllers
- One Year Warranty

FULL or ½-Height

16K RAM Card – Apple II+

- 2-Year Warranty

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APPLE or IBM JOYSTICK

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Compatible for either:
APPLE II and APPLE IIe
OR
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MULTIFUNCTION CARD

- 64K to 384K RAM
- Parallel Port
- Serial Port
- Clock Calendar
- Software Included
- 1-Year Warranty

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- Expandable to 512K
- Fully compatible with IBM software
- Fully compatible w/IBM diagnostic utilities
- Serial Port Available
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MEMORY EXPANSION KIT

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DISKETTES

5¼''

ATHANA: (SOFT SECTOR w/HUB RING)

25 per package

SS/DD...... 29.75 or 1.19 ea.
DS/DD...... 34.75 or 1.39 ea.

LIFETIME WARRANTY ON ALL ATHANA DISKETTES

NO LABEL: (SOFT SECTOR w/HUB RING)

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SS/DD...... 24.75 or .99 ea.
DS/DD...... 29.75 or 1.19 ea.

2-YEAR WARRANTY ON ALL BULK DISKETTES

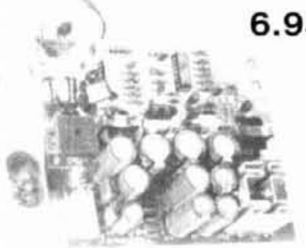
ROBOT KITS!
PEPPY

2-way sensor detects noise or solid objects in its path. When front sensor contacts an obstacle or hears a loud noise (hand-clap), Peppy automatically turns to the left. Uses 2 AA and 1 9V battery (not included).

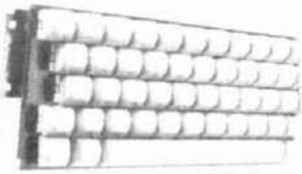
MV-916 $24.95

Reg. Power Supply
Model 4A/PS (99/4)
3 DC Outputs:
12V @ .4A, +5V @ 1.1A
−5V @ .2A Highly Filtered

6.95



KEYBOARD (99/4)



48 keys 4''x10'' **6.95**

TERMS: Minimum order $10.00. For shipping and handling, include $2.50 for UPS ground or $3.50 for UPS Blue (air). For each additional air pound, add $1 for UPS Blue shipping and handling. California residents must include 6% sales tax; Bay area and LA residents include 6½% sales tax. Prices are subject to change without notice. We are not responsible for typographical errors. We reserve the right to limit quantities and to substitute manufacturers. All merchandise subject to prior sale.

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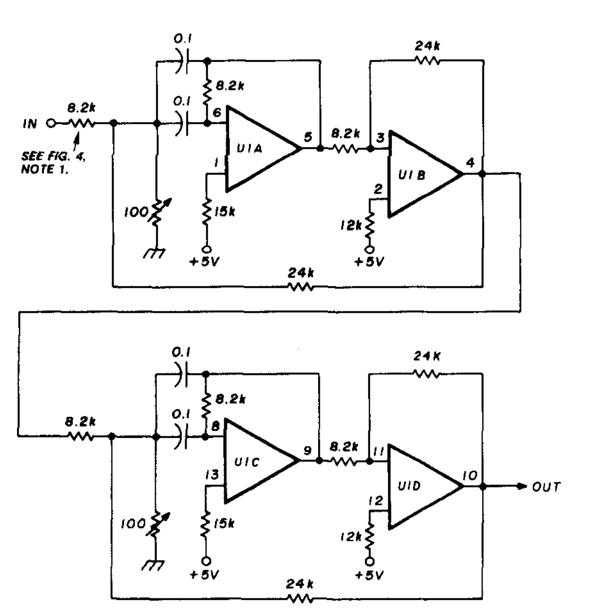
DoKay

run RTTY
on your VIC-20

Interface allows VIC-20 to command transceiver for CW and RTTY operation

It all started at a meeting of our ham club, when KB4SM sold me a Kantronics *Hamsoft*™ cartridge (He had just upgraded to *Hamtext*™.) So I plugged *Hamsoft* into my VIC-20 and rigged my key to see how the unit would print my CW. I found it to be very critical of my fist, with such aberrations as "the" coming out as "6E." After rigging a diode to rectify the receiver audio, I found that it was just as critical of the other fellow's keying, so I felt better.



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

fig. 1. Schematic of bandpass filter (from W4AYV, *ham radio*, April, 1979, page 47.

Next came the RTTY, which was the main purpose of the project. I used an LM567 chip, and sure enough, it printed out some of the stuff that I tuned in. At this point it became only too obvious that extra selectivity was essential, so I looked in some back issues of *ham radio* and found an active bandpass filter design[1] using an LM3900 chip that I had picked up at the Memphis Hamfest (**fig. 1**). With these two units combined, reception was quite tolerable. I hooked a pair of high-impedance headphones at the output of the bandpass filter to aid in tuning in the signal.

The next consideration was transmission of RTTY and CW. National Semiconductor's *Linear Handbook* provided a circuit for the AFSK oscillator. The output was found to be a square wave of several volts, peak-to-peak. This was far too much signal to substitute for the usual microphone output, so a three-section RC filter was introduced. In order to load the RC oscillator as lightly as possible, the input resistor was chosen to be about 39 kilohms, with the others being 12 kilohms. (These may be increased if the audio output should still be somewhat high.) Capacitors were all 0.01 μF ceramic. The output was now a nice looking sine wave, necessary for a clean signal.

circuit adjustments

To vary the frequency of the AFSK oscillator, a smaller capacitor is switched in and out of the frequency-determining circuit by a 2N2222 transistor. When used in this function, no DC voltage is applied to the collector of the transistor.

The AFSK oscillator frequency is given as the reciprocal of the RC product. A capacitor of 0.068 μF was combined with a resistor of 6.8 kilohms. The smaller capacitor that is switched for frequency

By Henry S. Keen, W5TRS, Fox, Arkansas 72051

change is 0.0056 μF. If you have a frequency counter available, the AFSK oscillator output may be established and adjusted, if necessary. It is more important to obtain a frequency shift of 170 Hz, when using an SSB transceiver, because corrections are automatic as the signal is tuned in.

The frequency of the bandpass filter is now adjusted to peak at the frequency of the AFSK oscillator by means of the two 100-ohm trim pots.

Next, the position of the potentiometer controlling the PLL frequency of the decoder, when its frequency coincides with that of the AFSK oscillator, must be determined. This is done by applying DC to both chips and either comparing the signals on a scope or combining them through a temporary resistive network, and adjusting to zero beat. This is the normal operating position. Any minor differences between your frequency and that of the other station will be compensated with this control, so that your transmitting frequency will not be affected.

For CW operation, a PNP transistor operates a reed relay that keys the transmitter directly. Although the transistor might do the job without the relay, I did not want to take a chance that some problem with the transmitter might damage the computer.

space and mark signals

The decoder chosen makes use only of the mark signal and ignores the space signal. Although marginal signals may better be handled by a decoder that makes use of both components, this is a satisfactory arrange-

table 1. Parts list.

quantity	description
1	LM3900 quad amplifier
2	LM567 tone decoders
1	14-pin socket
2	8-pin sockets
5	0.1 μF Mylar capacitors
5	0.01 μF ceramic disc capacitors
1	0.1 μF ceramic disc capacitor (can use Mylar
2	1.0μF tantalum capacitors
1	0.068 Mylar capacitor
1	0.0056 Mylar capacitor
2	1 kilohm 1/4-watt resistors
1	3 kilohm 1/4-watt resistor
6	8.2 kilohm 1/4-watt resistors
4	24 kilohm 1/4-watt resistors
4	12 kilohm 1/4-watt resistors
2	15 kilohm 1/4-watt resistors
1	6.8 kilohm 1/4-watt resistor
1	2N2222 transistor
1	2N3906 transistor
1	5-volt reed relay
1	LED
1	DPDT slide or toggle switch
3	miniature 2-circuit jacks
1	miniature 3-circuit jack
1	5 kilohm linear potentiometer
1	4 × 5 inch PC board, copper on one side
1	18-pin edge connector (optional)

suitable housing box for interface

appropriate cables and connectors to mate with transceiver

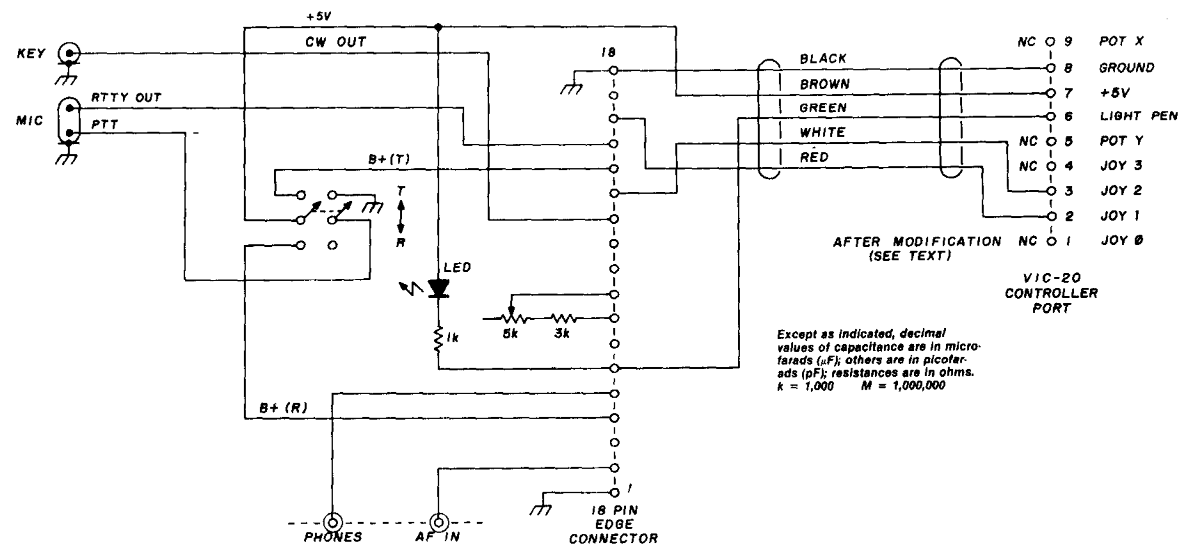


fig. 2. Additional interface wiring.

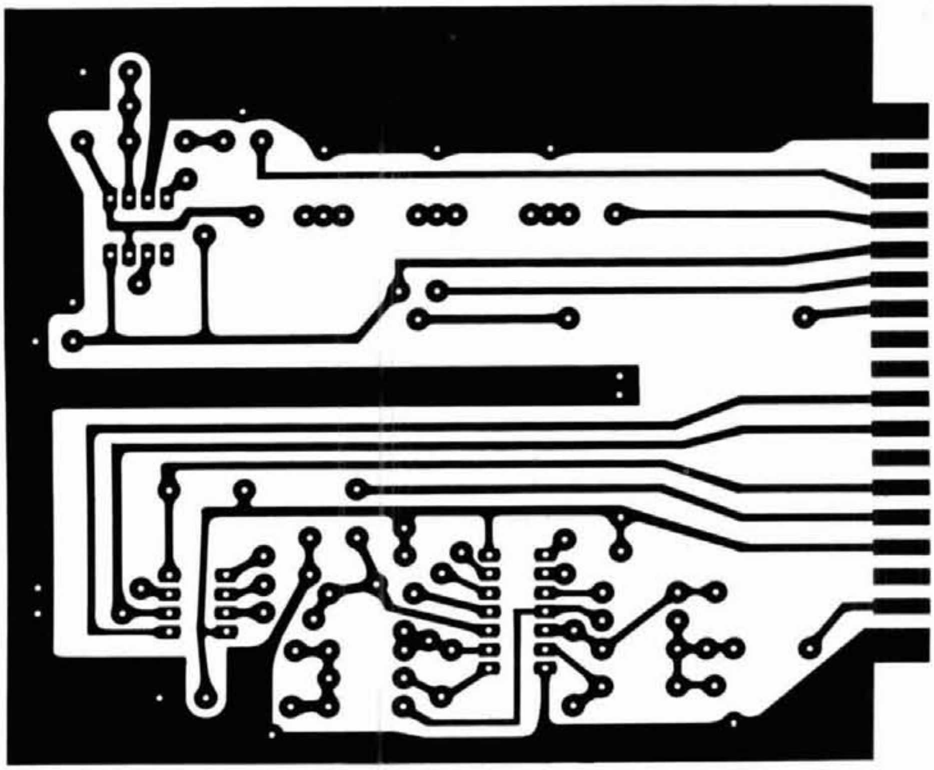


fig. 3. Foil side of PC board for the CW/RTTY.

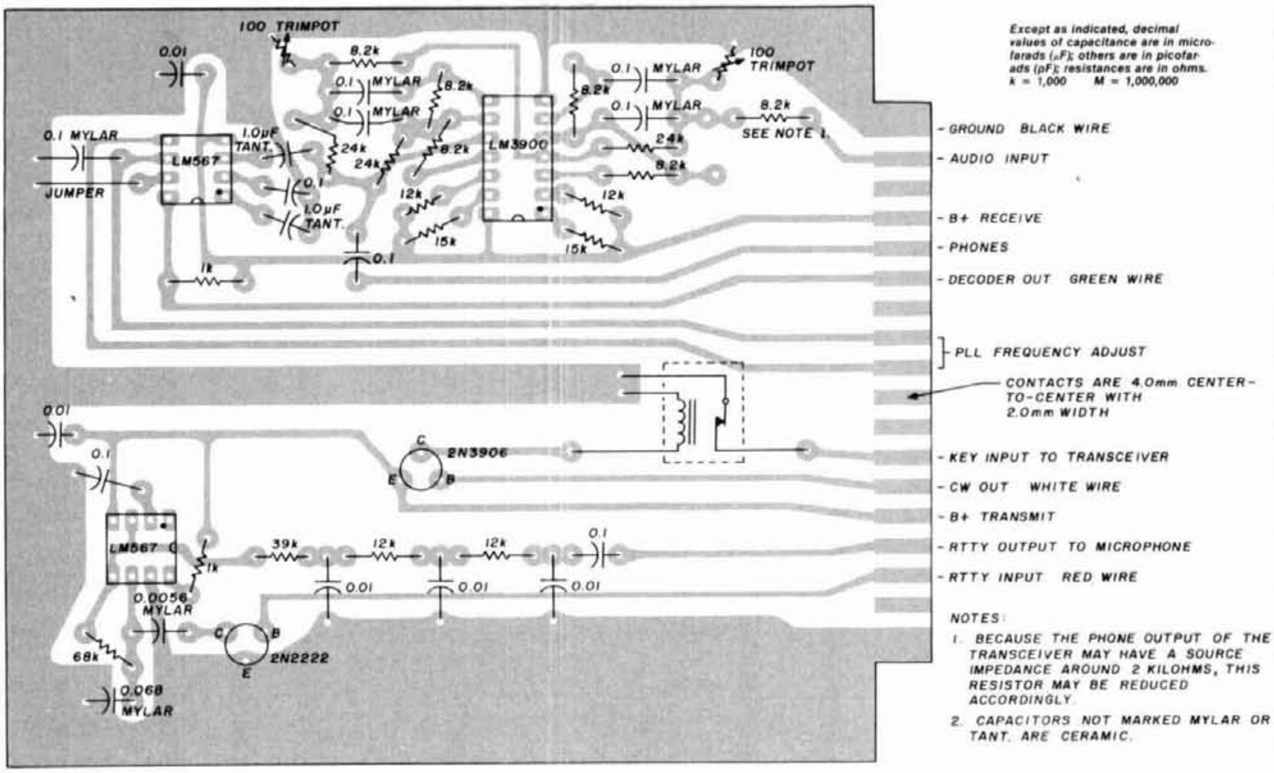


fig. 4. Component side of PC board for the CW/RTTY interface.

ment; signals that read R1 on the meter will usually give good print. A minor advantage appears with this decoder when widely different frequency shifts are encountered, because they present no problem to the mark-only decoder.

Most RTTY transmission and reception is on the LSB mode of the transceiver. In transceivers with provisions in the IF channel for a CW filter, this filter will usually be found to be tied into the USB mode. A number of stations operate this way, with inverted signals. Reception is easily corrected by adjusting the PLL frequency.

Accommodating transmission to the USB mode may be done in several ways. A friend assures me that if I press the K button instead of T, when activating the system, the transmission will not be inverted when on the USB mode. I haven't tried it, so I'm not sure. Another way would be to locate a transistor inverter stage between the decoder output and the computer light pen input terminal, and another in front of the NPN keyer transistor.

power supply notes

A separate power supply was originallly used for the interface, but one day the bandpass filter as well as the decoder and AFSK oscillator were run from a 5-volt common supply. Because little if any difference could be detected, 5 volts was used for the entire interface. The importance of this finding was that now the power to the interface could be supplied by the computer eliminating a separate DC source. The VIC-20 can supply +5 volts at up to 100 mA; the interface requires about 20 mA.

The cable that comes with the *Hamsoft* cartridge has five wires: black for ground, green for demodulator input, white for CW output, red for RTTY output, and brown for RTTY output (inverted). This last capability was exchanged for +5 volts supply, by removing the pin from hole No. 1 and moving it to hole No. 7. This probably eliminates the previously mentioned possibility of inverting the signal with the K for T change. However, the elimination of a separate power supply is well worth the effort.

additional wiring

Additional wiring included a DPDT slide switch for transmit or receive. The +5 volts is switched from the bandpass filter and decoder on receive to the AFSK oscillator on transmit. Also, the other section of the switch grounds the PTT line to the microphone connector of the transceiver. An LED is connected from the decoder output to the +5 volts, through a limiting 1 kilohm resistor. All of this extra wiring is shown in **fig. 2.**

The system also receives ASCII quite well. PLL adjustment on ASCII seems a bit more critical than on RTTY, but copy seems every bit as good. I have not tried ASCII on transmit because the slowest RTTY speed is still too fast for my typing ability.

The resistors used in the system are 5 percent. Any frequency determining capacitors should be Mylar or the equivalent. Ceramics are OK for bandpass, but I used the little Tantalum type wherever possible.

The system was mounted on a PC board, arranged to plug into an 18-pin edge connector, mounted in a 4 × 6-inch (10 × 15 cm) console-type box. Any PC board arrangement should do; perf board would be a bit messy.

The PC board as seen from both sides is shown in **figs. 3** and **4.** A parts list is provided in **table 1.** The reed relay as mounted on the PC board is shown in schematic form because these items vary considerably, and because I believe some freedom should be left to the individual builder.

Hamsoft™ and *Hamtext*™ are manufactured by Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66044.

references

1. *Linear Handbook*, National Semiconductor, 2900 Semiconductor Drive, Santa Clara, California 95051.
2. Nat Stinnetee, W4AYV, "Active Bandpass Filter for RTTY," *ham radio*, April, 1979, page 46.

ham radio

✓ 168

april 1985

short circuit

"smart" frequency counter

The parts list for **fig. 2** of WA5VQK's "Smart Frequency Counter" (October, 1984, page 41), omitted from the text of that article, reads as follows:

quantity	value/number	approximate cost
2	10k resistors	.02
3	470 resistors	.03
1	560 resistors	.01
8	75 resistors	.08
4	0.1 μF capacitors	.24
1	0.01 μF capacitor	.02
1	10 μF capacitor	.15
1	2 μF capacitor	.12
1	15-35 pF variable capacitor	1.00
6	2N3904 transistors	.36
1	SN74LS00 IC	.20
2	SN74LS390 ICs	3.00
1	MC14499P IC	3.32 *
1	MC68705P35 IC	27.26 **
1	4.0 MHz crystal	4.00
4	common cathode 7-segment displays	2.00
		$41.81

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More Details? CHECK—OFF Page 160

VHF/UHF WORLD

Joe Reisert W1JR

stacking antennas: part 1

There are two basic ways to obtain high antenna gain in the VHF/UHF spectrum. Either you build a high-gain antenna with a single feed (such as a parabolic dish or long Yagi), or you build a number of single-feed devices and array or stack them for higher gain. Parabolic dishes, which require only a single feed system, have been built with gains exceeding 60 dBi (dB above an isotropic radiator). (To convert dB over a dipole to dBi, add 2.15 dB.) Hence they are especially popular on EME, where high gain is necessary.

However, high gain dish type antennas can get quite large. For example, let's see what size dish would be necessary for EME. The minimum recommended antenna gain for 2-meter EME is 20 dBi. This would require a dish approximately 33 feet (10 meters) in diameter. The minimum recommended gain for 70 cm EME is 25 dBi, which would require an 18-foot (5.5 meter) diameter dish.[1] Furthermore, parabolic dishes are usually only 50 to 60 percent efficient and can present structural problems, especially for those locations where wind and snow are prevalent.

Yagis are replacing dishes

In recent years, the Yagi antenna has become very popular, particularly on 70 cm and lower frequencies. Its popularity is justly deserved because if properly designed, it can exceed 70 to 80 percent efficiency with only moderate wind load. A properly designed 20 dBi-gain Yagi would, however, require a boom length of about 13 wavelengths — 89 feet (27 meters) at 2 meters! A 30-foot (9 meter) 4.4 wavelength boom design is about the longest practical 2-meter Yagi, but it would have a gain of only about 16 to 17 dBi, 3 to 4 dB lower than desired for EME. Therefore, when high gains are required, two or more Yagi antennas are arrayed or stacked to obtain the required gain.

general principles in stacking

It is often said that every time you double the number of Yagis, you double (add 3 dB) the overall gain. It is also common to hear that the proper stacking distance for a Yagi is two thirds of the boom length. Are these statements true? *No.*

But don't lose heart. Since I'm all too often asked about proper stacking distance, I decided that it's time to update the material I've been distributing since my first talk on the subject (at the Central States VHF Conference in Kansas City in 1977) and present it here along with additional data.

This subject deserves more than just a set of charts or tables that quickly become obsolete as new designs appear. Also, there are many practical aspects of stacking that are often ignored. Therefore, I've decided to first discuss stacking concepts in depth.

Since the material required to thoroughly cover the subject of stacking is extensive, I've decided to devote two monthly columns to this topic. In **part 1** (this month), I'll discuss the theory of stacking and provide the examples, tables, and charts required for a first cut. **Part 2** (next month), will discuss the practical aspects of stacking and provide suggestions on how to properly use the material presented in part 1. At the conclusion of part 2, you should have all the necessary material to determine the proper stacking for any Yagi antenna of your choosing, even designs that are not yet available!

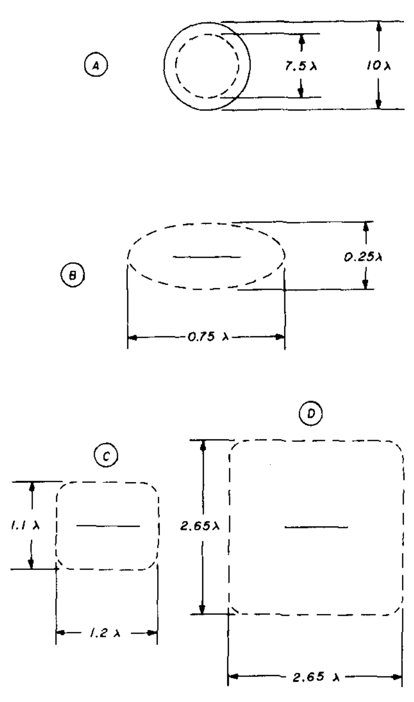


fig. 1. Typical effective apertures (shown within dotted lines) of some common antennas. *(A)* is a 10 wavelength diameter parabolic dish; *(B)* is a half wavelength dipole; *(C)* is a 1 wavelength Yagi; and *(D)* is a 10 wavelength Yagi.

table 1. Data on typical popular 2-meter through 23-cm Yagi designs including gain, boomlength, E and H beamwidths, E and H side lobe levels and recommended stacking distances. Data is believed to be accurate and has been gleaned from tests, data sheets, etc.

Yagi description	gain (dBi)	boom length (λ)	E B.W. degrees	H B.W. degrees	E S.L. (dB)	H S.L. (dB)	recommended stacking distance in E & H planes (λ)
general designs:							
NBS 3 element	9.25	0.4	57	72	23	12	1.00 × 0.60
NBS 5 element	11.35	0.8	48	56	22	13	1.20 × 0.90
NBS 6 element	12.35	1.2	40	42	19	12	1.40 × 1.10
NBS 12 element	14.40	2.2	34	36	17	13	1.55 × 1.40
NBS 17 element	15.55	3.2	28	33	16	12	1.80 × 1.35
NBS 15 element	16.35	4.2	26	29	17	13	1.95 × 1.75
2-meter designs:							
Cushcraft Jr. Boomer	14.40	2.2	34	36	17	13	1.55 × 1.40 (note 1)
Lunar 11 element	14.50*	2.6	31	34	17*	13*	1.65 × 1.50
F9FT 16 element	14.80	3.0	32	34	22	18*	1.80 × 1.50 (note 1)
KLM 2M-13LBA	15.00	3.1	28	33	18	15*	1.80 × 1.60 (note 1)
Cue Dee 15 element	15.15	3.1	30	32	16*	12*	1.70 × 1.40
Cushcraft Boomer	15.55	3.2	28	33	16	13	1.80 × 1.55
KLM 2M-16LBX	16.50	4.1	26	29	18*	15*	2.00 × 1.75 (note 1)
135 cm designs:							
Lunar 11 element	14.50	2.60	31	34	17*	13*	1.65 × 1.50
Cushcraft Boomer	16.35	4.2	26	29	17	13	1.95 × 1.75 (note 1)
KLM 220-22LBX	17.75	6.65	22	25	17	14	2.45 × 2.00 (note 1)
70-cm designs:							
KLM 432-16LB	15.20	5.3	30.0	33.0	17*	14*	1.70 × 1.55
K2RIW 13 element	15.40	5.3	29.5	29.5	10	7	1.50 × 1.50
K2RIW 19 element	17.35	5.6	24.0	26.0	18	15	2.10 × 1.80 (note 1)
F9FT 21 element	17.40	6.6	24.0	26.0	13	10*	2.10 × 1.75 (note 1)
FLEXA-YAGI 23 el.	17.95	7.2	24.0	25.0	17	15*	2.40 × 2.00
Cushcraft 424B	18.00	7.6	19.0	22.0	14	12	2.20 × 1.80 (note 1)
KLM 432-30LBX	19.40	9.6	19.0	20.0	17	14	2.70 × 2.40 (note 1)
W1JR 31 element	19.60	10.5	18.0	20.0	17	14*	2.70 × 2.40 (note 1)
23-cm designs:							
Tonna 23 element	17.00	7.5	19.0	19.0	12	10	2.40 × 2.40 (note 1)
W1JR 45 EL LPY	20.7	15.7	18.0	20.0	15	13*	2.85 × 2.65 (note 1)

*Estimated

Note 1. In this case actual tests have shown that a specific optimum is preferred (see text).

fundamental aperture concepts

Let us first examine some different antennas, each with its specific "effective aperture" or capture area. Some typical examples are shown in **fig. 1.** Other examples are contained in reference 2.

A 10 wavelength diameter parabolic antenna is shown in **fig. 1A.** It is easy to see how it has collection properties similar to that of the human ear. Note, however, that a dish antenna is not 100 percent efficient because it does not collect signals very well near its edge. Using simple geometry, the physical aperture of a 10 wavelength diameter dish is 78.5 square wavelengths, but its effective aperture is only approximately 44 square wavelengths.

A half-wave dipole antenna is shown in **fig. 1B.** Its aperture is more difficult to visualize. Note that its aperture extends out horizontally about 0.75 wavelength in the E plane and vertically about 0.25 wavelength in the H plane, bulging near the center and forming an ellipse. It has an effective aperture of approximately 0.13 square wavelength.

A Yagi has a slightly differently shaped aperture. A short (1 wavelength) conventional Yagi (one whose elements all lie completely in the same plane) is shown in **fig. 1C** and has a somewhat rectangular aperture, being wider in the E (horizontal) plane than in the H (vertical) plane. Its aperture is approximately 1.3 square wavelengths . A properly designed conventional 10 wavelength long Yagi as seen in **fig. 1D** has an almost square aperture of approximately 7 square wavelengths.

*Additional recommended reading: *Significant Phased Array Papers*, edited by R.C. Hansen, PN 0-89006-019-3, Artech House, Inc., 610 Washington Street, Dedham, Massachusetts 02026 ($13.00 plus $2.50 postage and handling).

Some of you may want to research this subject further to see how I determined the apertures. If you know the individual antenna directivity gain[3] you can calculate the effective aperture using the following equation:

$$effective\ aperture = \frac{gain}{12.5} \quad (1)$$

where effective aperture is in square wavelengths and gain is over isotropic as a numeric. For example, a ½ wavelength dipole has a gain of 1.64 (2.15 dBi). Therefore it has an effective aperture of 0.13 square wavelengths. A 1 wavelength Yagi has a directivity gain of approximately 16.5 (12.2 dBi) and an aperture of approximately 1.32 square wavelengths. Likewise, if we have a 10 wavelength Yagi with a gain of 87.5 (19.4 dBi), the aperture will be approximately 7 square wavelengths, quite an aperture increase over the 1 wavelength Yagi.

The above equation does not reveal the width or height of the aperture. In the case of the 10 wavelength Yagi the aperture is approximately square (**fig. 1D**). Therefore the horizontal and vertical dimensions are approximately the square root of the aperture ($\sqrt{7}$) or 2.65 wavelengths. Since the 1 wavelength Yagi is slightly rectangular (per **fig. 1C**), the aperture will be slightly wider than it is high or approximately 1.2 by 1.1 wavelengths, respectively.

Once the concept of aperture is understood, it is easy to see what happens when we try to stack two identical antennas. When they are in close proximity, their apertures overlap, as shown in **fig. 2A**. Hence the capture area will not be doubled. Also the gain will not be double (3 dB increase) that of the single antenna. Furthermore, when identical antennas are too closely spaced, they introduce mutual impedance effects that can play strange games with the pattern, power distribution, and VSWR.

If we move the antennas apart until their apertures just touch, as shown in **fig. 2B**, we should produce almost twice the capture area (more on this later). Separating the two antennas further as shown in **fig. 2C**, will definitely double the capture area but is not desirable for reasons to be discussed shortly. Similarly, stacking in the vertical plane is also possible and yields a similar increase in gain.

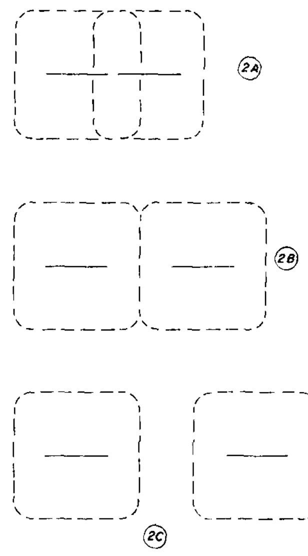


fig. 2. Effective apertures when stacking two square aperture antennas. *(A)* is too closely spaced; *(B)* is optimum spacing; and *(C)* is too loosely spaced.

stacking patterns

"So," you ask, "how does this relate to my Yagi?" First let us see what happens by looking at some typical antenna patterns. **Figure 3A** illustrates a typical antenna pattern for a 3-element Yagi.[4] Note that the half-power beamwidth is approximately 80 degrees and that this antenna pattern has very low side lobes.

We will stack two identical 3-element Yagis close together as shown in **fig. 2A**. The resultant antenna pattern is shown in **fig. 3B**. Note that the main beam narrows to about 50 degrees and the pattern is still very clean.

Next, let's separate the antennas further apart as shown in **fig. 2B**. The resultant antenna pattern is shown in **fig. 3C**. Note that the pattern beamwidth has become even narrower, 40 degrees, half the beamwidth of the original antenna. Also note that significant new lobes appear. These are properly referred to as "grating lobes" to differentiate them from the original single antenna's side lobes. The grating lobes in **fig. 3C** are only about 13 dB below the main beam. This separation is considered the optimum stacking distance.[5]

In **fig. 2C**, the antennas are spaced much further apart. The resulting antenna pattern is shown in **fig. 3D**. Note that in this case the main beam is approximately 20 degrees wide, or 25 percent that of the original antenna being stacked. Also note that the number of the grating lobes has increased to four, with maximum amplitude only 2 dB below the main beam. This is a technique often used by radio astronomers in interferometry. It allows very narrow beamwidths for greater accuracy in determining the position of extra-terrestrial objects. However, it is not usually desirable for Amateurs!

The patterns just shown all came from a "clean" Yagi. *If you look closely you will see that all the grating lobes and nulls were formed from within the area of the original antenna pattern.* This point is stressed because most Yagi antennas, especially those that are 1 wavelength or longer, usually have many side lobes *before* being stacked. The more side lobes you start with, the greater the chances are that the resulting pattern will be much "dirtier" and more complex than desired. Suppression of grating lobes is a difficult, if not impossible, task. Therefore, the real limitations when stacking antennas are the beamwidth, the side lobes in the antenna to be stacked, and the allowable level of the grating lobes. Despite the appearance that grating lobes are "robbing power" from the main beam, in actuality they are not since each grating lobe is very narrow. However, grating lobes are sources of extraneous noise or extra signal pickup, a killer on EME, and when there is lots of QRM. Incidentally, if you have many strong grating lobes, it is easy to accidentally peak your antenna on one of them instead of on the main beam!

To review, the optimum stacking distance for two identical antennas oc-

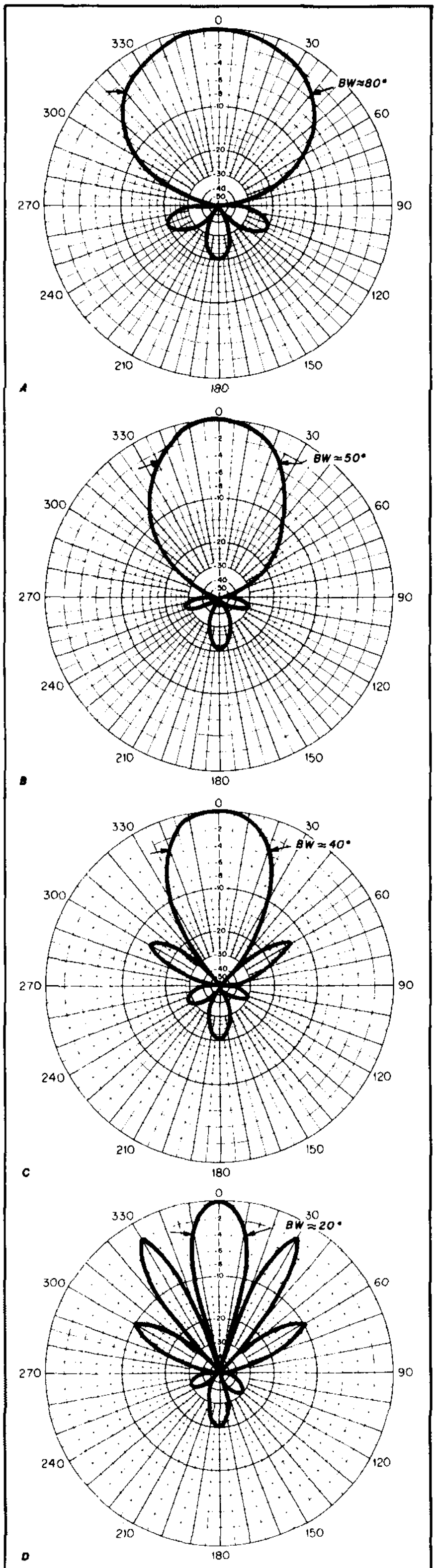
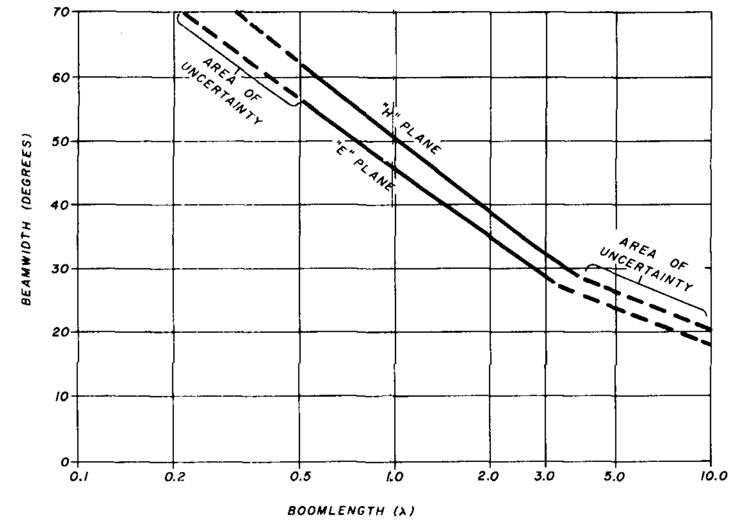


fig. 3. Typical antenna and stacking patterns. *(A)* is a typical 3-element Yagi; *(B)* two 3-element Yagis stacked 1/2 wavelength; *(C)* two 3-element Yagis stacked 3/4 wavelength; and *(D)* two 3-element Yagis stacked 2 wavelengths.



4. Typical beamwidths versus boomlength for well-designed Yagi antennas.

curs when the beamwidth of the array has been narrowed to about 50 percent and the grating lobes are approximately 13 dB below the main beam (more on this later). If four antennas are stacked in the same plane, optimum stacking would yield about 25 percent of the beamwidth of the original antenna while the grating lobes should still be 13 dB below the main beam.

actual stacking

There are two basic stacking methods: uniformly illuminated and shaped. Uniformly illuminated means that the antennas are all spaced the same distance apart in each plane and each is fed with the same amount of power. This method yields the maximum gain for its size and is the method most often used by Amateurs.

The shaped method is often used by professional antenna designers, especially in phased-array radars, where very low grating lobes are necessary. The individual antennas may be unequally spaced (sometimes one is completely left out!) and often are fed with different amounts of power. Since these techniques yield lower gain and are quite complex, they are usually not desired by Amateurs.

The subject that I have been discussing is called "pattern multiplication."[6] For more information on these techniques (pattern multiplication), see references 5 through 10.*

The most important parameters needed to determine optimum stacking distance are the beamwidths in the E and H plane of the antenna to be stacked. Also, the level of the first side lobe on the antennas to be stacked is important.

Most antenna manufacturers and antenna designers know the E plane beamwidth of their antenna very accurately since it is not difficult to measure. As discussed in reference 3, the beamwidths are often specified and accurate (in contrast to the gain claims). If the beamwidths are not known, they can be estimated from the "true" antenna gain. Several gain determining methods were described in detail in last May's column.[3] To save you even further time, I have prepared **table 1**, which lists many parameters of some of the most popular Yagi antennas.

The beamwidths of a Yagi antenna can be estimated if the boomlength is known. To assist you in this exercise, I have prepared a graph of E and H beamwidths versus boomlength for typical Yagi antennas (see **fig. 4**). All you need to know is the boomlength in wavelengths. For example, the E and H plane beamwidths of a typical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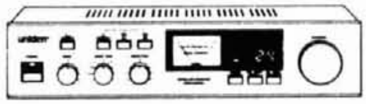

Little Wil by Wilson $ 899

Boman SR1500 System $1133

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M/A Com H1 System $2082

M/A Com T1 System $1982

Sigma Mark 2A Sys. $1133

Sigma Mark 3 BL Sys. $1362

Sigma Mark 5 BL Sys. $1433

STS MBS-SR System $1512

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Prodelin 10' Dish	$400
Ranger 11' Mesh Dish	$300
Magnum 12' Dish	$400

175

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MTI 2100	$225
MTI 4100	$345
Houston Tracker IV	$325
Houston Tracker IV +	$425

2 wavelength Yagi are approximately 35 and 39 degrees, respectively.

Also note in **fig. 4** that in a conventional Yagi the E plane is typically narrower than the H plane. While the E plane beamwidth is often available on data sheets, the H plane beamwidth is seldom shown (it is slightly more difficult to accurately measure). However, it can usually be "guesstimated" to be 10 percent greater than the E plane. For example, a Yagi with a 30-degree E plane beamwidth has a typical H plane beamwidth of 33 degrees.

Also, many Yagi antennas have side lobes that are so strong that they are equal to or greater than the idealized grating lobe desired after stacking! Therefore, it should be obvious that if the sidelobes on an antenna to be stacked are equal to or less than 13 dB below the main beam peak, they can't be optimally stacked. In this case the antennas must be placed closer together than optimum and consequently will yield lower stacking gain (more on this later)!

As a rule of thumb, the H plane side lobe level on a conventional Yagi is typically 3 dB stronger than the E plane lobe. Therefore, a Yagi with an 18 dB down sidelobe in the E plane probably has an H plane side lobe approximately 15 dB down from the main beam.

Now that we have determined the beamwidth and side lobe levels of our antenna, how do we determine the optimum stacking distance? For antennas with very low side lobes (at least 18 dB below the main beam):

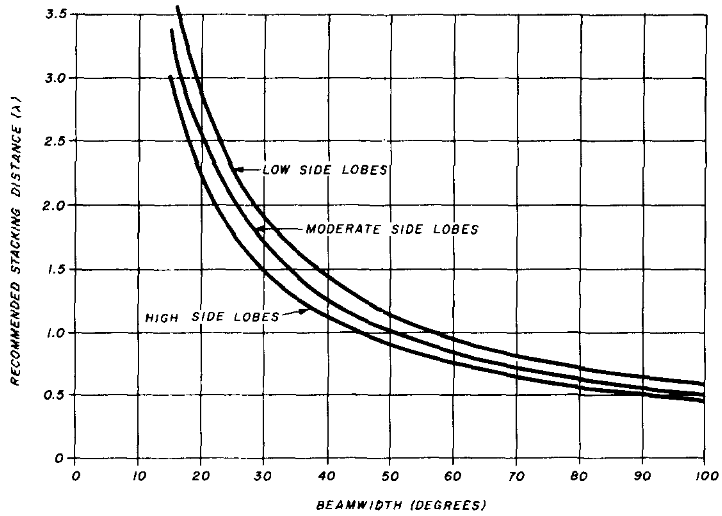
$$\textit{stacking distance} \approx \frac{57}{\textit{beamwidth}} \quad \textbf{(2)}$$

where stacking distance is in wavelength and beamwidth is in degrees.[10]

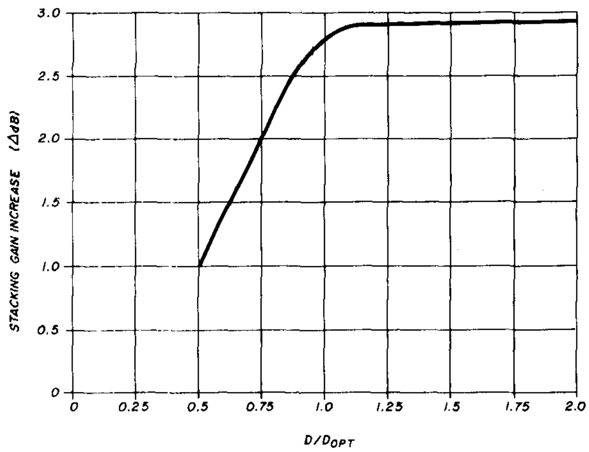
If the side lobes are typically 13-17 dB down, the usual situation, my tests have verified that:

$$\textit{stacking distance} \approx \frac{51}{\textit{beamwidth}} \quad \textbf{(3)}$$

If the side lobes are only 12 dB down (or less) (typical of the H plane of many Yagis):



5. Recommended stacking distance from *eqs. 2, 3,* and *4.* See text for proper usage and interpolation.



6. Typical gain increase due to stacking antennas. "D" is the stacking distance and "Dopt" is the recommended optimum from *eqs. 2, 3,* and *4.*

$$\textit{stacking distance} \approx \frac{45}{\textit{beamwidth}} \quad \textbf{(4)}$$

For simplicity, I have incorporated these formulas into a graph **fig. 5**), which is an updated version of the graph discussed earlier. It incorporates the recommended stacking for antennas with different levels of side lobes.

It can also be seen that there is room for interpolation using **eqs. 2, 3,** and **4** if desired.

For example, the NBS 17 element 3.2 wavelength Yagi has an E plane beamwidth of 28 degrees and a 16 dB down side lobe. Therefore use **eq. 3** (or the graph). Hence, the recommended E plane stacking distance is approximately 1.8 wavelength. The H plane beamwidth is 33 degrees but the side lobe is only 12 dB down. Using **eq. 4**, the recommended H plane stacking distance is approximately 1.35 wavelengths, quite a bit less than expected.

For those who do not want to make the required calculations, **table 1** also includes the recommended stacking distance for the antennas listed. In some cases, actual test measurements have been taken to determine the optimum spacing. Therefore, if the recommended stacking distance is different from that which you calculate (signified by note 1), it is the preferred value since actual tests have verified its validity.

stacking gain

So how much gain do you get if you use the recommended stacking distance? Günther Hoch, DL6WU, has carried out tests and presented some answers to this question.[10] I have incorporated his information in **fig. 6.** It can be seen that the gain approaches 3 dB, but only at very wide spacing as discussed earlier. A typical optimum value is about 2.8 dB. If you understack (i.e., position too closely) by about 25 percent (as illustrated in **fig. 2A**), the gain increase (from a single antenna to the array) will be reduced to about 2 dB and there will be almost no grating lobes! Obviously there is not a great degree of freedom when optimum gain and pattern are concerned.

Finally when an antenna has high side lobes, it must be stacked closer to control the grating lobes as shown by **eq. 3** or **4**. This represents a form of understacking and lower gain. When this situation, plus feedline losses and mutual coupling, are considered, *you are probably lucky to attain 2.5 dB even when the optimum stacking distance is used* (more on this subject next month).

After studying this subject and reference 3, it will become obvious that the level of the first side lobe is a very important antenna parameter. Unfortunately, most antenna designers rarely list this parameter, but instead often list the worthless front-to-side ratio! A change in the literature indicating the level of the first side lobe would be an improvement.

final evaluation

You should be able to test your pattern using the information just given, especially if your antennas are stacked in the horizontal plane. The test methods described in reference 3 should be sufficient. For those on EME, the sun can be used as a rough check. Always remember that if the antenna beamwidth is narrower than originally calculated and/or the grating *lobes are less than 13 dB down from* the main lobe, you have probably overstacked (i.e., positioned your antennas too far apart).

summary

Part 1 of this two-part series has been written to give you a feel for what happens when two or more antennas are stacked. Typical examples have been provided along with the equations and graphs necessary for determining optimum stacking distance. New or improved Yagi antennas or those I may have failed to mention can be quickly evaluated using the information provided in this article. Exact stacking distance is not critical since there are many compromises, as discussed.

Part 2 of this article will delve a little deeper into the subject, emphasizing the practical aspects of the subject with recommendations on how to obtain optimal performance in your particular situation.

acknowledgements

I would particularly like to thank Günther Hoch, DL6WU, Dave Olean, K1WHS, and Steve Powlishen, K1FO, for the test data they shared with me while I was preparing this article.

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upcoming VHF/UHF events

April 5:	*EME Perigee*
April 15:	*ARRL 2-meter Sprint Contest*
April 21:	*1400 UTC: predicted peak of Lyrids meteor shower*
April 21:	*ARRL 135-cm Sprint Contest*
April 26-28:	*Dayton Hamvention VHF/UHF Program (contact WA8ONQ)*
May 1:	*ARRL 70-cm Sprint Contest*
May 4:	*EME Perigee*
May 4:	*1300 UTC: predicted peak of Eta Aquarids meteor shower*
May 3-5:	*West Coast VHF Conference (contact W6RXQ)*
May 9:	*ARRL 23-cm Sprint Contest*
May 18,19:	*Eastern VHF/UHF Conference (contact W1EJ)*
May 19:	*ARRL 6-meter Sprint Contest (tentative)*

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5.595-.500/4, 5.595-.500/4/CW 4 pole 500 cycles wide CW. Impedance 800ohms 15pf In/800ohms 0pf out.	19.99
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MOTOROLA	4884863B01	11.7MHz 2pole 15KHz bandwidth	5.00
PTI	5350C	12MHz 2pole 15KHz bandwidth	5.00
PTI	5426C	21.4MHz 2pole 15KHz bandwidth	5.00
PTI	1479	10.7MHz 8pole bandwidth 7.5KHz at 3dB, 5KHz at 6dB	20.00
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FRC	ERXF-15700	20.6MHz 36KHz wide	10.00
FILTECH	2131	CF 7.825MHz	10.00

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	SFE10.7MS	10,7MHz 230KHz +-50KHz at 3dB , 570KHz at 20dB	2.50
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SD1212-8	4.95	SD1305	3.00	SD1452	20.00	SRF1431	40.00	SD1262	15.00	SD1413-1	18.00	SD1539H	100.00	SRF2918	15.00
SD1212-11	4.95	SD1307	3.00	SD1452-4	24.00	SRF1834	40.00	SD1263	15.00	SD1416	28.00	SD1542H1	170.00	SRF2919	15.00
SD1212-16	4.95	SD1308	3.00	SD1453H1	20.00	SRF2053-3	60.00	SD1263-1	15.00	SD1422-2	24.00	SD1544	26.00	SRF3071FF	50.00
SD1214-7	5.00	SD1311	1.00	SD1454-1	48.00	SRF2092	50.00	SD1272	10.95	SD1428	24.00	SD1545	33.00	SS4006	25.00
SD1214-11	5.00	SD1317	8.00	SD1477	35.00	SRF2147	22.00	SD1272-1	10.95	SD1428- 6084	12.00	SD1546H1	55.00	SS4132	15.00
SD1216	12.00	SD1319	2.50	SD1478	21.00	SRF2225	15.00	SD1272-2	10.95	SD1429-2	15.00	SD1561	79.00	TA7686	15.00
SD1219-4	15.00	SD1345-6	5.00	SD1480	53.00	SRF2264	25.00	SD1272-4	10.95	SD1429-3	14.90	SD1574-1	6.95	TA8559	15.00
SD1219-5	15.00	SD1347-1	1.00	SD1484	1.50	SRF2265	100.00	SD1278	13.75	SD1429-5	15.00	SD1575	6.95	TA8561	15.00
SD1219-8	15.00	SD1365-1	2.50	SD1484-5	1.50	SRF2281	5.00	SD1278-1	13.75	SD1430	12.00	SF4557	25.00	TA8562	15.00
SD1220	8.00	SD1365-5	2.50	SD1484-6	1.50	SRF2371	15.00	SD1278-5	13.75	SD1430-2	18.00	SK3048	5.00	TA8563	15.00
SD1220-1	9.50	SD1375	7.50	SD1484-7	1.50	SRF2347	50.00	SD1279-1	18.00	SD1434	28.00	SL501-59	15.00	TA8564	15.00
SD1220-9	8.00	SD1375-6	7.50	SD1488	22.85	SRF2356	38.00	SD1279-3	18.00	SD1434-5	28.00	SL501-173	15.00	TA8894	15.00
SD1222-8	16.00	SD1379	15.00	SD1488-1	28.00	SRF2378	16.00	SD1281-2	8.00	SD1434-9	28.00	SM7714	5.00	TIS189	3.55
SD1222-11	7.50	SD1380-1	1.00	SD1488-7	27.00	SRF2572	25.00	SD1283	10.00	SD1438	26.00	SRF112	15.00	TP312	2.50
SD1224-10	18.00	SD1380-3	1.00	SD1488-8	28.00	SRF2584	40.00	SD1283-2	10.60	SD1441	56.00	SRF395	50.00	TP1014	5.00
SD1225	18.00	SD1380-7	1.00	SD1499-1	36.00	SRF2597	25.00	SD1283-3	10.00	SD1442	15.00	SRF750	36.00	TP1028	15.00
SD1225-1	15.00	SD1405	21.00	SD1511H3	75.00	SRF2741	40.00	SD1283-4	10.00	SD1444	3.25	SRF769H	20.00	TRW3	5.00
SD1229-7	10.95	SD1408	25.00	SD1520-2	18.00	SRF2747	40.00	SD1289-1	15.00	SD1444-8	3.25	SRF887K3	2.50	TXVF2201/HF	450.00
SD1229-16	10.95	SD1409	18.00	SD1522-4	33.00	SRF2767H	40.00	SD1290-4	15.00	SD1444-9	3.25	SRF989K	15.00	V222-2	25.00
SD1232	4.00	SD1410	18.00	SD1528-1	24.00	SRF2821	25.00	SD1290-7	15.00	SD1446	4.03	SRF1005	50.00	V4101E	20.00
SD1240-8	15.00	SD1410-3	21.00	SD1528-3	34.00	SRF2822/2N6603	13.50	SD1300	1.25	SD1450-1	28.00	SRF1018	5.00	V415	5.00
SD1244-1	14.00	SD1410-6	21.00	SD1530-2	38.00	SRF2857	20.00	SD1301-7	3.00	SD1451	15.00	SRF1074	50.00		

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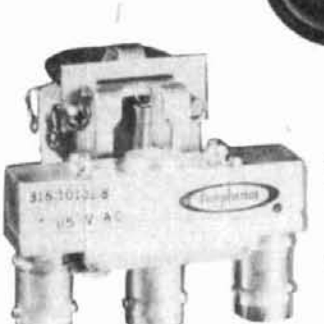

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CONFIRMING ORDERS: We would prefer that confirming orders not be sent after a telephone order has been placed. If company policy necessitates a confirming order, please mark "CONFIRMING" boldly on the order. If problems or duplicate shipments occur due to an order which is not properly marked, the customer will be held responsible for any charges incurred, plus a 15% restock charge on the returned parts.

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

FOREIGN ORDERS: All foreign orders must be prepaid with a Cashier's Check, or Money Order made out in U.S. FUNDS ONLY. We are sorry but C.O.D. is not available to foreign countries and letters of credit are unacceptable as a form of payment. Further information is available on request.

HOURS: Monday thru Friday 8:30 a.m. to 5:00 p.m. Saturdays 8:30 a.m. to 4:00 p.m.

INSURANCE: Please include 25¢ for each additional $100.00 over $100.00, UPS ONLY. All insured packages are shipped thru UPS only. If you wish to have it shipped through the post office there is a $5.00 fee which is additional to the shipping, handling and insurance.

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PARTS: We reserve the right to substitute or replace any item with a part of equal or comparable specification.

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PRICES: Prices are subject to change without notice.

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RESTOCK CHARGES: If parts are returned to MHZ ELECTRONICS, INC. due to customer error, the customer will be held responsible for all fees incurred and will be charged a 15% RESTOCK CHARGE with the remainder in CREDIT ONLY. The following must accompany any return; A copy of our invoice, return authorization number which must be obtained prior to shipping the merchandise back. Returns must be done within 10 DAYS of receipt of parcel. Return authorization numbers can be obtained by calling (602) 242-8916 or notifying us by post card. Return authorizations will not be given out on our 800 number.

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

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

OUR 800 NUMBER IS STRICTLY FOR ORDERS ONLY (800) 528-0180. INFORMATION CALLS ARE TAKEN ON (602) 242-8916 or (602) 242-3037.

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RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

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COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

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

WANTED: Circuit diagram/Gladding Islander — Pearce Simpson. Harvey Gilbert, VP9AN, Box MA124, Somerset, Bermuda. Call collect (809) 29-40112 weekends.

WANTED: SCR 274N/BC — ARC5/R-27 RCVRS only. Complete or parts. Send description, cost. Will pay shipping. W8KSW, Walt Evanyk, 3200 Sherrye Drive, Plano, Texas 75074.

LETTERING and text for QSL pasteup and label usage. Stamp brings circular. Labelcraft, PO Box 412, W. Sand Lake, NY 12196.

QSL Announcement: I regret to inform you I have sold the SAMCO Travel-Pak QSL Kit Co. Thank you for your QSL patronage. 73, "SAM" A. Moles, PO Box 412, W. Sand Lake, NY 12196.

ELECTRONIC CMOS keyer kit pcb & parts only $9.95 plus $1.50 shipping. WI res. add 5% tax. Send for free information. Bel-Tek, PO Box 125H, Beloit, WI 53511.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

WANTED: Old Crosley Radio Model 50, 51, 52, & "Pup". K4NBN "No Bad News".

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SSB Crystal Filter, 9.0 MHz, 2.4 kHz bandwidth, 6 pole, $17.50; two for $30.00. 4CX250B chimneys, $5.00, three for $12.00. ICL7117 LED DVM chips, $8.00; three for $20.00. Dentron C.A.P. transceiver, 100 watts, solid state, $275.00; ACPS, $85.00. GLA1000-C linear, 160-15-10, $285.00. All items new, postpaid, data. National receiver collection, SASE list. Wanted, automobile radio vibrators, new. W.E. Delage, PO Box 493, Northfield, Ohio 44067. (216) 274-2012.

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SPECTACULAR Savings. ICOM, Kenwood, Yaesu, Panasonic, Regency, Bearcat, MFJ, antennas, coax, CW/RTTY, much more! Free UPS shipping and insurance to 48 states. 25 page picture catalog $1.00 (refundable). Galaxy Electronics, Box 1202HR, 67 Eber Ave., Akron, Ohio 44309. (216) 376-2402 9-5 PM EST.

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

NEW President Grant AM-SSB CB. With power mic $125.00 plus $5.00 UPS. MO only. R.D. Carter, Box 418, Vass, NC 28394.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

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

CHASSIS and cabinet kits. SASE K3IWK.

HELP. Need schematic Atronics KB105MP CW keyboard. Company disappeared. Cavett, 8570 Herbert, Pennsauken, NJ 08109.

U.S. and FOREIGN Callbook, Radio Handbook and other ARRL books, 1950 to 1975 edition. Ham Radio, QST, 73, CQ magazines, second-hand for my private collection. Send price list to: Vittorio Bruni, Poste Succ. 8 — 05100 Terni Succ. 8, Italy.

IMRA, International Mission Radio Association, helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, NY 10538.

ANTIQUE RADIO Classified — The National Publication for Buyers and Sellers of old radios — published monthly. Sample copy $1.50. A.R.C., 9951-9 Sunrise, Cleveland, Ohio 44133.

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

AMATEUR RADIO CLASSES: Sponsored by the Hampden County Radio Association starting February 26, Agawam High School, 7 PM. Novice, Tech, General, Advanced or Extra. There is no charge for the classes. Text books must be purchased but are less than $10. Classes meet weekly for 10 weeks. For more information or to sign up contact Art Zavarella (413) 786-9115.

VE TEST SESSION: Sponsored by the Hampden County Radio Association, Saturday, May 18, 9 AM, Hampden-Wilbraham Regional High School,621 Main Street, Wilbraham, Mass. Exams for all license grades will be offered. Send completed FCC Form 610, a copy of current license and check for $4.00 payable to ARRL-VEC to: Yorke Phillips, K1BXE, 235 Ames Road, Hampden, MA 01036 prior to April 18, 1985.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

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

AMATEUR RADIO CLASSES: Sponsored by the 19-79 Amateur Radio Association. Evening classes start March 19 at the Chelsea (Mass) High School. Open to anyone interested in becoming a ham. Classes are free with a small cost for materials only. For information: 19-79 Amateur Radio Association, PO Box 171, Chelsea, MA 02150 include phone number.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49082.

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Coming Events

ACTIVITIES

"Places to go..."

OHIO: Dayton Hamvention, April 26, 27, 28, Hara Arena and Exhibition Center, Dayton. Admission $8 advance, $10 at door. Good for all three days. Banquet $14 advance, $16 at door. Flea market space $17 in advance for all three days. Technical, ARRL and FCC forums. New products and exhibits. Special group meetings. YL forum. International VHF/UHF conference. Amateur of the Year Award. Special achievement awards. Pre-registration starts January 1, 1985. For further information; Dayton Amateur Radio Association, Box 44, Dayton, OH 45401 or phone (513) 433-7720.

OHIO: The 16th annual B*A*S*H will be held on the Friday night of the Dayton Hamvention, April 26 at the Convention Center, Main and Fifth Streets. Parking in adjacent city garage. Admission free to all. Sandwiches, snacks and COD bar available. Live entertainment. Two exciting awards and many others. For further information contact the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401.

ARIZONA: The Cochise Amateur Radio Association (CARA) invites you to the inauguration of the CARA Training Facility and Range, a 40 acre complex in Cochise County, 5 miles east of Sierra Vista on Mason Road. The dedication will be our annual Hamfest, May 4 and 5. A flea market is planned and all tailgaters are welcome. For information: The Cochise ARA, PO Box 1855, Sierra Vista, AZ 85636. Att: KB7HB.

WISCONSIN: The Madison Area Repeater Association (MARA) announces its 13th annual Madison Swapfest, Sunday, April 21, Dane County Exposition Center Forum Building in Madison. Doors open for commercial exhibitors and flea market sellers at 8 AM. General admission 9 AM. Everything for hams, computer hobbyists and experimenters. All-you-can-eat pancake breakfast and Bar-B-Q lunch will be available. Admission $2.50/advance; $3.00/door. Children 12 and under admitted free. Flea market tables $4.00 each/advance; $5.00/door. Reserve early. Talk in on WB9AER/R, 146.16/.76. For reservations or information: M.A.R.A., PO Box 3403, Madison WI 53704.

CALIFORNIA: The 43rd annual Fresno Hamfest, May 3, 4 and 5, Tropicana Lodge of Fresno, 1406 N. Blackstone. Program includes tech talks, swap tables and flea market, transmitter hunts, CW contest, ARRL forum, commercial exhibits, eyeball QSO's, buffet dinner and more. Registration $24.00 before April 19, 1985. $26.00 after that date. For information: Jane Price, WA6HSW, 2353 W. Simpson, Fresno, CA 93705.

ILLINOIS: The Kishwaukee Amateur Radio Club's annual Hamfest, May 5, DeKalb County Fairgrounds, Suydam Road, Sandwich. Donation $2.00 advance; $3.00 gate. Inside tables $5.00 each. Free parking. Outside areas for tailgaters. Overnight camping, no hookups. Coffee and donuts available for early birds. Food wagon thereafter. Talk in on 94, 13-73. For tickets write: K.A.R.C., Box 334, Sycamore, IL 60178. Our 30th Year!

COLORADO: The Grand Mesa Repeater Society's 6th annual Western Slope Amateur Radio and Computer Swapfest, Saturday, April 20, 10 AM to 4 PM. Location TBA in Grand Junction. Free admission. Swap tables $5.00 each. Indoor Swapfest, Amateur Radio exams, auction and refreshments. Talk in on 146.82 and 449.200. For tables or information: SASE to Larry Brooks, WB0ECV, 3185 Bunting Avenue, Grand Junction, CO 81504 or call (303) 434-5603.

SOUTH CAROLINA: The Blue Ridge Amateur Radio Society is sponsoring the 46th annual Hamfest and Electronic Flea Market, American Legion Fairgrounds in Greenville. Saturday, May 4 8 AM to 5 PM. Sunday, May 5 8 AM to 3 PM. Admission $3.00 advance and $4.00 at gate. VEC walk in exams, Wouff Hong ceremony, ARRL State Convention, Saturday night banquet, ARES, QCWA, indoor dealer displays, indoor/outdoor flea market, food, beverages, snacks and camping. Early dealer/flea market setups with advance registration. For advance tickets and VEC exam info; Sue Chism, N4ENX, PO

Box 6751, Greenville, SC 29606. For additional information: Rancy Rice, WD4ADK, 1401 W. Parker Rd., Greenville, SC 29611.

OKLAHOMA: The Great Plains ARC will sponsor its 4th annual N.W. Oklahoma Eyeball & Swapmeet, Sunday, April 14, starting 9 AM in Mooreland. Admission $2.00. Dealer and swap tables available at no charge. VE tests given. Campsites available. Local airport. Covered dish dinner at noon. For further information: Gordon Richmond, NR5L, Rt. 1, Box 12, Mooreland, OK 73852 or Gerald Bowman, Box 356, Mooreland, OK 73852 or call (405) 994-5394. (405) 994-5453.

NORTH CAROLINA: The Raleigh Amateur Radio Society's 13th annual Hamfest, Sunday, April 14. 8 AM to 4 PM. NEW LOCATION Jim Graham Building, NC State Fairgrounds, Hillsborough Street, Raleigh. Pre-registration $3.50. $5.00 at the door. One flea market space, table, two chairs $5.00 (ours only please). Vendor setup Saturday 4-10 PM, Sunday 6-8 AM. Hamfest social Saturday night. Special interest meetings. Amateur FCC exams. Send Form 610 with copy of current license and check or MO for $4.00 to: W.C.A.R.S./V.E.C., Mr. John Johnson, WM4P, 2118 Lyndhurst Drive, Raleigh, NC 27610. CW and homebrew contests. Talk in on W4DW (146.04/146.64) and K4ITL (146.28/146.88). For further information: RARS Hamfest, PO Box 17124, Raleigh, NC 27619.

MICHIGAN: 1985 Blossomland Blast, Sunday, October 6, 1985. Write "Blast", PO Box 175, St. Joseph, MI 49085.

INDIANA: The Putnam County Amateur Radio Club's third Auction and Flea Market, April 6, Putnam County Fairgrounds, north of Greencastle on US 231. Doors open for setup at 0600. Flea Market 0800. Flea market tables $2.00 each. Admission $3.00. Children under 12 free. Auction starts 1300. Food and beverages available. Commercial exhibitors welcome. For information SASE to John S. Underwood, K9IIB, RFD 1, Box 10, Fillmore, IN 46128 or call (317) 246-6335.

NEW YORK: Indoor/outdoor Flea Market sponsored by the Suffolk County Radio Club, Sunday, May 5, 8 AM to 3 PM, Republic Lodge No. 1987, 585 Broadhollow Road (Rt. 110), Melville. Refreshments available. Free parking. Admission $2.00. (Spouse and kids free). Indoor tables $7.00, outdoor space $5.00, includes one admission. Talk in on 144.61/145.21 and 146.52. For information: Richard Tygar, AC2P. (516) 643-5956 evenings.

MISSOURI: The PHD Amateur Radio Association is sponsoring the State ARRL Convention, Saturday and Sunday, April 13 and 14, Trade Mart Building II, Kansas City Downtown Airport. Doors open 9:30-5:30 both days. Commercial setput 7-9 PM Friday, 7-9 AM Saturday. Saturday evening banquet. Special guest: Dale Clift, WA3NLO, ARRL General Manager. Forums include ARRL, computer, FCC, VE, DX, PR, CW contest, homebrew contest and much more. Exams by PHDVEC Friday 5 to 7 PM, Saturday and Sunday, 8 AM. Send application with $1.00 and SASE to PHDVEC, PO Box 11, Liberty, MO 64068-0011 by April 8, 1985. Registration $4.00 for both days. Banquet $10.50. Swap tables $10.00 for both days, includes one registration per table. Free parking. RV's welcome but no hookups. Talk in on 146.34/94. Send registrations to: PHD ARA, PO Box 11, Liberty, MO 64068-0011. Phone (816) 781-7313 or 452-9321.

ARKANSAS: The Northwest Arkansas ARC will hold its 5th annual Hamfest/Swapfest, Saturday, May 4, Rogers Youth Center, 315 West Olive, Rogers. 8 AM to 4 PM. Setup 7 AM. Commercial/flea market tables $2.00. Admission free. Walk in FCC exams given 10 AM and 1:30 PM. Talk in on 16/76 and 52 simplex. For information SASE to: Ray Watson, N5HAP, 714 Maple Drive, Springdale, AR 72765 or Dave Perry, KE5QZ, 3201 N. 13th, Rogers, AR 72756.

CALIFORNIA: The 13th annual Sacramento Valley Amateur Radio Hamswap, Sunday, May 5, Placer County Fairgrounds, Roseville. 9 AM to 3 PM. Talk in on 145.190 and 224.78, K6IS repeaters. Free parking. For advance tables, tickets, information: Carl Schultz, KA6KWB, 2942 Gwendolyn Way, Rancho Cordova, CA 95670. (916) 336-9111.

MASSACHUSETTS: The Hampden County Radio Association Flea Market, Sunday, May 5, rain or shine, West Springfield Lodge of Elks, Morgan Road, West Springfield. 9 AM to 3 PM. Admission $1.00 per person. Tables $3.00 each. Dealers $3.00 per vehicle display. Food/refreshments available. Talk in on 147.105 up 600. For information: Paul Kress, WA1ZKT (413) 568-8291 or Steve Nelson, WA1EYF (413) 596-8216.

NEVADA: The Ziegfeld Showroom of the MGM Grand Hotel will host the first 1985 earth station industry Banquet, held in conjunction with the SPACE/STTI Las Vegas show, Monday evening, April 1 at 6 PM. Festivities include dinner, special guest speakers and presentations. All topped off with Jubilee, the MGM's musical extravaganza. Tickets $50.00 per person. Call SPACE (703) 549-6900. Credit cards accepted or mail check to 709 Pendleton Street, Alexandria, VA 22314 before March 15. Banquet tickets are not refundable.

OHIO: The Medina Two Meter Group is sponsoring a Hamfest, May 12, Medina County Community Center Building, Lafayette Rd., State Rt. 42 SW. 8 AM to 2 PM. Setup 7 AM. Refreshments and free parking. Tickets $3.00 advance, $3.50 at door. Tables $6.00. Flea market space $2.00. Talk in on 147.63/03, K8TV/R. For table reservations and tickets: PO Box 452, Medina, Ohio 44258 or (216) 725-5021.

LOUISIANA: BRARC Hamfest, May 11 and 12, Baton Rouge. Free admission. VE exams Saturday and Sunday, 30 day advance registration only. Send SASE, 610 and check for $4.00 to ARRL/VEC, George Perry, W5LVX, 17424 Lady Constance, Greenwell Springs, LA 70739. For further information SASE to Rick Pourciau, N5HHF, 879 Castle Kirk, Baton Rouge, LA 70808.

GEORGIA: The Athens Amateur Radio Club (formerly the N.E. Georgia ARC) will sponsor a Hamfest, April 21, 8:30 AM to 3:30 PM, Athens Vocational-Technical School, Highway 29, Athens. Registration is free. Talk in on 147.285. For information: Norman Archibald, KB4IIA, PO Box 225, Athens, GA 30603.

NEW MEXICO: The UNM and Westside ARC's are co-sponsoring a tailgate swapfest, April 20, 10 AM to 2 PM MST, UNM North Campus parking lot, Tucker Avenue and University Blvd., Albuquerque. There is no charge but bring your own tables. Talk in on 147.75/147.15 and 449.3/444.3 repeaters. For information SASE to Robert A. Scupp, WB5YYX, 648 Marquis Drive NE, Albuquerque, NM 87123. (505) 296-6546.

CALIFORNIA: Flea Market and FCC examinations. April 13, May 11, June 8, July 13, August 10 and September 14. Novice thru Extra exams given. Information call (408) 255-9000. Foothill College, Los Altos, CA. 73 Gordon, W6NLG VEC.

ILLINOIS: The Moultrie Amateur Radio Klub (MARK) Hamfest, Sunday, May 5, 8 AM to 3 PM, Moultrie County 4-H Center Fairgrounds, Cadwell Rd., 5 miles east of Sullivan. Heated indoor/large covered outdoor Flea Market. No charge to vendors. Vendors setup Saturday. No overnight hookups. Talk in on 655/055 and 52. For information: MARK, PO Box 79, Sullivan, IL 61951 or call Vernon Jack, K9SWY (217) 728-7596.

NEW HAMPSHIRE: Springfest '85, the 5th annual Flea Market/Hamfest, sponsored by the Great Bay Radio Association. Saturday, April 20, 9 AM to 3 PM, Somersworth Armory, Blackwater Road, Somersworth. Admission $1.00. Tables $8.00 includes one admission. Free parking. Food and refreshments available. Talk in on 146.40/147.00. For information/table reservations: Great Bay Radio Association, PO Box 911, Dover, NH 03820.

ILLINOIS: The Centralia Wireless Association's annual Hamfest, Sunday, May, Kaskaskia College Gymnasium, 3 miles NW of Centralia. Doors open 7 AM to setups. No charge for flea market and exhibit space. Some tables available. Admission to Hamfest is free. Food and refreshments available. Exams for all license classes except Novice will be given at 9 AM. Send completed Form 610, copy of current license and check for $4.00 payable to ARRL/VEC to Lou Hodges, W9IL, Route 1, Box 62A, Centralia, IL 62801 by April 5, 1985. For further information: David Conder, KA9QPC (618) 532-2772 or Lou Hodges, W9IL (618) 533-4724.

MASSACHUSETTS: The Framingham Amateur Radio Association's annual Spring Flea Market, Sunday, April 14, Framingham Civic League Bldg., 214 Concord St. (Rt. 126) downtown Framingham. Doors open 10 AM. Sellers setup begins 8:30. Admission $2.00. Tables $10.00 includes one free admission. Pre-registration required. Bargains galore! Contact Jon Weiner, K1VVC, 52 Overlook Drive, Framingham, MA 01701. (617) 877-7166.

MASSACHUSETTS: The Montachusett Amateur Radio Association's Flea Market, Saturday, April 27, Knights of Columbus Hall, Electric Avenue, Fitchburg. Doors open 9 AM to 3 PM. Dealer setup 8 AM. Admission $1.00. Tables $8.00 each. Refreshments available. Free parking. Talk in on 144.85/145.45 and 146.52 simplex. For tables send check payable to M.A.R.A., Jim Beauregard, KB1AY, 7 Mountain Avenue, Fitchburg, MA 01424.

MINNESOTA: The Arrowhead Radio Amateur Club announces "Swapfest '85", Saturday, May 11, Holiday Inn, 207 West Superior Street,, downtown Duluth. Doors open 8 AM for vendors. General admission 10 AM. Admission $4.00. 4' tables $5.00. Plenty of food. Free parking. Talk in on 146.34/94 repeater. For information: Bill Cossette, NØBKL, 15 Manitou Street, Duluth, MN 55808.

MINNESOTA: The Rochester Amateur Radio Club's 8th annual Hamfest, Saturday, April 20, John Adams Junior High School, 1525 NW 31st Street, Rochester. Doors open 8:30 AM. Large indoor flea market for radio and electronics items, refreshments and plenty of free parking. Talk in on 146.22/82. For further information: RARC, c/o WBØYEE, 2253 Nordic Ct., NW, Rochester, MN 55901.

OPERATING EVENTS

"Things to do..."

APRIL 13: Connecticut QSO Party sponsored by the Candlewood ARA, 1100Z 13 Apr to 1100Z 14 Apr, rest period 0500 to 1000Z. Send signal report, QSO number, ARRL section or Ct. County for stations worked inside CT. Club station W1QI counts 5 points per band/mode. Mail by 5 May 1985 (SASE for results) to CARA c/o R. Dillon, N2EFA, Box 143, Bethel, CT 06801.

MAY 18: ARMED FORCES DAY. The annual Armed Forces Day Communication Test. CW, SSB, RTTY and SSTV. Cross band contacts — military to Amateur cross band operations 18/1300 UTC to 19/0245 UTC May 1985. Military stations participating in cross band operations: Air, NMH, NPL, NAM, NMN, NZJ, NAV, NPG, WAR. Military stations will transmit on select frequencies and announce the specific Amateur band frequency being monitored. The CW and RTTY broadcasts will be a special Armed Forces Day message from the Secretary of Defense. Transcriptions of CW and/or RTTY receiving tests should be submitted "as received". Time, frequency and call sign of military station copied and name, call sign and address of individual submitting entry must be indicated. Entries must be postmarked no later than 25 May 1985. Send to following military commands: AIR — AFD Test, 2045CG/DONJM, Andrews AFB, DC 20331-5000. NAM, NAV, NPG — AFD Test, 4401 Massachusetts Ave., Washington, DC 20390-5290. WAR-AFD Test, Commander, USAISC, Att: AS-OPS-CM, Ft. Huachuca, AZ 85613-5000.

APRIL 10, 11, 17, 18: All licensed women operators throughout the world are invited to participate in the DX-YL North American YL contest. DX YLs call "CQ North American YL" and NA YLs call "CQ DX YL". All bands may be used. Stations may be worked and counted once on each band and mode. Exchange station worked, QSO number, RS(T) state or country. For more information: Marty Silver, NY4H, 3118 Eton Road, Raleigh, NC 27608.

MAY 1, 2: Indiana Month of May Contest: Be the first to work 500 Indiana contacts. Exchange RST, state, privince, or country, name and county (Indiana stations). Send copy of log, dup sheet and score sheet by June 30, 1985 to Russ Ryle, N9DHX, Southern Indiana QRP Group, PO Box 2466, Bloomington, IN 47402.

APRIL 30: Amateur Radio operators will have a rare opportunity to work an English Renaissance sailing ship when the Godspeed sails on a 10 week voyage from London, England to Jamestown, Virginia. The original Godspeed was one of three square-rigged vessels which brought the first permanent English settlers to the New World in the winter of 1606-7. Rigging the ship's radio systems has been coordinated by Neil Tanner, WA4CHQ. The Captain of the Godspeed is George Salley, KA4FVB. Special QSL cards have been designed. For more information contact Jamestown/Yorktown Fdn., PO Drawer JF, Williamsburg, VA 23187.

APRIL 20: Spring SSB Contest. 1200 UTC Saturday, to 2400 UTC Sunday. May operate a maximum of 24 hours. Exchanges: Members give RS, state/province/country and QRP ARCI membership number. Non-members give RS, state/province/country and power output. Stations may be worked once per band. Separate log sheets for each band must be received by May 21, 1985. Send logs to QRP ARCI Contest Chairman Eugene Smith, KA5NLY, PO Box 55010, Little Rock, AR 72225.

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NEW products

antenna bridge

The new MFJ-204 Antenna Bridge lets you trim your antenna quickly and easily for its best performance.

The antenna bridge will give an accurate reading of your antenna resistance up to 500 ohms and will cover all the ham bands up to 30 MHz. When used to measure the resonant frequency of your antenna, it allows checking to see if the resonant frequency is higher or lower than desired. You can then lengthen or shorten your antenna based on the information gathered.

Priced at $79.95, plus $4.00 shipping, the MFJ-204 Antenna Bridge has a frequency counter jack for precise frequency measurement and can also be used as a signal generator. Housed in a sturdy black aluminum cabinet, the unit is very compact (4 × 2 × 2 inches) and operates on a single 9-volt battery or 110 VAC using MFJ's AC adapter (MFJ-1312, $9.95).

For further information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #312 on Reader Service Card.

low/medium power coax attenuators

A series of 33 broadband, high-performance attenuators with six continuous-power ratings from 2 to 75 watts is now available from Bird Electronic Corporation. The 8300 series of 50-ohm fixed attenuators covers a frequency range from DC-4 GHz with a VSWR below 1.25, except for the 75-watt and the 2-watt units, which cover DC-2GHz. Models 8308 (75 watts), 8306 (25 watts), 8305 (15 watts), 8304 (10 watts), and 8303 (5 watts) are rated at a peak power of 3 kW with pulses to 5 microseconds wide, are available with attenuation levels of 3, 6, 10, 20, or 30 dB and have male N input/female N output connectors.

The 8300 series attenuators can be used in tandem for odd dB values, or in connection with Bird's TENULINE® high power attenuator (to 4000 watts) for additional attenuation.

For details, contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139.

Circle #313 on Reader Service Card.

low noise amplifier

The R.L. Drake Company has announced the introduction of a new low-noise amplifier for consumer satellite television and other applications. The new Model 2574 provides a noise temperature better than 100 degrees Kelvin and utilizes a 15-volt DC power supply with an output of 3.7 to 4.2 GHz.

The price of the Model 2574 LNA is $195.00 (retail).

For further information, contact R.L. Drake, 540 Richard Street, Miamisburg, Ohio 45342.

STV filtering

Phantom Engineering, Inc. has introduced an improved version of their popular variable bandwidth filter, the IFP-1. The new IFP-1X replaces the IFP-1. It still features the fingertip IF bandwidth selection and IF gain control. The 70-MHz filter allows the bandwidth on the user's receiver to be progressively narrowed with the IFP-1X's four-position signal selector which, in cases of small dishes and/or terrestrial interference, has shown remarkable results.

The improvement comes from the ability of the IFP-1X to pass all control signals below 10 MHz that are used on quartz synthesized type receivers. Another improvement is the addition of the power supply with each unit at no additional cost.

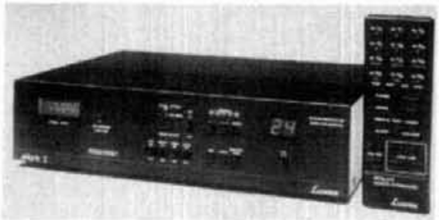
For more information, contact Phantom Engineering, Inc., 16840 Joleen Way, Bldg. E, Morgan Hill, California 95037.

Circle #304 on Reader Service Card.

STV receiver

Luxor North America Corp.'s new Mark 2 remote-control STV receiver is a single integrated satellite component. With a built-in stereo processor and Dolby noise reduction, its advanced "block conversion" technology allows several TV sets in home, building, or neighborhood to share one STV antenna while enjoying independent channel selection.

Designed for use with regular, component TV, VCR, home or studio stereo, signal descrambler, dish positioning systems, and other audio-video equipment, Mark 2 is easily preprogrammed and



operated by the infrared hand-held Remote Commander. An optional Remote Infrared Sensor allows use of the Commander in a room different from the receiver. An antenna positioning system can be remote-controlled. Digital channel and LED tuning indicators, and a signal strength meter, monitor receiver performance.

For details, contact Luxor North America Corp., P.O. Box 32, Bellevue, Washington 98009-0032.

Circle #315 on Reader Service Card.

tuning indicator kit

Heath Company has expanded its Amateur Radio line to include the HD-3006 Crossfire Tuning Indicator for quick and easy tuning of RTTY transmissions. Sixteen LEDs make up the HD-3006's visual display: eight vertical LEDs identify mark signal strength; eight horizontal LEDs indicate space signal strength. Tuning the HD-3006 for maximum vertical and horizontal display provides a strong signal for computers or RTTY printers. Each LED bar requires approximately 14 dB no-signal to signal voltage ratio for full operation. Minimum input signal is 0.3 VAC RMS or 0.5 VDC. Maximum signal is 15 VAC RMS or 15 VDC.

The HD-3006 Crossfire Tuning Indicator has a wide voltage range and is compatible with almost any interface/terminal unit that has oscilloscope outputs for tuning. The AC/DC cube-type power supply is included in the kit.

For information and a free catalog, contact Heath Company, Dept. 150-435, Benton Harbor, Michigan 49022. In Canada, write Heath Company, 1020 Islington Avenue, Dept. 3100, Toronto, Ontario, M8Z 5Z3.

Circle #314 on Reader Service Card.

FCC/VE test guides

Test guides for every Amateur Radio class of license are now available from Gordon West's Radio School.

The test guides list all 500 test questions plus the multiple-choice answers in an attractive 8-1/2 × 11 inch manual. The questions, the distractors (wrong answers) and the right answers are listed exactly as they will be found on ARRl or W5YI volunteer examinations. The General and Advanced class test guides list 500 questions; the Extra class test guide, 400; and the Novice class test guide, 100.

Each test guide also includes study notes listing reference material from which the questions were derived and sources of further information about the answers. Formulas for solving the problems are also incorporated into each test guide.

Also included are several pages of instructions to the applicant on locating a Volunteer Exam Coordinator and how to sign up for a local volunteer-administered examination. Also included are the necessary test forms that applicants must fill out ahead of time, including the new FCC Form 610 (revised).

All test guides have been updated to reflect recent rewordings of FCC test questions. This will allow students to spot any format change in any of the FCC-approved questions.

Study guides are priced at $19.95 plus $3.00 postage. (Be sure to specify the license class you want.) Exclusive stereo Radio School 4-set cassette theory tapes are also available at $39.95; each set of four theory tapes includes Amateur Radio "sounds" that help illustrate specific questions on the examinations. (Be sure to specify the license class covered by the theory course you are requesting.)

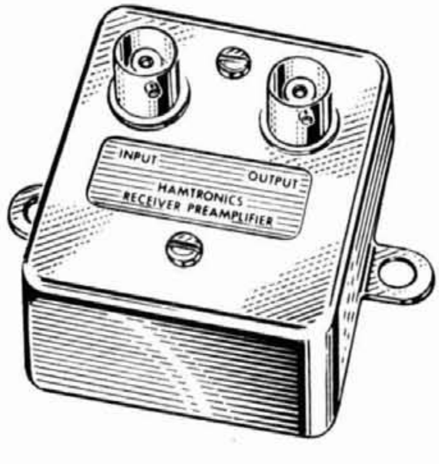
For more information on study guides, code and theory training tapes, and a colorful catalog on instructional materials for volunteers who give the Amateur Radio exams, contact Gordon West, RADIO SCHOOL, 2414 College Drive, Costa Mesa, California 92626.

Circle #302 on Reader Service Card.

NEW products

new Hamtronics® 800 MHz receiver preamplifier

A new 800-960 MHz version of its popular GaAs FET preamp is now available from Hamtronics, Inc. The LNG-800 preamp features a dual-gate GaAs MESFET with built-in diode protection against static discharge damage. The unit has 11 dB of gain with a 1.5 dB noise figure. It is easy to install, operates on 13.6 VDC, and measures only 2 × 2 × 1-1/2 inches. The LNG series of GaAs FET preamps is available also for the high band and UHF band. Preamps in the LNG series, including the LNG-800, cost $49 plus $3 shipping.



For complete information on the LNG preamps, and other Hamtronics products for VHF and UHF, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535.

Circle #107 on Reader Service Card.

J.I.L. SX-400 general coverage scanning monitor receiver

The new SX-400 general coverage scanning monitor receiver from J.I.L. Corporation is designed to cover AM/FM signals in the 26-520 MHz region. This frequency range should be of interest because of the unit's complete coverage of the FM/TV and the 220-MHz band as well as military frequencies above 225 MHz. A complete line of converters for expanding the SX-400's range from 100 kHz to 1.4 GHz — making this unit one of the most versatile scanners on the market — is also available.

The SX-400 has a 20-channel computer controlled memory. Sensitivity is 0.5 μV FM/1 μV AM in the 26-300-MHz range and 0.5 μV FM/2 μV AM in the 300-520-MHz range. The unit has both a fast (8 characters per second) and slow (4 characters per second) scan rate and has a variable 0 to 4-second) delay rate. An automatic noise limiter has been added to reduce pulse noise interference on AM and FM if a filter is used to facilitate the reception of TV and FM broadcast services. For high fidelity reception of signals, channels can be spaced in either 5 or 6.25 kHz steps on VHF and 10 or 12.5 kHz steps on UHF.

An easy-to-use keypad is included for programming frequencies. The unit weighs approximately 7-3/4 pounds (3.5 kg), measures 11.8 × 3.5 × 8.3 inches (30 × 9 × 21 cm) and runs on 13.8 VDC.

The SX-400 is directly interfaceable through the RC-4000 data interface to the NEC-8801A computer. This capability allows both high speed reprogramming of the scanner's channels and an almost unlimited number of channels to be contained in the computer's memory. The computer also provides a complete record of which frequencies were received and the time at which they were received.

Other accessories for the SX-400 include an 800-1400 MHz downconverter for AM/FM (RF-8014); a 500-800 MHz downconverter for AM/FM (RF-5080); a 100 kHz-30 MHz upconverter for AM/SSB/CW (RF-1030); RF attenuators, AF gain control, Delta tuning, IF noise blanker, provision for three external antennas, and squelch. When using the RF-1030 in conjunction with the other converters CW/SSB and AM/FM in the 100 kHz-1.4 GHz range can be received. The suggested retail price is $574.90.

For more information on the SX-400 and its accessories, contact J.I.L. Industries, 17120 Edwards Road, Cerritos, California 90701.

Circle #303 on Reader Service Card.

Butternut Electronics HF2V

Butternut Electronics has just released a new two-band vertical antenna, the HF2V. Designed for 80 and 40 meters, the HF2V can be modified to operate on 160, 30, and 20 meters. The overall height is 32 feet (9.75 meters) and the antenna weighs 13 pounds (5.9 kg). The HF2V is designed to match 50-ohm cable and will give a 2:1 or less VSWR bandwidth of approximately 65 kHz on 80/75 and full-band coverage on 40 meters. Its power rating is 2 kW PEP/1 kW CW (slightly less when using the 160-meter base resonator). The HF2V is designed to operate as a 1/4-wave vertical on 40 meters and a loaded antenna on 80 meters.

For more information, contact Butternut Electronics, 405 E. Market Street, Lockhart, Texas 78644.

NEW products

automatic gain control (AGC-4)

The AGC-4 automatic gain control kit from Barrett Electronics is for any Amateur who needs to keep audio levels constant. In addition to the obvious repeater, autopatch, phone patch, or HF transmitter applications, the AGC-4 can be used in RTTY or SSTV reception with minor component adjustments. As featured in the September, 1984, issue of *ham radio*, the AGC-4 will hold a constant output of 2 volts P – P (±2 dB) with any input level ranging from –36 dBm to +10 dBm. At the center of the AGC range, the total harmonic distortion is 0.5 percent or less with a frequency response (–3 dB) that spans 40 Hz to 20 kHz. With the input shorted, the noise floor is –42 dB below the AGC output level. The AGC-4 was originally designed for the Collins/Autogram "IC" series broadcast mixer and can be operated single ended or with user supplied level or impedance matching transformers. The kit includes a drilled 2-3/4 × 1-5/8-inch printed circuit board, parts, and instructions. The unit is priced at $28.00 including U.S. shipping.

For details, contact Barrett Electronics, 525 North 2150 West, West Point, Utah 84015.

Circle #316 on Reader Service Card.

surveillance receivers

A new generation of HF surveillance receivers has been introduced by Cubic Communications with the R-3030, which features two completely independent receivers in the same rack mount normally required for a single unit.

During the first nine months of introduction, the U. S. Navy contracted for nearly 1000 receivers in the series. Principal selling points included size, light weight, and a wide range of standard advanced features.

The basic receiver is fully functional without costly options, although customer-specified enhancements such as a special data bus or up to six selectable bandwidths also are available.

The advanced modular design (in which plug-in modules are secured by 1/4-turn fasteners and can be replaced in 30 seconds or less) simplifies maintenance and improves operation. In addition, each module is independently shielded to protect circuits from electromagnetic interference as well as potential handling and storage damage.

In shock tests conducted for the Navy, the R-3030 demonstrated its rugged construction by passing a full 2000-foot-pound "hammer test" with a level A rating. The receiver was bolted to a platform, which then was struck by a 400-pound hammer swinging in successive arcs of one, three and five feet. The test was then repeated with the R-3030 turned 90 degrees.

The Navy's "level A" rating means the unit not only retained all components in place, but also continued to work perfectly after the shocks, which are meant to simulate the force of a direct hit to the ship by a non-nuclear torpedo. The reliability level resulting from this rugged construction means more than 5000 hours mean time between failures.

The R-3030's expanded performance features include: tuning in less than 8 milliseconds over a full 5 kHz-30 MHz range, 30 percent fewer parts than conventional radios and capability for full computer control. The modular, building-block approach makes it easier to meet special requirements for data bus connectors, bandwidths, AGC settings and so on.

Fault isolation is both simple and comprehensive, providing quick detection and easy replacement. All faults are reported automatically in three simultaneous modes: data bus, front panel annunciator, and LED on the faulty module. Field repair requires only removing the specified independently shielded module and plugging in a replacement. Each module is labeled and marked with a diagonal coding stripe to prevent improper installation. No special tools, alignment or adjustment are required.

The R-3030 also provides such standard features as: 100 memory channels, IEEE-488 or RS-232 bus connector modules, EMI/EMC shielding, five bandwidths (0.5-8 kHz), five operating modes (LSB, USB, AM, FM, and CW) versatile sweep and scan modes and minimal power input (approximately 35 watts per receiver).

The 48-pound dual system in a compact 5-1/4 × 19 inch rack chassis saves valuable operations and parts storage space in the field while offering effective coverage of virtually any general purpose or surveillance requirement.

For more information, contact Cubic Communications, 305 Airport Road, Oceanside, California 92054-1297.

Circle #306 on Reader Service Card.

transfer lettering fixative

Once a curiosity, dry transfer lettering is now part of the design engineer's and draftsman's tool kit. Even though the letters stick to virtually all surfaces, permanence can be a problem. Unprotected transfer lettering wears off with use. General purpose aerosol sprays aren't much help, because they contain aggressive solvents that cause dry transfers to dissolve, wrinkle or even float out of position before repeat coats can be applied to build up a protective film.

Two products from DATAK can help provide greater permanence: Dakakoat™ acrylic spray can be used to protect transfers applied to painted or unpainted surfaces on metal and plastics. It will not, however, form a smooth film on porous materials such as paper and tracing vellum; Datakoat rests on top of the transfer, but *will sink into the surrounding areas of porous* materials.

The transfer lettering on porous surfces can be protected with Hardkoat,™ a unique spray for use on rubdown transfers applied to paper and other porous surfaces. A single coat softens and penetrates the transfer ink, then glues it to the surface. No additional coats are needed unless severe weathering is expected.

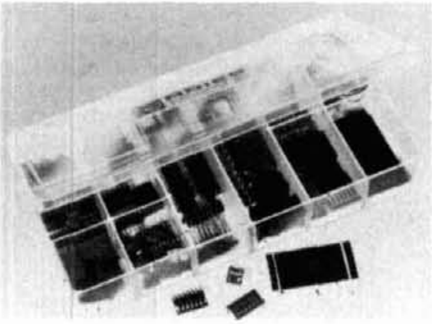
Hardkoat is supplied in 12-ounce spray cans in either gloss or matte finishes. The price for either is $4.75.

For further information, contact The DATAK Corporation, 65 71st Street, Guttenberg, New Jersey 07093.

Circle #307 on Reader Service Card.

engineer's sample case

Aries Electronics, Inc. has made available a Component Engineer's sample case, Part No. SB-100, containing over 100 pieces of various connector products the company makes. Included are sockets, Vertisockets,® elevator



sockets and single-row sockets (both stamped and collet pin versions), headers, programmable headers, switches, shorting plugs, jumper assemblies, etc. Worth over $100 if purchased individually, these parts come in a handy plastic case and sell for $30.00.

For additional information, contact Aries Electronics, Inc., P.O. Box 130, Frenchtown, New Jersey 08825.

Circle **#308 on Reader Service Card.**

DTMF encoder mike

Midland LMR has introduced an optional DTMF encoder microphone for its Midland SYN-TECH™ line of 2-way FM mobile radios. Available with or without automatic number identifier (ANI) capabilities, the new dynamic amplified microphones incorporate an integral DTMF encoder capable of generating the 16 standard DTMF digits 0-9 plus *, #, A, B, C and D. The

NEW products

company had previously announced availability of a DTMF decode option for its SYN-TECH radios.

The new SYN-TECH DTMF encoder microphones have a front-mounted keypad which can be enabled and disabled by an on-off switch, with current status indicated by a red LED. An internal annunciator provides audible confirmation of each keypad switch closure. Single tones can be generated, also, by simultaneously pressing two keys in the same column or row. Top-located up/down switches allow direct control of channel selection from the microphone.

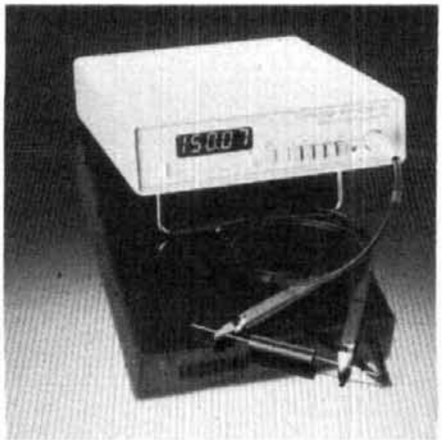
The SYN-TECH DTMF microphone with ANI capability features an ANI output which is jumper selectable from one to eight digits. The ANI sequence is automatically sent at the beginning of a transmission if a pre-set time interval has elapsed since the previous transmission. The ANI code sequence can be activated at any time by pressing the * or # key, and can be "strapped" to give single or multiple sequences when activated.

For more information on the new Midland SYN-TECH DTMF options, contact Midland LMR, Marketing Department, 1690 N. Topping, Kansas City, Missouri 64120.

Circle #309 on Reader Service Card.

micro-ohmmeter

Cambridge Technology, Inc. has introduced the Model 510, a low-cost, 4-1/2 digit, micro-ohmmeter designed to measure the resistances of switch and relay contacts, transformer and motor windings, connectors, or any other low



resistance devices. It has five ranges from 19.999 milliohms to 199.99 ohms, full-scale, 1 micro-ohm resolution, and a basic accuracy of 0.02 percent.

Three measurement modes are provided. The continuous DC mode is useful for making measurements on inductive components and the switched DC mode removes the effect of thermal voltages, the largest source of error in low resistance measurements. A pulsed mode is provided for thermally sensitive devices such as fuses. The standard unit comes with 4-terminal Kelvin test clips and a parallel BCD interface. Optional limits comparators, battery operation, and an RS-232 interface will be available in the first quarter of 1985.

For further information, contact Cambridge Technology, Inc., 2464 Massachusetts Avenue, Cambridge, Massachusetts 02140.

Circle #311 on Reader Service Card.

automatic phone patch system

NCG Co. has just announced its new Hotline 007 simplex autopatch unit. When connected between your radio and the telephone line, it enables initiation and reception of telephone calls without operator assistance. The Hotline 007 uses a unique method of signal processing that eliminates annoying squelch tails and chirps that are heard on other units. The unit uses a field programmable five digit access code to eliminate unauthorized use. In the event that you drive out of communications range, the Hotline 007 has a variable 3 to 12 minute time-out timer. When someone is calling, the Hotline 007 will page you with a CW message. To answer the call, simply send your 5-digit access code and the Hotline 007 will connect you to the phone call. You can also program the unit to refuse any calls that start with a 0 or 1. NCG also has DTMF microphones and telephone handsets available as options.

For further information and prices, contact NCG Co., 1275 N. Grove, Anaheim, California 92806.

Circle #301 on Reader Service Card.

one millionth satellite earth station installed

The Board of Directors of the Society For Private and Commercial Earth Stations (SPACE) recently celebrated the installation of the one millionth satellite earth station. SPACE is the National Trade Association representing the Manufacturers, distributors, and dealers of satellite earth stations.

Information about the Society for Private and Commercial Earth Stations, (SPACE), is available from SPACE, 709 Pendleton Street, Alexandria, Virginia 22314.

The Problem Solver. . .

The RF Wattmeter Model 81000-A from Coaxial Dynamics, Inc. does more than provide accurate rf measurements. Testing of transmission lines, antennas, connectors, filters and related components can reveal unknown problems and assure optimum equipment performance.

The 81000-AK Wattkit features this easy-to-read RF Wattmeter (pictured here), with its optional carrying case and an array of elements and accessories. Coaxial Dynamics elements can be purchased separately for use in other manufacturer's Wattmeters. For more information on the 81000-A Wattmeter or any of the complete line of Coaxial Dynamics RF products and OEM components please contact Coaxial Dynamics, Inc.

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196-214	"	BNC connector	6.95
196-224	144-UP	BNC conn. adj. angle	7.95
196-814	220-225	BNC connector	6.95

5/8 WAVELENGTH

Model No.	Freq. MHz	Description	Price
191-210	"	5/16-32 for old TEMPO	22.95
191-214	"	BNC connector	19.95
191-219	"	PL-259 w/M-359 adpt.	22.95
191-810	220-225	5/16-32 for old TEMPO	22.95
191-814	"	BNC connector	19.95
191-940	440-450	5/16-32 for HT-220	22.95
191-941	"	1/4-32 stud	22.95
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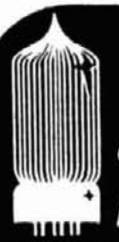

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More Details? CHECK—OFF Page 160

THE GUERRI REPORT

Ernie Guerri
W6MGI

toward "softer" software

Early electronic data processing systems required that people communicate with them in their own language — *machine* language. But it soon became clear that communication between people and machines in binary, or some related code, was *cumbersome* — for the *people* — and "higher-level" languages had to be devised that would be more recognizable to people and would automatically do conversion to machine language. FORTRAN, COBOL, BASIC, *Ada*, and LISP are all the products of a relentless move toward computer languages that are closer to human experience, yet still accommodate more complex machine architectures. This trend toward "softer" software is aimed primarily at user friendliness — the keep-it-simple concept. But there's a price to pay. Not everything is simple. In an attempt to simplify the languages, software architects have to choose the complexity of the functions they want to implement with relatively simple program statements. In this simplification process much computational power can be lost.

Computer architects and software designers must now work hand-in-hand to assure that commands given by humans can be interpreted by machines whose logic has been optimally structured to perform specific algorithmic functions. As machines become more complex, finding the best combination of machine structure and software language is more difficult. If we structure the machines to easily interpret "plain English," then the machines may be limited by what can be said in English. A compromise between language simplicity and machine functionality is in order. The best compromise has yet to be determined.

RF sonar serves medical needs

During 1985 some of the key biological functions of astronauts aboard the U.S. Space Shuttle will be monitored by a process called *echography*. In this process, body functions are observed by radio frequency sonar operating in the 3 to 5 MHz region. A small antenna probe is used to illuminate the tissue at very low power, and the echo returns form a time-sequenced image of the tissue. The data is stored as real-time video, and can then be digitized and image-enhanced. Resolution is very good; images are formed at rates up to 50 per second, and time-motion analysis can be updated as often as 5000 times per second. This can give very detailed information about heart movement and the flow of blood through vital organs. (You fellows with 75-meter beams, please — no elevation rotators!)

synthetic rocks

Researchers in Australia have developed a synthetic rock material that may have important application in the disposal of radioactive material. Major problems involved in disposing of high-level radioactive waste include site safety, disintegration of surrounding material, and the ability of the waste containers to remain intact for the required 10,000+ years. It's not too difficult to find geographical areas which can be extrapolated to remain stable for several hundred thousand years, but finding containers whose atomic structure is not destroyed by years of intense radiation has been an elusive process.

The newly developed "Synroc," as its developers call it, is made of three natural minerals whose atomic structure is such that high-level radiation converts the rock to a glassy material. The glassy material is atomically stable and resists cracking and subsequent leakage. The Los Alamos National laboratory is now evaluating the material by embedding high-level waste with a short half-life (less than a year) into the Synroc and then extrapolating the results. Considering the very long half-life of the more common waste, the validity of extrapolation will have to be shown. In any case, this may be a major development in what has otherwise been an unyielding problem.

GaAs high speed ICs make progress

Silicon semiconductor material has been the mainstay of the modern electronics industry for about 25 years. It has been continually improved and now offers good speed, high yields, and best of all, low cost. However, we are now pushing the limits of silicon physics and device geometry to get further improvements. Gallium Arsenide (GaAs) and an associated family of semiconductor materials offer the higher speeds made possible by high electron mobilities. Microwave GaAs devices have been available for some time, but the price has been fairly high due to the higher material cost and poor yields. Obtaining flat, defect-free epitaxial wafers for the fabrication of GaAs ICs has been a very difficult and costly process.

However, new techniques permit material growth and processing with sufficient quality to permit small-scale integration on a commercial basis. Simple digital circuits containing tens of devices are available in the market, and arrays with a few hundred elements will be readily available during 1985. Work is progressing on very high-speed memories with cycle times of 500 picoseconds, and complex functional circuits with 1-GHz digital and 5-GHz analog circuits sharing the same real estate should be available in another year or two.

This next generation of circuit developments will offer yet another milestone in speed and capability in practically every domain of telecommunications. Amateurs have already seen the benefits of GaAs devices in low-noise VHF/UHF amplifiers, and will next see comparable performance in the signal processing sections of new radio equipment.

ham radio

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