


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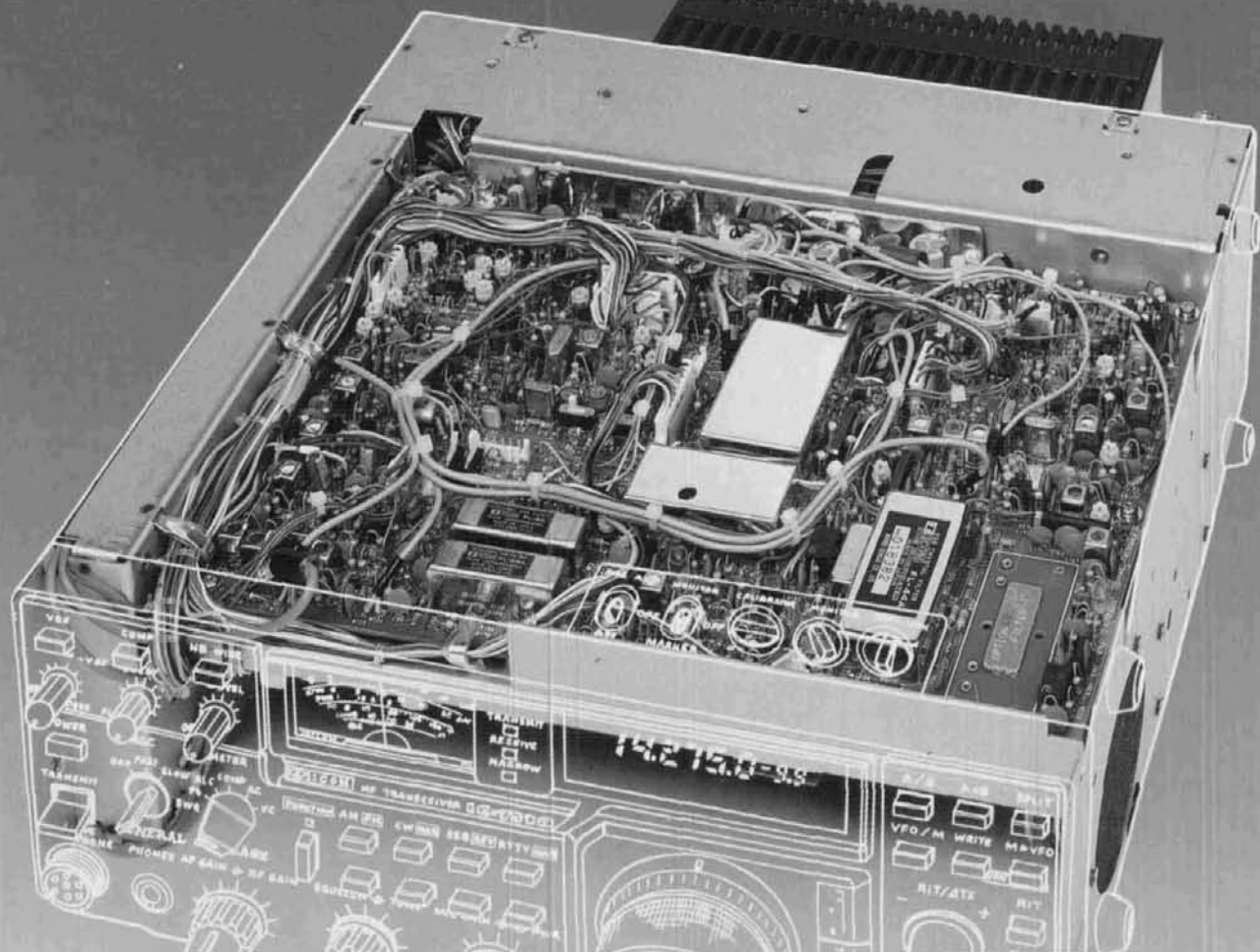
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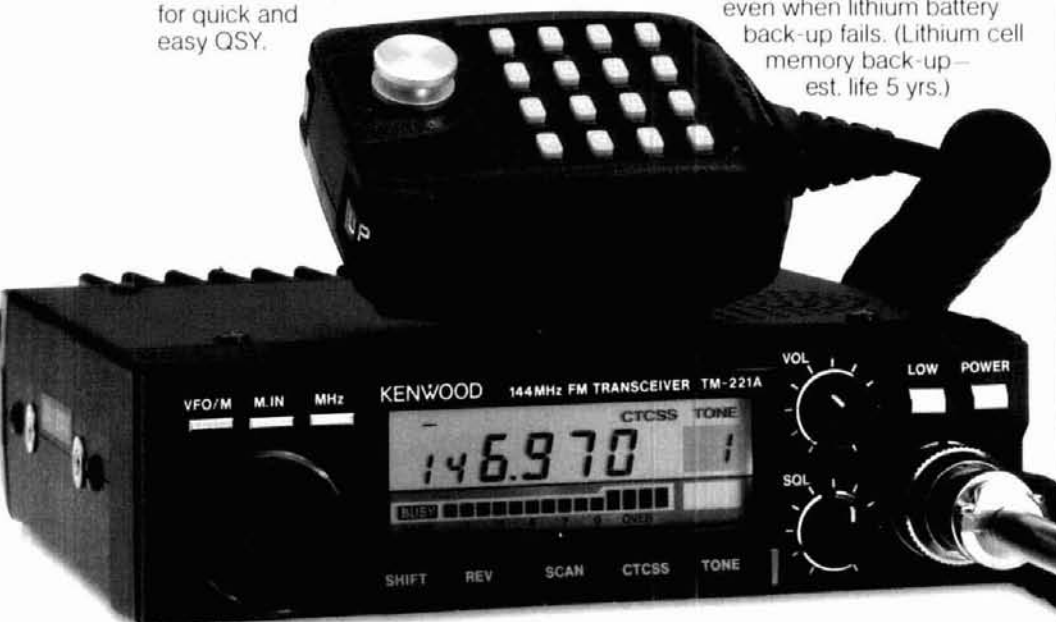
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Optional telephone-style handset remote controller RC-10 is specially designed for mobile convenience and safety. All front panel controls (except DC power and RF output selection) are controllable from the RC-10. One RC-10 can be attached to **either or both** TM-221A and TM-421A with the optional PG-4G cable. When both transceivers are connected to the RC-10, **cross band, full duplex repeater** operation is possible. (A control operator is needed for repeater operation.)



Optional Accessories:

- RC-10 Multi-function handset remote controller
- PG-4G Extra control cable, allows TM-221A/TM-421A full duplex operation
- PS-50/PS-430 DC power supplies
- TSU-5 Programmable CTCSS decoder
- SW-100A Compact SWR/power/volt meter (1.8-150 MHz)
- SW-100B Compact SWR/power/volt meter (140-450 MHz)
- SW-200A SWR/power meter (1.8-150 MHz)
- SW-200B SWR/power meter (140-450 MHz)
- SWT-1 Compact 2 m

- antenna tuner (200 W PEP)
- SWT-2 Compact 70 cm antenna tuner (200 W PEP)
- SP-40 Compact mobile speaker
- SP-50B Mobile speaker
- PG-2N Extra DC cable
- PG-3B DC line noise filter
- MC-60A, MC-80, MC-85 Base station mics.
- MC-55 (8-pin) Mobile mic. with gooseneck and time-out timer
- MA-4000 Dual band antenna with duplexer (mount not supplied)
- MB-201 Extra mobile mount

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contents

8 real coax:
impedance and phase relationships
Forrest Gehrke, K2BT

19 simple modifications and adjustments
for the TS-930S
Marv Gonsior, W6FR

24 packet radio conference bridge
Bill Wade, WD5HJP

29 the TEXNET packet-switching
network
part 2: hardware design
Thomas H. Aschenbrenner, WB5PUC, and Thomas
C. McDermott, N5EG

47 practically speaking
grounding and shielding, part 1
Joe Carr, K4IPV

55 ham radio techniques:
Xingiang Province: the last frontier
Bill Orr, W6SAI

63 a three-tube 4CX250B linear amplifier
Dragoslav Dobricic, YU1AW

74 VHF/UHF world
33 cm: update
Joe Reisert, W1JR

89 understanding noise figure
Mike Gruchalla

118 advertisers index
and reader service
6 comments
97 DX forecaster
116 flea market

114 ham mart
102 new products
4 reflections
50 short circuits



**220 AT RISK —
ACT NOW**

comments close April 6.

REFLECTIONS
REFLECTIONS
REFLECTIONS

novice enhancement and the 220 MHz band: the FCC giveth...

Novices are back on phone — for the first time since the 1960s — as this issue of *ham radio* arrives in your mailbox. In its Report and Order of Novice Enhancement (effective March 21, 1987), the FCC expanded the Novice 10-meter allocation to 28.5 MHz and awarded new VHF/UHF allocations at 1-1/4 meters and 23 centimeters. Phone operation is now permitted on all three bands.

In brief, on 10 meters Novices (and Techs, who share Novice low band privileges) may now use digital (F1B) as well as CW from 28.1 to 28.3 MHz, plus CW and SSB from 28.3 to 28.5 MHz. The power limit remains 200 watts. On 1-1/4 meters, they're authorized all modes from 222.1 to 223.91 MHz, with a 25-watt power limit. On 23 centimeters, the Novice band is 1270-1295 MHz (all modes), but with only 5 watts.

There's one important limitation: while Novices will be permitted to *use* repeaters, they won't be permitted to be repeater control operators or trustees. Note: contrary to a widely circulated but erroneous interpretation of the new rules, Novices' signals may be repeated outside the Novice subbands, just as Techs' signals can be repeated on 10-meter cross-linked repeaters or through the satellites. This means that Novices will be able to communicate directly with higher class Amateurs on the other VHF/UHF bands through cross-linked repeaters.

They'll have to pass a more difficult exam, however. Instead of 20 questions, they'll face 30 — with the additional ten testing knowledge relevant to the new privileges.

So what's this going to mean for Amateur Radio? First, it means there are now 80,000+ Amateurs with phone and digital privileges on 10 meters and first-time access to all modes on VHF and UHF. Despite the diminished sunspot cycle activity, 10-meter operations should see a big boost. At its worst, 10 generally shows some life — and greatly increased activity on a "dead" band often confirms that its not really dead, but only lacking users! One-and-a-quarter meters should benefit at least as much as 10 — with equipment costs and performance about the same as 2 meters, and a well-established repeater system in place, this band could even become as busy as 2 meters! One unknown is just how well the newcomers will be received; the message is mixed on this, with some repeater groups already planning cross-band links to further expand Novice horizons, while others vow they'll install elaborate coding systems to keep Novices off their machines — or even shut their repeaters down entirely.

Twenty-three centimeters will also see some increase in activity, with good quality commercial equipment available from several suppliers. However, costs are high and activity low, so the effects here are not likely to be nearly as extensive.

What new Novice privileges will mean to the Amateur Radio service is yet to be determined. Though enhancement doesn't make the Novice a no-code entry-level license, it does radically augment its benefits by the addition of phone privileges. If this results in a big boost in newcomers, it may temper enthusiasm for a code-free license — perhaps the perceived problem with the Novice license hasn't been its requirement for 5 wpm code capability, but rather that its only former benefit was to permit the holder to work CW (and CW is *work* when you're new at it).

One thing enhancement seems sure to do is increase activity among present Novices, and that's good for Amateur Radio. The dropout level for Novices has always been unacceptably high, but the new phone privileges should substantially change that. Why? Because Novices will find both their investment (in new equipment) and return on that investment (fun) increasing to the point where they'll not only be unwilling to let their licenses expire, but will be stimulated to upgrade. All things considered, then, enhancement appears to be a big plus for Amateur Radio.

Before we become too euphoric, however, let's consider this:

...and the FCC taketh away!

Excitement over Novice enhancement was still growing — in fact, the news was still being spread — when the FCC dropped the other shoe . . . squarely on 220 MHz. In a Notice of Proposed Rule Making dated just two days after the Report and Order on Novice Enhancement, the Commission proposed reallocating 220-222 MHz to land mobile service use only!

As justification, the FCC cites "light loading" of the band in comparison to 2 meters, using ARRL's Repeater Directory listings as evidence. Even assuming this "evidence" is accurate (which it isn't, since the 220-MHz band supports many control links and a large, dynamic, and growing number of packet and other data communication users who've come to 220 MHz because 2 meters couldn't accommodate them), it's obvious the Commission's right hand doesn't know what its left is doing! You don't give 80,000 new users access to a finite band of frequencies, then cut that band by 40 percent! It's worth noting, at this point, that this proposal *did not* come from the FCC's Personal Radio Bureau, where Amateur Radio and our problems are well known and appreciated, but from the Office of Engineering and Technology, where they obviously aren't!

That's not to say this proposal wouldn't have made better sense 10 years ago. In fact, back then it might have even been to our advantage. Why? Because, along with taking 220-222 MHz away from Amateurs, the FCC is also proposing removing fixed and mobile service allocations from the 222-225 MHz segment and making that portion solely Amateur (though shared

(continued on page 50)

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70 cm

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Compact all mode transceivers

It's the "New Sound" on the 2 meter band—Kenwood's TR-751A! Automatic mode selection, versatile scanning functions, illuminated multi-function LCD and status lights all contribute to the rig's ease-of-operation. All this and more in a compact package for VHF stations on-the-go!

• Automatic mode selection, plus LSB
144.0 144.1 144.5 145.8 146.0 148.0 MHz

CW	USB	FM	USB	FM
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- Optional front panel-selectable 38-tone CTCSS encoder
- Frequency range 142-149 MHz (modifiable to cover 141-151 MHz)
- High performance receiver with GaAs FET front end
- VS-1 voice synthesizer option

- 25 watts high/5 watts adjustable low
- Programmable scanning—memory, band, or mode scan with "COM" channel and priority alert
- 10 memory channels for frequency, mode, CTCSS tone, offset. Two channels for odd splits.
- All mode squelch, noise blanker, and RIT
- Easy-to-read analog S & RF meter

- Dual digital VFOs
- Semi break-in CW with side tone
- MC-48 16-key DTMF hand microphone and microphone hook included
- Frequency lock, offset, reverse switches
- Digital Channel Link (DCL) option

Optional accessories:

- CD-10 call sign display
- PS-430, PS-30 DC power supplies
- SW-100A/B SWR/power meter
- SW-200A/B SWR/power meter
- SWT-1 2 m antenna tuner
- SWT-2 70 cm antenna tuner
- TU-7 38-tone CTCSS encoder
- MU-1 modem unit for DCL system
- VS-1 voice synthesizer
- MB-10 extra mobile mount
- SP-40, SP-50B mobile speakers
- PG-2N extra DC cable
- PG-3B DC line noise filter
- MC-60A, MC-80, MC-85 deluxe base station mics.
- MC-43S UP/DOWN mic.
- MC-55 (8-pin) mobile mic.
- MA-4000 dual band antenna with duplexer



Actual size front panel

TR-851A

70 cm SSB/CW/FM transceiver

The same winning features are yours on 70 cm with the TR-851A!

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comments

thoroughly debated by the membership.

**Dick Green, KA1LBW,
and Mike McAmis, WA3ECT**
Etna, New Hampshire 03750

a question of goals

Dear HR:

When we read about the [ARRL Board's approval in principle of a new Amateur Radio museum ("It Seems to Us," *QST*, October, 1986), we thought it was a nice idea.

Then we read an editorial by Craig Clark, N1ACH, in *ham radio* ("Reflections," October, 1986) and kicked ourselves for not thinking about this issue more carefully.

Craig pointed out, we think correctly, that the project will consume enormous amounts of time, energy, and money. In light of the Board's ambitious goals, set in 1984, to recruit new hams and ARRL members, it just doesn't make sense to divert our focus to an equally ambitious undertaking — one that probably will not increase the number of new licensees significantly.

We're all for keeping HQ and W1AW in good shape, but how many of us get a chance to visit Newington? Will a museum in Connecticut inspire non-hams across the nation to get a license? Wouldn't it be better to raise and spend \$2.7 million on recruiting new hams?

For that kind of money we could buy, stock, staff and operate several mobile displays which could present ham radio exhibitions and give license exams at schools, museums, fairs, clubs, and shopping centers throughout the country. Let's take Amateur Radio to the people instead of expecting them to come to us.

We urge all League members to read Craig's editorial and convey their feelings to their Division Directors immediately. This important issue should be

low-band operation

Dear HR:

K2RR's articles in the May and June issues ["Secrets of Successful Low-Band Operation," pages 16 and 17, respectively] prompt me to dash off a few lines on that subject.

Let me preface my remarks by setting the time frame and locations for my experiences on 80 and 40 meters. In the years 1956 through 1966, I was fortunate to operate from three locations in Europe and two in North Africa.

I will do my best to recall a few of the QSOs and without reference to the old log books, which are not now at hand.

October, 1958 saw me in North Africa on 40 meters (though on a few occasions, I did get down on 80 meters). One memorable QSO was with the late Jim Mills, ZL2BE. Jim ran several big rhombics, and knew that I could "borrow" some even bigger ones from the military. Well, to make a long story short, we had a QSO or two, using long path, short path, and probing the other paths as well. Hams at both ends of the circuit were somewhat amazed to hear us on what they considered a dead band. Needless to say, I have fond memories of QSOs like that, so many years ago. No one is more pleased than me to see the tremendous strides that have been made in the design of antennas and equipment in recent years.

Now, let me jump ahead to the 1960s and recount a few experiences on 80 meters. On 80, most any evening, I would be working other Europeans. We, of course, were in darkness, but the east coast of the U.S. was still in daylight. As the evening wore on, out of the noise would come the signal of, say, W1FRR (now

W1FC). Fred would say, "Hey, Warren, I've been calling you for an hour now. Didn't you hear me?" My retort: "No, Freddie, not a peep until just now."

One-way skip? Perhaps so, but how do you explain it? Was my signal going to the "F" layer, bouncing off the Atlantic in mid-path, then traveling in the "E" layer and then down to Fred? Theory has it that since my "E" layer had dissipated, and Fred's had not, the early evening path employed both "F" and "E" layer from east to west, or dark to daylight.

Now, let's see what we can do to explain the bending, or azimuthal divergence of radio signals. With the advent of large, directive arrays on the ham bands, Amateurs are experiencing what the commercial stations knew decades ago. Let me give you an example: a commercial point-to-point station in Slough, England, ran daily traffic with a similar station in Capetown, South Africa. The path between them was normally North-South, as you would expect. But on some occasions, both stations would have to slew their antenna arrays to a point off South America. I don't recall when this situation was first observed, nor do I remember when I read about it, but suffice it to say that it was not in recent times. By the way, this effect has occurred on numerous occasions over the past several years.

How do you explain this bending thing? Well, far more knowledgeable persons than I have explained it this way: *bending can take place when the layer is spherically stratified*. What makes this bending different from backscatter or side-scatter is that the signal is not diluted or fragmented, as it may be in a scattered signal.

I have many other thoughts on low-band DXing, but will save those for another time.

Kudos on your fine magazine and the folks who make it all happen.

Warren Lufkin, W0SII
Cheyenne, Wyoming 82001

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YES!
220 MHz

220: Kenwood Style!

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The first comprehensive
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TM-3530A—25 watts of 220 MHz FM—Kenwood style! Features include built-in 7-digit telephone number memory, auto dialer, direct frequency entry and big LCD. All this makes the TM-3530A the most sophisticated rig on 220 MHz!

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- 23-channel memory for offset, frequency and sub-tone



- Big multi-color LCD and back-lit controls for excellent visibility
- Optional front panel programmable 38-tone CTCSS encoder **includes 97.4 Hz**

- Frequency lock switch
- Digital Channel Link (DCL) option
- High performance GaAs FET front end receiver

TH-31BT/31A

Kenwood's advanced technology brings you a new standard in pocket/handheld transceivers!

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- BT Series has built-in tone
- Repeater offset: -1.6 MHz, reverse, simplex
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TH-31BT/31A optional accessories:

- **HMC-1** headset with VOX
- **SMC-30** speaker microphone
- **PB-21** NiCd 180 mAh battery
- **PB-21H** NiCd 500 mAh battery
- **DC-21** DC-DC converter for mobile use
- **BT-2** manganese/alkaline battery case
- **EB-2** external C manganese/alkaline battery case
- **SC-8/8T** soft cases with belt hook
- **TU-6** programmable sub-tone unit
- **AJ-3** thread-loc to BNC female adapter
- **BC-6** 2-pack quick charger
- **BC-2** wall charger for PB-21H
- **RA-9A** StubbyDuk antenna
- **BH-3** belt hook

- 16-key DTMF pad, with audible monitor
- Center-stop tuning—**another Kenwood exclusive!**
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- HI/LOW power switch (adjustable LOW power)



TH-31BT with DTMF pad shown.
Optional RA-9A attached.



TM-3530A optional accessories:

- **TU-7** 38-tone CTCSS encoder
- **MU-1** DCL modem unit
- **VS-1** voice synthesizer
- **PG-2N** extra DC cable
- **PG-3B** DC line noise filter
- **MB-10** extra mobile bracket
- **CD-10** call sign display
- **PS-430** DC power supply
- **MC-60A/MC-80/MC-85** desk mics.
- **MC-48B** extra DTMF mic. with UP/DOWN switch
- **MC-43S** UP/DOWN mic.
- **MC-55** (8 pin) mobile mic. with time-out timer
- **SP-40** compact mobile speaker
- **SP-50B** mobile speaker
- **SW-200B** SWR/power meter
- **SW-100B** compact SWR/power meter

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real coax: impedance and phase relationships

The device between
the transmitter and antenna
deserves better treatment

K2BT, author of the definitive series on phased verticals¹, examines in his own precise manner the extremely important and often neglected component that gets your signal from here to there. — Ed.

We all know that when using coaxial cable at radio frequencies, it may have a significant loss which must be considered. It's quite clear that a cable having a loss of 3 dB/100 feet at 30 MHz will deliver only half the available power to the antenna if we need to use 100 feet. But how many of us are aware that there's a further loss if the load is mismatched to the line? Or that a lossy coax can make a badly matched antenna look good in the shack? In fact, it's possible for a quite low-loss cable, if long enough, to indicate a 1:1 SWR at its input even though the output is open-circuited! And though coax losses at lower frequencies ordinarily can be neglected, as the SWR goes much above 5:1 we still can sustain a non-trivial loss and a different input impedance than we might have expected due to the high mismatch. (Incidentally, additional loss due to load mismatch can occur with any lossy circuit, not just coax.)

Those of you who've seen graphs of additional loss attributable to SWR may have wondered, as I did, just how this loss was calculated. These effects may be appreciated in a general way, but we might be hard pressed to attach numbers to some of these situations. The mathematical formulations dealing with this subject, strewn with Greek letters and unfamiliar trigonometric symbols, give the appearance of such formidability that we are dissuaded from looking any

further. And so we often approach the problem "assuming the lossless case...", knowing this is not reality but hoping we haven't missed it by much.

The purpose of this article is to help with such assessments and to place some of the calculations into a simplified and standardized format for ease of programming computers and calculators.

common approach: interpolation

Trying to assign loss at a given frequency when the ratings are given at other frequencies is the first problem we encounter. Most cable manufacturers publish loss figures at various frequencies — for example: 1, 10, 50, 100 MHz and so on, which are called attenuation constants and defined as the loss in decibels per 100 feet with a matched load. But if the operating frequency is 21 MHz, what now? We can "eyeball" interpolate, but it turns out we can do it to a high degree of precision using the manufacturer's ratings.

definitions

Before proceeding, we must come to some general understanding:

1. A transmission line is a uniform system consisting of two parallel conductors. "Uniform" means that the materials and geometry of the line and its surrounding medium remain constant throughout the length of the line. This applies whether the line is parallel wire, coaxial, stripline, or other configuration. This understanding about uniformity also applies to the electrical circuit coefficients that result from the materials chosen for this line.

2. The currents in the line conductors flow only in the direction of the line. It may seem redundant to state this in discussing distributed circuit transmission lines. However, under certain conditions some or all of the current may propagate around the conductors

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instead of along them. Such transmissions are known as "waveguide modes" and are not explainable by distributed circuit theory. For lines with conductor separations that do not exceed a few inches, these waveguide modes are not supported except at frequencies in the very high GHz range. However, although these modes do not propagate any appreciable distance at the frequencies we are interested in, they are excited by any discontinuities in the line, whether caused by too sharp twists or bends or abrupt physical terminations of the line. These excitations die out quickly from their point of origin, traveling a distance no greater than a few times the conductor separation of the line. However, they are responsible for anomalous behavior that occurs in the vicinity of such discontinuities.

For now, so that "discontinuities" are not simply read as being only characteristic impedance variations, it should be clearly understood these are *physical* discontinuities. For example, the junction of an RG-8 line with an RG-174 line (two lines with the same characteristic impedance but very different physical dimensions) is such a discontinuity.

3. The frequency range of interest is generally 1 MHz to 1 GHz, a range more than sufficient for most Amateur Radio interest. Although coaxial transmission lines are used above and below this range, the equations presented in this article must be used with more knowledge about certain specific characteristics of the line. For example, as the frequency is reduced below 2 MHz, the characteristic impedance of the coax will have an increasingly significant reactive component. This may not be a show-stopper, but including that component in calculations will improve accuracy. As frequency is progressively lowered (less than 100 kHz), not only will we see a more reactive characteristic impedance, but the resistive component will rise — slowly at first — and then more rapidly. As we go above 1 GHz, losses at some point may rise disproportionately and unpredictably because of various problems such as insufficient braid coverage in flexible line allowing power loss by radiation. Waveguide mode anomalies — for example, those due to distortions caused by having been bent around too small a radius — also become more troublesome at these frequencies because the excited modes are operating over what are now larger fractions of wavelengths.

4. All discussion and equations assume steady-state conditions and no active elements.

ABCD network parameters

In a previous article discussing design of drive networks for phased vertical arrays,¹ I presented material on the ABCD network matrix parameters. ABCD matrix algebra was devised for the purpose of simplifying calculations of input/output impedances, current,

and voltage for four-terminal networks and for facilitating cascaded calculations when chaining networks. The parameters are defined in such a manner that whether a passive network is a length of coax, a Pi or T circuit, or anything else, so long as the parameters are identical — regardless of the differences in individual components or hookup — the input/output relationships are unchanged at the frequency for which the parameters are identical. For the purposes of the referenced article, I felt the subject of dealing with real coax to be an undue digression even though definitely related, particularly for phased arrays at higher frequencies, or even at lower frequencies when driving an element through an excessively mismatched line. Indeed, it is to fill in that gap that this article was written.

For the lossless case, the ABCD matrix parameters for a length of coax are:

$$\begin{matrix} A & \begin{bmatrix} \cos \theta & j Z_0 \sin \theta \\ j(\sin \theta)/Z_0 & \cos \theta \end{bmatrix} & B \\ C & & D \end{matrix}$$

θ = Electrical length (degrees or radians)
 Z_0 = Characteristic impedance

In order to account for loss, these parameters must be defined differently. We arrive at them and can go back to the lossless case simply by substituting zero for the loss terms.

$$\begin{matrix} A & \begin{bmatrix} \cosh \Gamma \ell & Z_0 \sinh \Gamma \ell \\ (\sinh \Gamma \ell)/Z_0 & \cosh \Gamma \ell \end{bmatrix} & B \\ C & & D \end{matrix}$$

Γ = $(\alpha + j\beta)$ Propagation constant
 ℓ = Physical length
 Z_0 = Characteristic impedance

At first glance this matrix looks much the same, but there are important differences: the familiar circular functions (sin, cos) are replaced with hyperbolic functions (sinh, cosh), and a term called the "propagation constant" appears for the angle. The propagation constant combines loss and length in the form of complex numbers.

propagation constant

Engineers refer to many of the coefficients involving transmission lines as "constants," a term not well chosen considering the inconstancy of some of them as frequency is changed. The propagation constant consists of two coefficients, the real term α , called the attenuation factor, and the complex imaginary term β , the phase factor. Both are unit factors — that is, they refer to a value per unit length. The calculation for any particular section of line implies a specific length; if the terms of the propagation constant are multiplied by the length in consistent units, whether length is in meters or feet will not matter because the result is total loss and total angle for that length of line. Angular units may be degrees or radians, consistent with the trigonometric functions being used, but the unit of loss must be in nepers.

attenuation factor α

American coaxial cable manufacturers typically use the decibel as the unit of attenuation, usually rating a cable in terms of decibel loss per 100 feet. We'll need to convert these loss figures into nepers. Both the neper and the decibel are logarithmic functions of the ratio of power flowing at two points in a circuit. Their mathematical definitions are:

$$\text{Decibel} = 10 \log_{10} (P1/P2) \quad (1A)$$

$$\text{Neper} = 0.5 \log_e (P1/P2) \quad (1B)$$

The decibel is based on decimal logarithms while the neper uses the natural or Napierian base, 2.71828... which is usually written as e . Appropriately, neper is a Latinized spelling of Napier, in honor of the Scottish mathematician who invented logarithms. The neper is by far the larger unit; for example, a 2:1 ratio of power is expressed as 3.0 decibels but only 0.35 neper. More precise conversion factors are:

$$1 \text{ neper} = 8.686 \text{ decibels}$$

$$1 \text{ decibel} = 0.1151 \text{ neper}$$

Even the decibel is usually too large to be used "per foot" for coax. Consequently, attenuation is quoted in decibels per 100 feet.

Rated loss figures for transmission lines are given in large MHz intervals (1, 10, 50, 100, and so on), which brings us back to the question of an accurate interpolation method for frequencies between them. We can produce an empirical interpolation by taking advantage of the known loss versus frequency characteristics of cable materials. Over the frequency range of interest there are two major loss contributions: loss in the conductors and loss in the dielectric material. We can say then that the total loss is equal to these two components:

$$\alpha_T = \alpha_C + \alpha_D \quad (2)$$

From various references we learn that the loss of round copper wire due to skin effect varies as the square root of the frequency. For any given cable this can be stated as:

$$\alpha_C = m \sqrt{f} \quad (3)$$

where m is a constant for any given cable type. The dielectric loss is found to vary directly with frequency:

$$\alpha_D = nf \quad (4)$$

where n is a constant for a given cable depending on the insulation used, its thickness, and so on. For a first approximation and to the degree that the constants m and n remain independent of frequency:

$$\alpha_T = m \sqrt{f} + nf \quad (5)$$

Using the rated loss α_{T1} and α_{T2} at two frequencies f_1

and f_2 for any given cable and these identities for the loss components, two simultaneous equations can be written. Substituting and solving for m in one equation and then back-substituting to solve for n we get:

$$m = \frac{\alpha_{T1} \cdot f_2 - \alpha_{T2} \cdot f_1}{f_2 \cdot \sqrt{f_1} - f_1 \cdot \sqrt{f_2}} \quad (6)$$

and

$$n = \frac{\alpha_{T2} - m \cdot \sqrt{f_1}}{f_1} \quad (7)$$

Let's test this interpolation method with a real example where we know in advance the result we should expect:

Amphenol rated RG-58/A coax:

0.44 dB/100 feet loss at 1.0 MHz

1.4 dB/100 feet loss at 10.0 MHz

3.3 dB/100 feet loss at 50.0 MHz

Using the loss values for 1.0 MHz and 50.0 MHz and solving for the two constants m and n in eqns. 6 and 7 specific to RG-58/A coax between these frequencies:

$$m = 0.435604 \text{ and } n = 0.004396$$

Let's see how well we can predict the manufacturer's rating at 10 MHz using these constants in eqn. 5:

$$\alpha_T = 0.435604 \cdot \sqrt{10} + 0.004396 \cdot 10 = 1.42 \text{ dB loss/100 ft.}$$

The manufacturer gave his ratings to two-digit accuracy; we can say we came tolerably close. This interpolation method is quite accurate over the span of frequencies 1 MHz to 1 GHz when using a 10:1 range of rated frequencies. For many applications, even larger ranges are acceptable, as the 50:1 ratio of this example shows. A word of caution: the constants for m and n shown above apply only to RG-58/A. Don't try to use the same values for m and n with another cable type unless the loss ratings are identical.

phase factor β

Technically the phase factor is defined in angular units per unit length, radians per meter in the MKS system. Multiplying this value by the total line length would give us the angular electrical length. Since frequency modifies this factor, a value independent of frequency has been found more useful. Furthermore, just as with attenuation, our concern is the line total. This more practical ratio is the velocity factor, vf , usually given as a decimal fraction or a percentage.² For our purpose it obviates the need to know β , but gives us the means to derive it if required. Velocity factor is a ratio of the propagation speed of electromagnetic waves through the line as compared to its speed in a vacuum. This latter velocity is the same as the speed of light, designated c , and is equal to 299.8×10^6 meters (983.5×10^6 feet) per second. If we divide either of these values by frequency in Hz,

we obtain the free space full wavelength, respectively, in meters or feet. Multiplying this length by the velocity factor of the line, one obtains the full-wave *electrical* length. Since there are 360 degrees or 2π radians in a full wave, we can also state this length in angular units.

A source of much confusion to many Amateurs — all the more confusing when electronics professionals sometimes thoughtlessly equate them — is the difference between angular electrical length and angular current or voltage phase change. Not helping matters are the descriptive terms used — for example, “delay line,” referring to a phasing line, with “delay” implying a time-related factor. There are applications in which a propagation time delay is truly meant (radar, for instance), but usually we mean relative phase change, which is *not* time related. (If it were, we’d never be able to advance it!) On the other hand, electrical length *always* refers to time; it is determined by the time required for a particular frequency wave to travel through a medium. For any line, *phase difference* can be varied by changing the termination impedance, but propagation time is invariant, being related only to the velocity of electromagnetic waves in the medium.

As we have seen, propagation time can be defined in several ways, all referring to velocity. The most obvious, time itself, is not easily measured directly. The more commonly used measuring sticks are wavelength and angle. Frequency, velocity, and time are directly and unchangeably related for any medium; when a piece of transmission line is said to be a quarter-wavelength long, this means the propagation time is 1/4 of the time required to travel a complete wavelength at a particular frequency. Since frequency is involved and there are 360 degrees in a complete wave cycle, we can also state it as the number of degrees represented by the time interval. But never confuse this angular representation of propagation time with the current or voltage phase angle difference that may exist between the input and output of this line; they have quite different meanings. It’s true that if we match a line with its characteristic impedance, its angular electrical length will be the same as the phase difference, but keep in mind how that coincidence is brought about. Antenna elements seldom meet that requirement without help; your chances of a coincidence is little better than your chance of being hit by a meteor.

Phase change is the angular difference in the voltage or current cycle measured between two points in a circuit under steady-state conditions. Furthermore, in most instances we are concerned with *relative*, not absolute phase change. Note the word “circuit”; phase change may occur in any circuit and can be made positive as well as negative in certain circuits.

An apt analogy for explaining the difference between angular phase change and angular propagation time is the count of the number of cars of a train between two markers along a railway. At the steady-state condition (a continuous line of cars between markers) only car length, not the speed of the train, would influence the count. Conversely, the time required for a particular car to pass by both markers — the propagation time — owes nothing to the count of cars between markers, only to velocity.

If propagation time has nothing to do with phase difference, then why so carefully measure the length of these “delay” lines? This stems from a need for a specific current or voltage phase difference. In my railway analogy, suppose a particular car count is wanted between markers. If the length per unit (degree, radian, or railroad car) can be controlled, then the total count can be predetermined. If 90 degrees is the phase difference desired and the termination impedance is made a pure *resistance*, then a quarter-wave line — that is, a 90-degree line, will produce the required phase change. The impedance qualification is absolutely essential, because the nature of the termination predetermines, or controls, the phase change. If the load isn’t a pure resistance, the phase change won’t be 90 degrees and all the care going into measuring the length of this line will be as useful as rearranging the deck chairs on the Titanic. For emphasis, let’s put this another way: it’s entirely possible, with certain termination impedances, for a 70-degree electrical length line to have a 90-degree phase difference in current.

solving complex hyperbolic functions

There are two approaches to solving complex hyperbolic functions: expansion of the complex hyperbolic function into functions of real circular and real hyperbolic angles, and direct evaluation of the complex exponents of ϵ . The first method is useful where real hyperbolic functions are available in calculators or computer programs or from textbook tables.

Expansion method.

$$\cosh(\alpha + j\beta)l = \cosh \alpha l \cdot \cos \beta l + j \sinh \alpha l \cdot \sin \beta l \quad (8)$$

$$\sinh(\alpha + j\beta)l = \sinh \alpha l \cdot \cos \beta l + j \cosh \alpha l \cdot \sin \beta l \quad (9)$$

These relationships expand the complex hyperbolic functions into functions of real circular and real hyperbolic angles. For every term in the expansion, including the j terms, the real function values may now be substituted. Compared to direct evaluation, which follows, this method offers simplicity of calculation, provided real hyperbolic function values are available. (You C language aficionados finally may have found a use for your hyperbolic functions, which are a standard component of most C compiler math libraries.)

From these expansions, we can get the terms for the lossless case for coax calculations. If α is zero, then the real hyperbolic sines equal zero and the hyperbolic cosines equal one resulting in:

$$\cosh(\alpha + j\beta)l = \cos \beta l \quad (10)$$

$$\sinh(\alpha + j\beta)l = j \sin \beta l \quad (11)$$

This means that the lossless case need not be provided as a separate calculation in a program. If the loss is entered as zero, a program using hyperbolic functions will carry out the calculations correctly. The angular units for these calculations had best be radians. Besides, published tables and computer functions for real hyperbolic always use radians for angular units. Loss must be in nepers.

Direct evaluation. For most of us the method using complex exponents of e (the Napierian base 2.71828...) offers the better solution because it requires access only to real circular functions:

$$\cosh(\alpha + j\beta)l = \frac{e^{(\alpha + j\beta)l} + e^{-(\alpha + j\beta)l}}{2} \quad (12)$$

and

$$\sinh(\alpha + j\beta)l = \frac{e^{(\alpha + j\beta)l} - e^{-(\alpha + j\beta)l}}{2} \quad (13)$$

It can be shown that:

$$e^{\alpha l} \cdot e^{j\beta l} = e^{\alpha l} \cdot (\cos \beta l + j \sin \beta l) \quad (14)$$

and

$$\frac{1}{(e^{\alpha l} \cdot e^{j\beta l})} = \frac{1}{[e^{\alpha l} \cdot (\cos \beta l + j \sin \beta l)]} \quad (15)$$

Note that eqns. 14 and 15 are the rectangular equivalent of the polar vector representation:

$$e^{\alpha l} \angle \beta l$$

and

$$e^{-\alpha l} \angle -\beta l$$

Either eqns. 14 and 15 or eqns. 16 and 17 may be used. Since many scientific calculators have polar/rectangular conversions, eqns. 16 and 17 would be a good choice. The problem is now reduced to one that can be handled by calculators or computers with real circular functions (phase may be in degrees if this suits the purpose). Loss must still be in nepers.

At line 140 of the complex hyperbolic functions, note the statement $S = 10^{(S/20)}$. This is the decibel-to-neper conversion. It avoids the use of the irrational number e . Instead, it is stated as a fractional exponent of 10. Because this equality may not be immediately obvious, here is its explanation:

As stated previously:

$$\text{dB} = 0.1151 \text{ neper} = \log_e \sqrt{P1/P2}$$

$$\text{or } e^{0.1151 \text{ dB}} = \sqrt{P1/P2}$$

$$\text{and dB} = 10 \log_{10} P1/P2$$

$$\text{or } 10^{\frac{\text{dB}}{10}} = \sqrt{P1/P2}$$

Taking the square root of both sides of this equation:

$$10^{\frac{\text{dB}}{20}} = \sqrt{P1/P2}$$

therefore

$$10^{\frac{\text{dB}}{20}} = e^{0.1151 \text{ dB}}$$

Put it down to the fact I don't like having to remember and type irrational numbers or odd fractional conversion factors. (Another of my favorites is $PI = 4 \cdot ATN(1)$. Let the computer remember PI , and to as many places as it can muster!)

example explains

I've always found that when learning a new calculation procedure, nothing breeds confidence like being able to duplicate the results of a step-by-step example calculation. I've therefore chosen an example which highlights some of the surprising things that can happen with real coax and a high mismatch.

The frequency is 14 MHz, using RG-58/A 50-ohm coax. The load is $10 - j10$. The electrical length is 2 wavelengths (4π radians or 720 degrees); since the velocity factor is 0.66, the physical length is 92.74 feet. The load current will be assumed to be $1 + j0$ amperes. Using the interpolation equations and the rated loss at 10 and 50 MHz, the loss at 14 MHz is 1.67 dB/100 ft. For our line length, the loss is 1.55 dB or 0.178 neper.

$$\text{COSH}(0.178 + j4 \cdot \pi) = 1.0159 + j0$$

$$\text{SINH}(0.178 + j4 \cdot \pi) = 0.1794 + j0$$

The ABCD matrix parameters:

$$A = 1.0159 + j0$$

$$B = 8.9612 + j0$$

$$C = 0.0035 + j0$$

$$D = 1.0159 + j0$$

Network calculations:

$$AZ_L + B = 19.1206 - j 10.1593$$

$$CZ_L + D = 1.0517 - j 0.0358$$

$$Z_{IN} = (AZ_L + B)/(CZ_L + D) = 18.490 - j 9.029$$

$$E_{IN} = I_L \cdot (AZ_L + B) = 21.652 \angle -27.983 \text{ (degrees)}$$

$$I_{IN} = I_L \cdot (CZ_L + D) = 1.052 \angle -1.952 \text{ (degrees)}$$

VSWR at load: 5.2:1

VSWR at input: 2.8:1

Total power into coax: 20.47 watts

Power dissipated in load: 10.00 watts

Total loss: -3.11 dB 10.47 watts

Normal loss: -1.55 dB 4.28 watts

SWR loss: -1.56 dB 6.19 watts

Efficiency: 48.8 percent

The first thing to notice is how this relatively small loss of 1.55 dB has given the VSWR as measured at the input of the line an optimistic look. The high

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Simple BASIC program calculates complex hyperbolic functions

Using the polar form for calculating the complex hyperbolic functions requires surprisingly few BASIC statements. It may be instructive to list these, which can be set up as subroutines to be called by the subroutines which are calculating the ABCD matrix values.

```
100 'Sinh S=dB total loss, T=total radians; s+jt returned.
110 FL = 1: GOTO 140
120 'Cosh S=dB total loss, T=total radians; s+jt returned.
130 FL = 0
140 S = 10^(S/20): GOSUB 170: S3 = S: T3 = T: GOSUB 190
150 IF FL = 1 THEN S = (S3 - S) / 2: T = (T3 - T) / 2: RETURN
160 S = (S3 + S) / 2: T = (T3 + T) / 2: RETURN
170 'Polar to rectangular; S=Mag., T=Angle (rad.); s+jt returned.
180 S2 = S: S = S2 * COS (T): T = S2 * SIN (T): RETURN
190 'Invert 1/(s+jt); s+jt returned.
200 U = 1: V = 0
210 'Divide (u+jt)/(s+jt); s+jt returned
220 Y = (S * S) + (T * T): S = S / Y: T = T / Y
230 'Multiply (u+jt)*(s+jt); s+jt returned
240 W0 = (S * U) - (T * V): T = (S * V) + (T * U): S = W0: RETURN
```

Comments prior to each subroutine indicate the operation performed, parameters to be passed, and parameters returned. All calculations are complex. Some of these routines are general purpose functions: polar-to-rectangular conversion, inversion, division, and multiplication.

Adding square root and rectangular-to-polar subroutines would complete a minimum set for handling the majority of complex algebra calculations. Though not necessary to the calculations in this article, they follow here:

```
300 'Square root. S=Real, T=j Term; s+jt returned.
310 'If S returned as 0, issue warning. J term may be minus or plus.
320 W0 = SQR ((S + SQR (S * S + T * T)) / 2)
330 IF W0=0 THEN T = -SQR (ABS (S)): S=0: PRINT "Sign OK j term?": RETURN
340 T = T / (2 * W0): S = W0 : RETURN
350 '
360 'Rect. to Polar. S=Real, T=j Term. S=Magnitude, T=Radians returned
370 PI = 4 * ATN (1)
380 Y = SQR (S * S + T * T): IF S = 0 THEN 450
390 W0 = ATN (T / S)
400 IF S > 0 THEN 440 ELSE IF T > 0 THEN 430 ELSE IF T = 0 THEN 420
410 T = W0 - PI: GOTO 480
420 T = PI: GOTO 480
430 T = W0 + PI: GOTO 480
440 T = W0: GOTO 480
450 IF T < 0 THEN 470 ELSE IF T = 0 THEN 480
460 T = PI / 2: GOTO 480
470 T = -PI / 2
480 S = Y: RETURN
```

To preserve accuracy, keep calculations in rectangular form as far as practical; avoid multiple polar/rectangular conversions except as necessary at input or output of data. By convention the variables S and U are real, while T and V are the j terms. (In polar/rectangular conversions, S is magnitude and T is the angle.) The returned variables are always S and T .

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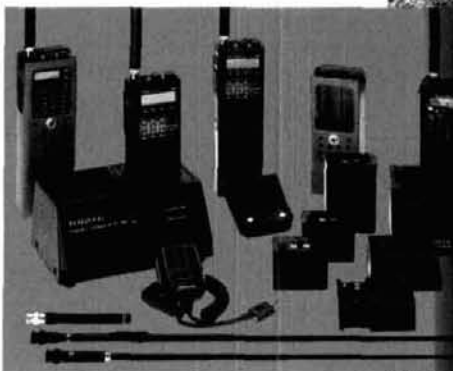
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- CTCSS encoder built-in. TSU-4 CTCSS decoder optional.
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- PB-2: 8.4 V, 500 mAh NiCd pack (2.5 W output)
- PB-3: 7.2 V, 800 mAh NiCd pack (1.5 W output)
- PB-4: 7.2 V, 1600 mAh NiCd pack (1.5 W output)
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- BC-7 rapid charger for PB-1, 2, 3, or 4
- BC-8 Compact battery charger
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- SC-12, 13 soft cases
- RA-3, 5 telescoping antennas
- RA-BB StubbyDuk antenna
- TSU-4 CTCSS decode unit
- VB-2530: 2m, 25 W amplifier
- LH-4, 5 leather cases
- MB-4 mobile bracket
- BH-5 swivel mount
- PG-2V DC cable
- PG-3C cigarette lighter cord with filter



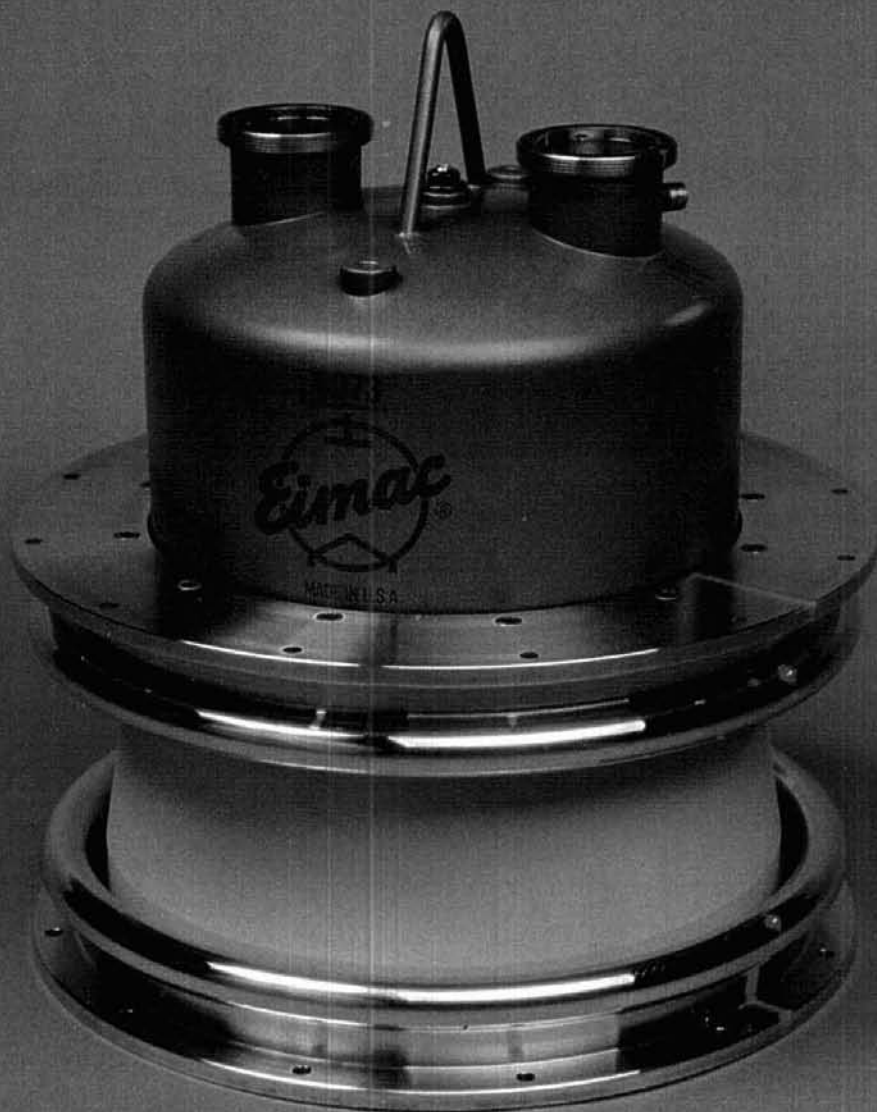
TH-215A

TH-215A shown

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For a data sheet and technical literature on the 8973, contact Varian EIMAC, 301 Industrial Way, San Carlos, CA 94070 or call (415) 592-1221, TWX 910-376-4893.

User	Application	Frequency (MHz)	Power Output	Pulse Length
JET	ICH*	25-50	1.5 MW	20 seconds
JT-60	ICH*	110-130	750 kW	10 seconds
JFT-2M	ICH*	10-40	1.5 MW	300 milliseconds
KFA-Textor	ICH*	29-53	1.5 MW	3 seconds

*ICH = Ion cyclotron heating



VSWR at the load has caused an additional loss of 1.56 dB. And even though this line is a multiple of a half-wave, where we'd normally expect to see the load impedance repeated at its input, it is in fact quite different.

I must emphasize that these effects result from line loss. If we could magically produce a lossless line, none of these effects would be seen, no matter how high the VSWR. For all that, had I chosen a much lower loss line, these effects would be drastically reduced. Reflected wave power, which would not be lost in a lossless line, gets a piece taken out during each traverse of the line — just as with forward wave power. This is the reason for the additional loss due to the high VSWR.

summary

Real coax, except in the most benign circumstances (good load match, very low loss), gives rise to effects which can be startlingly different from assumptions based on a lossless line. Nor is it entirely a problem of efficiency and optimistic VSWR. A case in point is the 4-Square vertical phased array, whose reference element requires a small part of the total drive power (on the order of 5 percent). Because of the high VSWR associated with the feedline of this element, the assumption of zero losses (even at 80 meters) results in a non-trivial error in the calculated input impedance of this feeder, leading to a drive current phase error. Since good f/b ratio is a vector subtraction of large numbers looking for a zero sum, a small error can lead to a large f/b ratio change.

ABCD matrix algebra again demonstrates its versatility and flexibility for four-terminal network calculations. Except for adding complex hyperbolic functions to my computer program, and coax loss interpolation, no other changes were necessary. Given the parameters, it doesn't matter to the ABCD program what sort of network it is calculating.

acknowledgments

My thanks again to Mason Logan, K4MT, and Bob Booth, WB6SXV, for their encouragement and assistance. They pointed the way; I wrote about it. A special thanks to Bob, who provided the interpolation method.

references

1. Forrest Gehrke, K2BT, "Vertical Phased Arrays, Part 5," *ham radio*, December, 1983.
2. Some cable ratings give the dielectric constant of the insulating material — e.g., polyethylene has a dielectric constant of 2.26, which equates to a velocity factor, $vf = 1/\sqrt{2.26} = 0.67$.

bibliography

Radio Engineers' Handbook, F.E. Terman, McGraw-Hill, pages 19 and 178.
Transmission Lines, Robert Chapman, Schaum Outline Series in Engineering, McGraw-Hill.

ham radio

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XF-9B-01	LSB	2.4 kHz	8	95.90
XF-9B-02	USB	2.4 kHz	8	95.90
XF-9B-10	SSB	2.4 kHz	10	125.65
XF-9C	AM	3.75 kHz	8	77.40
XF-9D	AM	5.0 kHz	8	77.40
XF-9E	FM	12.0 kHz	8	77.40
XF-9M	CW	500 Hz	4	54.10
XF-9NB	CW	500 Hz	8	95.90
XF-9P	CW	250 Hz	8	151.20
XF-910	IF noise	15 kHz	2	17.15

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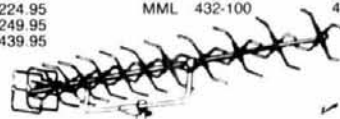
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simple modifications and adjustments for the TS-930S

Minor changes
let you align without
test equipment, improve
performance and reliability

The Kenwood TS-930S offers the serious Radio Amateur the opportunity to enjoy the high quality of an all solid-state, general coverage transceiver with a very effective built-in automatic antenna tuner, among a number of options. Furthermore, Kenwood's updated version, the TS-940S, provides all the same features plus improved cooling, increased memories, and keyboard data entry. For those using the TS-930S — and to some extent, those using the TS-940S — the following notes may be of interest.

memory backup

The TS-930S memory backup requires the use of three 1.5-volt AA batteries, which will easily last their shelf life. AA NiCads, however, may be substituted with good results; because the drain is only in the microampere region, recharging will be required only about once a year. This can be done by simply attaching some small clip-type probes to the polarity coded red and black leads that exit the battery housing on its left side (as viewed from the top, front of the radio).

i-f alignment

The rather laborious i-f alignment procedure outlined in the factory service manual may be greatly simplified by using the receiver's 100 kHz marker as a signal generator in conjunction with the S-meter. For convenience, and as a reference, the receiver's internal "S9" meter adjustment is used to set the marker

level to S9. This provides relatively accurate signal readings, consistent with those I would give by ear. Unfortunately, the incoming signals above that level will read high, but in reality those are only relative at best, and I prefer the convenience of accuracy below that point, where it's more useful. Some operators may prefer a setting of S8 as a compromise, depending on their particular needs.

The service manual's instructions for the alignment of L132 and L134 in the rf processor i-fs are quite complex. To simplify this procedure, you can use your voice as a signal generator while peaking the slugs as indicated on the *compression* scale of the radio's multimeter. Alternatively, a complex sound such as fan noise picked up by the microphone and detected in the monitor mode will suffice by just peaking the audio signal without the meter reading.

dial calibration

To align the TS-930's 10-MHz time base for basic dial calibration, the following procedure will provide a simple, yet very effective means of relatively accurate calibration. First, tune WWV to the highest frequency in which an S9 or stronger signal can be copied. With the receiver in the *a-m* mode, enable the 100 kHz marker and adjust the rf attenuator until the optimum beat note is heard in the form of the conventional "whooshing" sound, while adjusting TC1 for zero beat. The best results will be obtained when the two signals are of equal amplitude — i.e., when the strongest beat note is heard. Always perform this calibration when WWV is well above S9 to allow for the maximum adjustment range of the rf attenuator.

preventing failure in R400

In all TS-930S units with serial numbers below 4,100,000,* resistor R400 on the signal board is prone to failure. In my radio, the precursors of failure were

*There are obviously not 4.1 million TS-930Ss — the manufacturer's numbering system simply identifies products in this manner. Ed.

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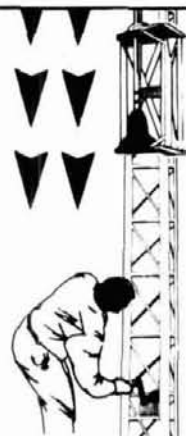
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suggested by a VOX malfunction and a 15-minute delay in the ALC meter reaching zero. Several other 930 owners have told me that they've found R400 in their units discolored and too hot to touch; this strongly suggests a high probability of failure.

The measured power dissipated in this resistor is 1/2 watt with a 1.5-k value. The original rated dissipation of this resistor was either 1/4 or 1/2 watt (the reason for this uncertainty is that my failed resistor wasn't returned to me, and the manufacturer doesn't specify resistors in their parts lists). As originally designed, it was the only hot resistor on the signal board. In all models numbered above 4,100,000, Kenwood changed its value to 2 watts, so be sure to inspect this resistor for discoloration and measure its value. The existing resistor may be replaced by crushing its body and soldering the new one to the remaining leads; if you do it this way, you'll have to remove the signal board. Of course, if you're removing the board any way, you might as well replace the resistor.

R400 is difficult to find because it's not located in the area of its sequential grouping, but instead at the top left center of the signal board, as viewed from the bottom side of the radio (facing you). If you have the service manual — a bargain at \$15, by the way — you can find R400 at the intersection of C and 1.5.

ALC considerations

Some operators have been experimenting with the ALC time constants, with the objective of increasing the 930's "talk" power. They believe that some minor improvement may be accomplished by decreasing the decay time of the ALC, since this varies by the net tolerances of the components involved. However, any alteration (delay) of the *attack* time should be avoided because doing so will only increase the intermodulation distortion products. A senior Kenwood factory engineer has advised that any ALC indication *above zero* causes an increase in distortion; therefore, ALC should be used sparingly and in accordance with the operating instruction manual. Should the decay time be found to be excessive (that is, about 5 seconds from half scale), it may be reduced by changing R240 — located at the intersection of 4 and C/D (see page 39 of the service manual) — on the signal board to a smaller value. Unfortunately, unless you remove the board, you'll have to crush it, salvaging its leads for the replacement.

rf decoupling

Like most solid-state equipment, the TS-930S exhibits some degree of rf susceptibility, even under very low VSWR conditions. Coil cords make excellent antennas. So do any other cables to external units such as speakers, phone patches, and keyers. The

liberal use of ferrites, consistent with good shielding practices, will usually take care of most problems. It's axiomatic that all external leads should be decoupled.

Evidence of external rf entering the radio and resulting in distortion may be found as follows. Wearing headphones, turn off the monitor and determine whether a garbled SSB signal is heard. Next, place the multimeter in the Vc position and note if it remains well regulated under full power conditions, especially with a power amplifier in operation. Poor Vc regulation is a sure indication of a problem, either because of rf or in the regulator circuit itself.

improving transmitted audio, greater IMD suppression

Users of radios with serial numbers below 3,080,001 may not be aware that a Kenwood factory modification was inaugurated at that point for improved transmitted audio quality — i.e. improved IMD suppression. The modification consisted of increasing the collector current of the rf drivers and final transistors, the MRF485s and the MRF422s in their quiescent state. The effect is more pronounced in the reduction of the higher, odd order products. The full details can be found in Kenwood's Factory Service Bulletin No. 867 (March 29, 1983), entitled "TS-930S SSB TX Tone Quality." The enhanced performance occurs by operating the drivers and finals deeper in Class AB1, by increasing their collector idling currents from 50 and 500 to 70 and 1100 mA, respectively. To determine if this mod has already been incorporated in your radio, place the multimeter in the Ic position and compare the value you read to the 1100 mA. To some extent, the results of this mod account for the difference in transmit quality between the earlier and later versions of the 930.

Thanks to W4CG, a 5-dB improvement in the IMD of the radio may be readily obtained by replacing Q1, the rf pre-driver (a 2SC2075) with a Motorola MRF485. It's essential that the MRF485 be of current manufacture because early devices exhibited a thermal runaway problem which has since been corrected by Motorola. The new, improved version is identified by a narrow white horizontal stripe across the face, just above the device identification number. The MRF485 is an ultralinear device, especially designed for SSB service and therefore just a bit better in this application than the 2SC2075, while being fully interchangeable in form, fit, and function. Therefore, no parameter change is required. Replacing the device is actually quite simple. It can be a bit of a chore in the 930 because of the tedium of removing the rf board from its mounting; in the 940, the job is much easier.

quieting a noisy cooling fan

In most radios, the power supply fan will become

noisy after a year or two, depending upon the particular fan unit and the number of hours of use. Kenwood has developed a simple modification that greatly reduces or eliminates it. The mod consists of adding a small spring clip (Part No. 602-0549-04) to laterally load the fan shaft. Described in Kenwood's Service Bulletin No 893, the mod takes only about 30 minutes.

I've also found that a light Teflon™-loaded lubricating oil such as Tri-Flow™ is an excellent lubricant for the front bearing of both fans. However, unfortunately, neither fan is designed for lubrication of the rear bearings.

increasing dial light life

In order to save the multimeter dial lights, which are costly and based on a special pigtail design, Kenwood recommends that they normally be operated in the "dim" position. However, a 12-volt Radio Shack unit (No. 272-1141, 79¢) can be substituted and wired in series for the 28-volt source.

ergonomic "face-lifting"

Adding some 1-inch rubber extensions to the front mounting feet will make viewing the controls easier, thereby helping the operator's posture and possibly lessening eye strain. (This will be especially true for contest operators.) These rubber extensions are obtainable at parts supply houses or hardware stores

at a nominal cost. I cemented mine to the existing feet with 5-minute epoxy.

balanced modulator stability

I've observed a frequent, random imbalance between the capacitive and resistive balance adjustments. While the radio undoubtedly meets its carrier suppression spec, it is rather disconcerting to observe the imbalance occurring. Therefore, it's a good idea to check the balance periodically, using another receiver as a detector or an oscilloscope with a dummy load across the output of the radio. While the amount of carrier may be small, it may become quite annoying when it's amplified by a linear and a beam. When using an auxiliary receiver as the detector, advance the Processor In control fully clockwise and reduce the Processor Out control counterclockwise to the fully off position. This provides the best combination for checking without audio feedback.

conclusion

All of the foregoing procedures may be accomplished without test equipment and without following the detailed factory instructions, yet they should yield the same results. I hope that these methods and observations will, in some ways, add to your operating pleasure while using these fine radios.

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ANTENNA POLARITY SWITCHER MODEL APS-1

The APS-1 is a self-contained control head designed to allow remote polarity switching of circular antennas such as the Mirage/KLM range of crossed yagis.



The APS-1 may be powered by the power adaptor (included) or may alternately be powered from a vehicle or other 13-17 VDC source.

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SPECIFICATIONS:

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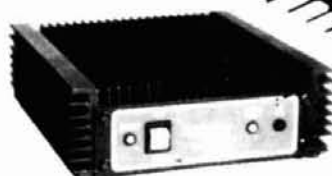
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packet radio conference bridge

AX.25-compatible bridge
links six stations
for routine or emergency
communications

From the beginning, packet radio has been essentially a point-to-point mode. Even with the TNC-2's multi-connect capabilities, it hasn't been possible for all participants in roundtable or net-type operations to be connected to everyone else in the net. Although makeshift arrangements have been devised, they've lacked the anti-collision and error-controlling capabilities of the AX.25 protocol.

In an effort to solve these problems, Tom Aschenbrenner, WB5PUC, has designed a truly error-controllable, AX.25-compatible conference bridge. Because it's combined with other network software components he's designed, the conference bridge is offered in two versions. The one described here fits into a replacement EPROM for a TNC-2 clone; the other is a part of the software installed in the 9600-baud network node controller board for TexNet.^{1,2}

A conference bridge module in a TNC-2 clone or a TexNet node provides full-protocol, multiple-station roundtable or net-type operation between packet stations. Typical operation is accomplished by each of the stations involved in the net connecting to the bridge-equipped node by using the Secondary Station Identifier (SSID) assigned to the conference bridge function. On the test nodes in operation in the Dallas area, the SSIDs are -2 and -3 on each node. The current version of the network software supports two independent conference bridges of six participants each. It's also possible to connect to the bridge through one digipeater if necessary.

typical conference bridge operation

To connect to a conference bridge, each station

connects as if it were connecting to any other packet station. A typical text sequence to a conference bridge would be:

```
C WB5PUC-2 < carriage return >
```

The operator's TNC does the connect routines.

The following will then appear on the operator's CRT:

```
***Connected to WB5PUC-2 < cr lf >  
(from TNC)
```

Welcome to the WB5PUC Conference Bridge. A Control-U shows all stations connected to this bridge.

At this point or at any other time, the response to a Control-U command to the bridge will bring up a text string listing the calls of all connected stations. For example:

```
N5EG-5 WD5HJP W5YR-7 WA5MWD-3  
connected
```

No additional commands are needed to operate the bridge.

In the normal operation mode, each operator receives a text string with a shorter header indicating the call of the originating station. For example:

```
WD5HJP > Transmitter power is now at 100  
watts.
```

```
N5EG-5 > OK Bill, try adjusting trimmer C15  
now.
```

```
W5YR-7 > What are you guys up to?
```

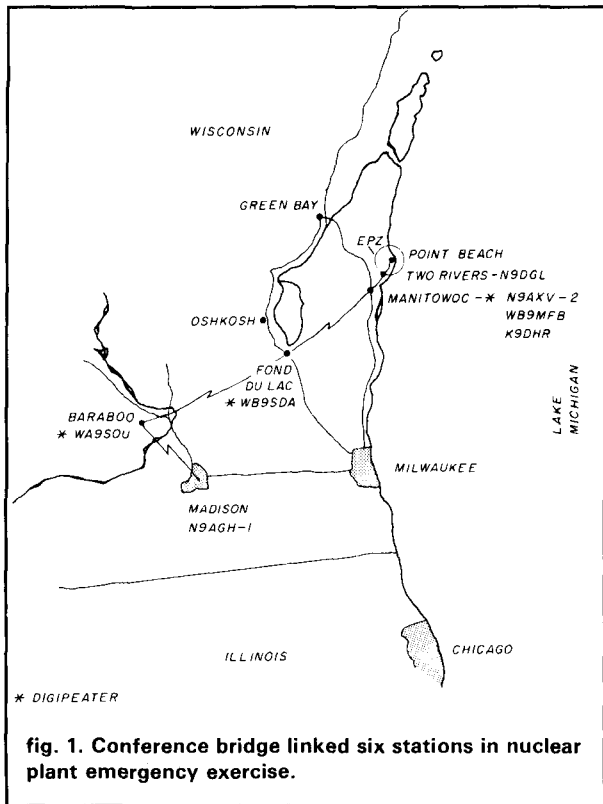
```
WD5HJP > Hi George, just adjusting the  
node's final amp.
```

When the QSO is over, those connected to the bridge simply disconnect as they would from any other packet connection, via a DISCONNECT command in the Command mode of the TNC.

operations test

A routine test of the emergency plan for the Point Beach, Wisconsin nuclear power plant in September, 1986, provided an intensive on-the-air test of an early version of the conference bridge software. Though not specifically designed for such use, the conference

Bill Wade, WD5HJP, Texas Packet Radio Society, 600 Via Sevilla, Mesquite, Texas 75150



bridge served as the hub of an emergency communications network that included a link to the state capital. Overall, the bridge performed well in its original form; later modifications have been implemented to facilitate its use in emergencies.

The Nuclear Regulatory Commission (NRC) requires annual testing of every nuclear power plant's emergency plan. This test is designed to evaluate the ability of plant personnel and the utility company holding the facility's license to cope with an accident. The NRC measures their ability to assess the extent of danger to the public, their effectiveness in recovering control of the plant, and their ability to minimize damage to the surrounding environment. A succession of events pushes the plant engineering staff through a series of critical decisions; events are programmed into the scenario to simulate damage to the power plant and motivate recovery action by the staff.

For a realistic overview of the performance of the allied agencies that would be involved in an actual incident, the test scenario includes a simulated evacuation. State, county, and municipal emergency units become involved in the plan when supervisory NRC engineers and the utility's power plant engineers have recognized a possible threat to public safety. At some point in the escalation of the situation, the plant staff recommends evacuation, triggering activation of a number of government safety, information, and public assistance centers.

At that point, the power plant staff contacts the county and state emergency units and delivers its assessment of the situation. The complement of offices that become active in the evacuation phase of the emergency plan are the county Emergency Operations Center (EOC); the state Emergency Government Office (EGO); the Department of Human Resources Reception Centers that will process evacuees; the Joint Information Press Center (JIPC), the official information center for the emergency; and the off-site power plant Emergency Operations Facility (EOF), from which the NRC and utility supervisory engineers make their recommendations to the state and county emergency government.

Each of the 102 nuclear power plants in operation in the United States is surrounded by an Emergency Protection Zone (EPZ) with a radius of 10 miles. This EPZ is the area considered under immediate hazard in case of any airborne release of radioactivity. The Point Beach power plant EPZ includes three towns — Two Rivers, Shoto, and Two Creeks — and extends into Lake Michigan.

test scenario

This particular scenario began with a hypothetical earthquake tremor, causing a series of leaks in the cooling loops of the "B" power plant. Subsequent "damage" to the plant caused some injuries and contamination of plant employees and allowed the release of radioactive steam into the atmosphere. By 8 AM, after a damage assessment, the county and state government were notified that there was a "public danger"; the county emergency office notified the Manitowoc RACES EC, who activated the amateur emergency system. Stations equipped with 2-meter fm voice and packet equipment were established at each of the operations centers listed above, with the exception of the plant Emergency Operations Facility (EOF), not normally accessible to the public.

RACES participation

The Amateur community in Manitowoc County is actively involved with the county emergency government office! During the test, the five-station RACES system was centered around the Manitowoc County EOC, which has a permanent Amateur Radio station linked via microwave to a dedicated voice repeater (WB9MFB, 145.19/.79) located at the county transmitter facility. Four remotely-sited, linked voting receivers increase the effectiveness of this 60-mile diameter service area, 2-meter fm voice system. For packet operations, a permanent conference bridge-equipped packet digipeater (N9AXV-2, 145.01 MHz) utilizing an MFJ-1270 TNC-2 is installed at the transmitter site. The voice repeater and packet digipeater are converted 100-watt Motorola transceivers.

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Other equipment used at all of the temporary public service Amateur stations included additional 2-meter fm voice transceivers for the RACES 145.19/.79 repeater. Voice fm was used to coordinate large file transfers and to back up the packet equipment. A variety of computers (chiefly IBMs and IBM clones and Commodore 64s) and TNCs (AEA, Kantronics, MFJ, and original TAPR 1's and 2's) were used without any compatibility problems. Each of the locations was equipped with the appropriate disk drives; some were also equipped with printers.

Operators at each of the centers connected to N9AXV-2, the conference bridge/digipeater at the county transmitter site. N9AGH-1 was at the Wisconsin State Emergency Government Office in Madison, the state capital. To reach the conference bridge, N9AGH-1 connected through two digipeaters, WA9SOU (in Baraboo) and WB9SDA (in Fond Du Lac), a link length of over 150 miles (see fig. 2). WB9MFB, at the county EOC, K9DHR, at the reception center in Manitowoc, and N9DGL, at the JIPC in Two Rivers, were all connected directly to the bridge.

Once the stations were connected, typical operation proceeded as in any other net. Net control was maintained by operators at WB9MFB, the county EOC. Traffic was passed simply by sending the text via packet from each of the sites. Since all stations on the bridge were getting identical copies of text, very little repetition was necessary.

impact of packet operation

The worth of the conference bridge and packet radio was demonstrated immediately. The test emergency was declared from the governor's office in Madison, and the message went out via the two-digipeater link to the conference bridge at the Manitowoc EOC. A parallel 75-meter phone system running from the state emergency office to the county EOC was also activated. The packet conference bridge delivered the message correctly, approximately 30 seconds ahead of the 75-meter phone link. The 75-meter system garbled the power plant emergency protection zone (EPZ) grid coordinates in the first message!

In some respects, use of the packet conference bridge became second nature. To simulate a system failure, Ron Shimek, WB9MFB, the county EC, disconnected power to the conference bridge, forcing the use of point-to-point packet communication. When they became aware of the system failure, packet operators set up point-to-point links to re-establish communication. Although throughput was slower, point-to-point operation did provide usable information. As a backup, a standby station had been set up to monitor activity and act as recorder for the entire test.



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test results

At the conclusion of the exercise, the Emergency Coordinator described the conference bridge as a key ingredient in the success of the test. In his report, he emphasized that federal observers had been impressed not only with the speed and accuracy of traffic handling, but also with the capability for time/date stamping of traffic (using the computer systems' real-time clocks) and for producing hard copy simultaneously at all sites.

Because all the stations in the emergency system were connected through the conference bridge, automatic status and warning updates were available at all sites. An NRC inspector mentioned to one of the participating Radio Amateurs that thanks to ham activity, he was never out of touch with any of the sites for the duration of the test. The Federal Emergency Management Authority (FEMA) observers indicated that they would encourage further use of packet radio and the conference bridge in future tests. WB9MFB's evaluation of the test suggested a broader role for the RACES group in both forthcoming tests and actual emergencies.

recommendations for future designs

The relatively few operation anomalies that caused some delays on the system were largely attributable to QRM originating outside the limits of the test area. The Manitowoc county transmitter site is about 3 miles from Lake Michigan; there is considerable channel activity from Milwaukee and Chicago to the south and from Michigan across the lake. All of the test activity took place on 145.01 MHz. Typically, there are periods of 10 to 20 minutes when the N9AXV digipeater squelch never closes. Selection of another frequency besides 145.01 MHz for operation would be desirable.

Connection of the conference bridge to a network system is also recommended. In this test, the state emergency government center was by necessity connected through a 145.01-MHz, two-digipeater link. Response time from that part of the system was proportionately slower, but still quite usable. Other situations without such a robust digipeater link would benefit from use of a backbone network system. The conference bridge software does allow connection to any network because of its compatibility with AX.25.

WB9MFB has suggested that a system monitor be set up in advance to record all activity during an emergency communications test. Packet operation offers a significant advantage in this regard, in that all text information is easily stored to disk. In this case, the system monitor provided a good backup to the NCS; later, when the EC and RACES group needed to do evaluation of their own, the stored information proved to be useful indeed.

Other recommendations include eliminating BBS activity on frequency during tests or actual emergencies to help avoid channel congestion. Operators from outside the test area occasionally interrupted RACES activity with inquiries about the test and the conference bridge.

implications of conference bridge operation

Amateurs who've handled traffic in hurricanes, tornados, or other disasters know that telephone systems are the first communications systems to become overloaded or destroyed. The Point Beach scenario demonstrated that utility companies and state agencies are still blindly tied to this relatively fragile communications resource. The ability of packet radio systems to handle high volumes of traffic quickly and accurately under difficult circumstances is being demonstrated regularly; as part of these systems, a conference bridge can provide reliable communications to the people that need it most.

references

1. Thomas H. Aschenbrenner, WB5PUC, and Thomas C. McDermott, N5SEG, "The TEXNET Packet-Switching Network — Part 1: System Definition and Design," *ham radio*, March 1987, page 29.

2. Thomas H. Aschenbrenner, WB5PUC, and Thomas C. McDermott, N5SEG, "The TEXNET Packet-Switching Network — Part 2: Hardware Design," *ham radio*, April, 1987 page 29.

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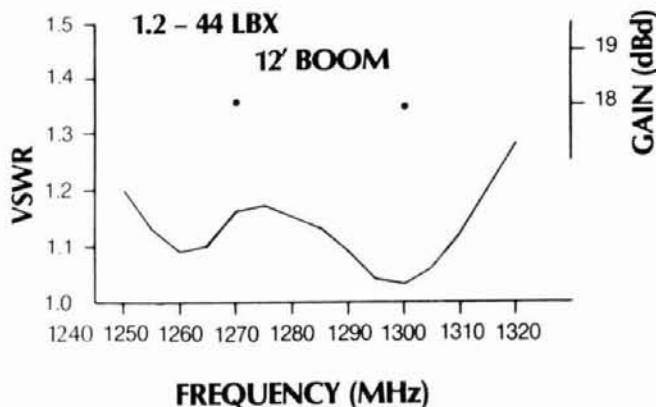
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the TEXNET packet-switching network

part 2: hardware design

In last month's article,¹ we discussed network algorithms and the software layering. This month, we'll focus on the design and testing of hardware for the network, and on the results that have been achieved to date.

system partitioning

Partitioning hardware to minimize the number of signals that must connect between units offers three benefits: simplified cabling, easy measuring and modification of individual units, and flexibility in the construction of the network.

Figure 1 is a block diagram of a TEXNET network node. There are four main pieces: a local area network (LAN) radio, which in this case is a 1200-baud AFSK modem and 2-meter radio; an inter-processor (IP) radio (a 9600-baud FSK radio and modem) for use as the high-speed network trunk; a node control processor (NCP) that contains the microprocessor and communications ICs; and the power supply, which contains a three-state float charger, battery, and circuitry for automatic uninterrupted power should ac power fail.

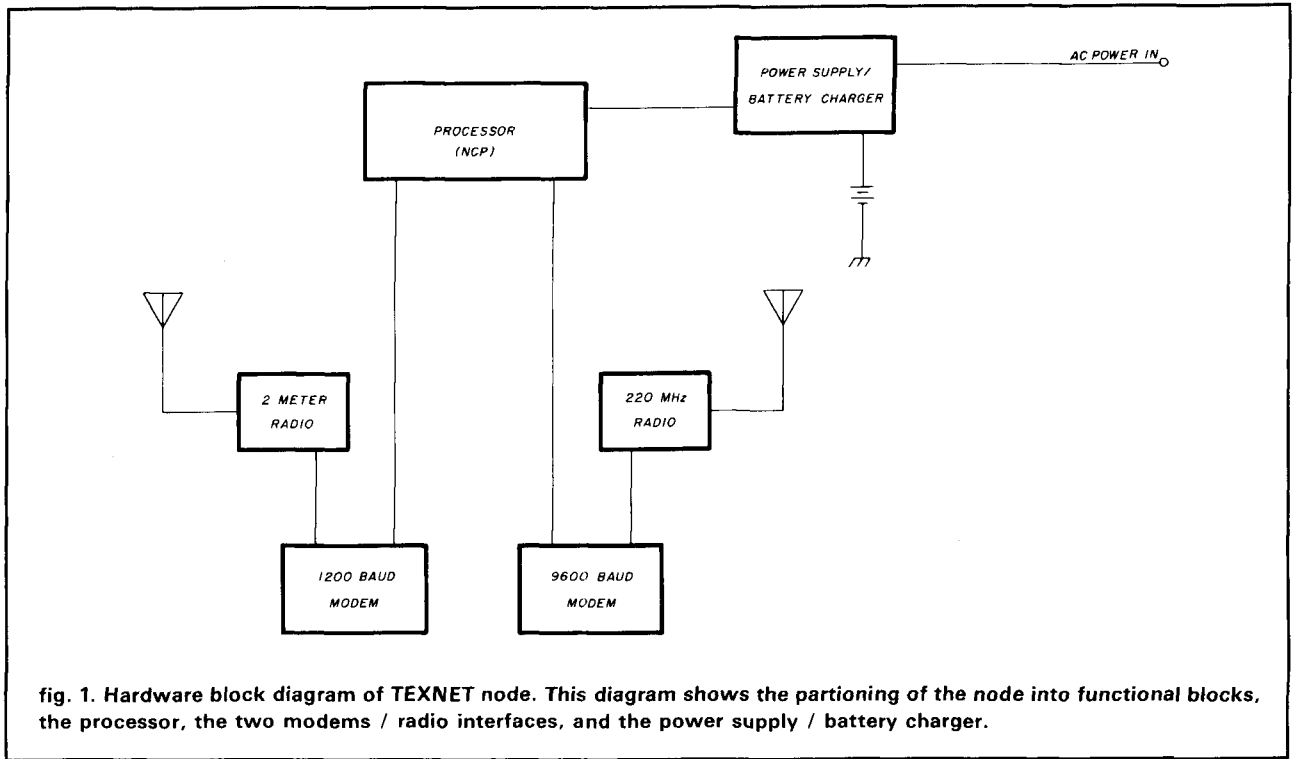
LAN radio and modem

This channel is the primary method by which users with TNCs and 2-meter radios connect to a network node. By connecting the modem separately from the processor, any modem can be used — 300 baud, 1200

baud, 2400 baud, or whatever might be desired. **Figure 2** is a diagram of the modem, which is similar to the TAPR TNC-1 modem. We chose to implement an active filter equalizer with op-amps instead of a switched-capacitor filter IC. The modem includes a 45-second time-out timer to disable the transmitter should the controller fail for some reason. The strap allows setting the EXAR demodulator VCO center frequency, but better results can be obtained by adjusting the VCO frequency control pot and observing the received "eye" pattern on an oscilloscope from an AFSK signal known to be good. An eye pattern is observed on an oscilloscope by synchronizing the scope trigger with the recovered clock and displaying the data. Since the data displayed is not involved in triggering the scope, a random display of all data sequences is shown, but the zero crossings are fixed in time on the screen, yielding an open area in the center of a data bit (known as the "eye"). **Figure 3** shows a typical eye pattern, the recovered clock, and the slicing level (which decides between a 1 and a 0). The basic decision circuit is shown in **fig. 4**.

The radio is an ICOM-IC22S, a popular 2-meter transceiver, with the frequency hard wired to 145.05

Thomas H. Aschenbrenner, WB5PUC, and Thomas C. McDermott, N5EG, Texas Packet Radio Society, P.O. Box 831566, Richardson, Texas 75083-1566



MHz. The transmit audio is injected after the microphone amplifier; the receive audio is tapped off prior to the audio PA in order to avoid the severe degradation of frequency response that results if the speaker and microphone leads are used for audio pick-off and injection.

IP radio and modem

The performance of the network trunks is very important in determining the overall throughput of the network as a whole. As each user sends traffic into a network node, all traffic is multiplexed (combined) onto the high-speed trunks. Thus the trunks carry a much greater amount of traffic than the user links. Because of this, we have decided to operate the trunks at 9600 baud, with rapid Transmit/Receive (T/R) switching. Rapid T/R switching is required because at 9600 baud, actual packets take relatively little time to transmit, and the T/R delay can determine the effective channel capacity.

Figure 5 illustrates the effective channel capacity versus T/R delay for a 9600-baud channel with no errors, and one acknowledgment packet for the entire transmission. Several different values are shown: one indicates the number of packets per transmission, another the number of bytes per packet, and a third for two values of DWAIT (digipeater waiting time, which allows a digipeater transmission priority). Because there are no digipeaters in TEXNET, DWAIT = 0. (A value of DWAIT = 80 ms is typical for 2-meter

channels.) Our experiments involved the use of a pair of Hamtronics FM-5 220-MHz fm transceivers and K9NG's modems. These radios are modified to operate FSK, and the received data signal is tapped off the quadrature detector in the receiver. Since these radios are PIN-diode switched between transmit and receive, we were able to make them operate with 40 ms T/R delay, although in practice 80 ms was allowed.

We encountered some difficulty in making the radios operate properly at 9600 baud. Apparently these problems were due in large part to variations in the performance of different FM-5 radios, which were designed not for high-speed data operation, but rather for fm voice operation. In an effort to improve the operation of the radios, a number of experiments were run, and modifications were made to the modem.

To understand this better, let's review the basics of frequency-shift-keyed (FSK) data transmission, the spectrum of a non-return-to-zero (NRZ) data signal at baseband, and the performance of i-f filters in the time domain. We'll see that all three have to be addressed properly to assure proper performance of the radios and modems.

An NRZ signal is one that toggles between logic 1 and logic 0 no more than once per bit period (see fig. 6). In an FSK system, two frequencies are transmitted — one for logic 1, the other for logic 0. At the output of the discriminator/quadrature detector at the receiver, the two frequencies are translated back to voltages. If the frequency of the transmitter were to

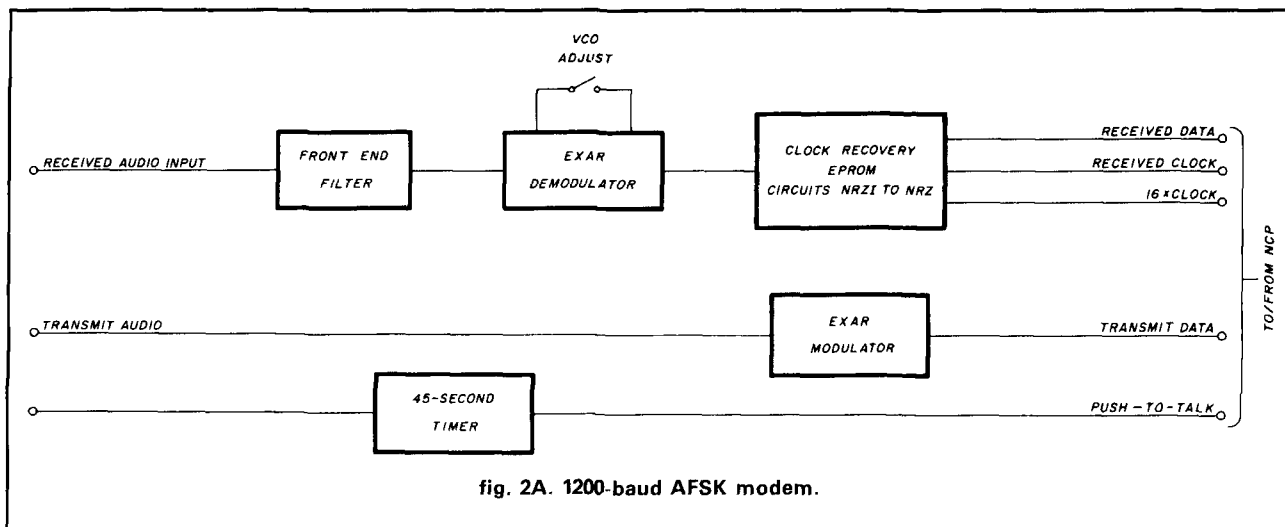


fig. 2A. 1200-baud AFSK modem.

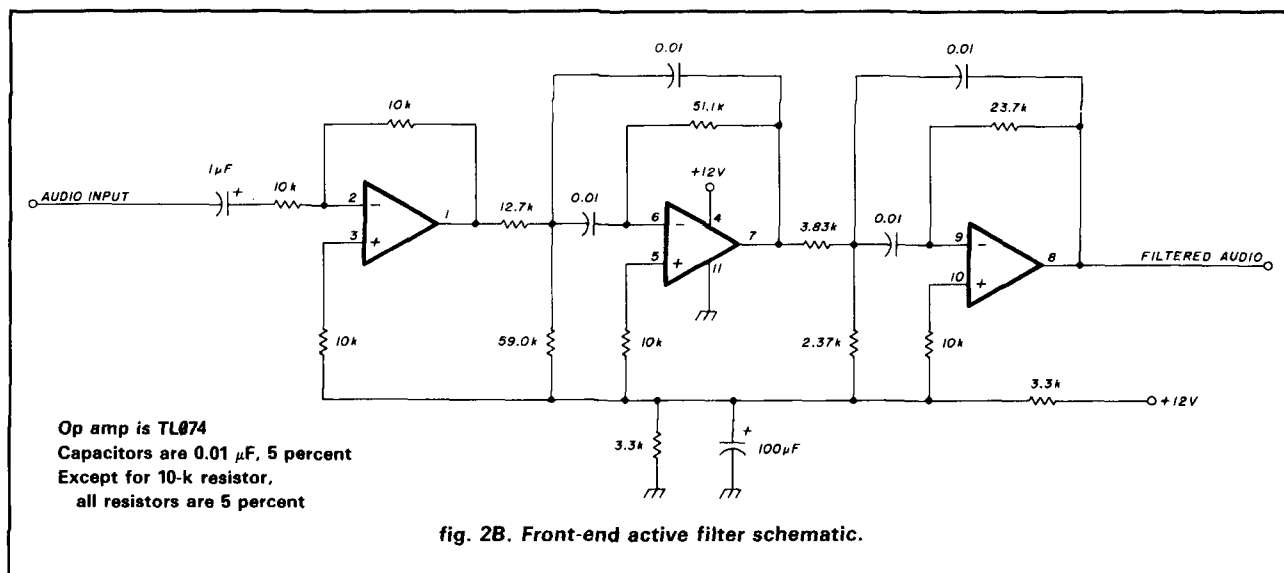


fig. 2B. Front-end active filter schematic.

vary slightly, then both the logic 0 and logic 1 voltages would also vary. A simple method to determine the correct "slicing level" for deciding between a logic 1 and a logic 0 is to choose the voltage halfway between the two. This is simple to do by using a low-pass filter with a time constant quite a bit longer than the data period to generate the slicing voltage level. Then the slicing threshold will "track" the logic 1 and 0 voltages automatically (see fig. 7). This requires, however, that the transmitted data have, on the average, the same number of ones as zeros; if they don't, the recovered slicing level will be biased off the true center point (fig. 7).

In HDLC, the code used for AX.25, there is no guarantee that the code will be dc-balanced (i.e., have same number of ones as zeros). In fact, the flag character contains two zeros and six ones, thus having a large dc offset from one-half. A simple way to solve

the problem is to use a pseudo-random scrambler to cause some apparent randomization of the data sequences. A self-synchronized descrambler is used on the receiving modem to recover the original bit stream. This is the method used on the K9NG modem to send and receive data, a 17-stage scrambler being used. With this arrangement, the average number of ones and zeros is nearly the same. Certain sequences into a scrambler can, however, produce long strings of ones or zeros. If a long string were to occur, our low-pass filter in the receiver would tend to drift off the center voltage, halfway between the 0 and 1. So we must compromise the time constant of the low-pass filter in the receiver slicing level circuit (which we would like to make very long) with the need for rapid T/R switching, where we need to acquire the proper level quickly. In addition, any capacitors used to couple the analog signal must have long time constants; if

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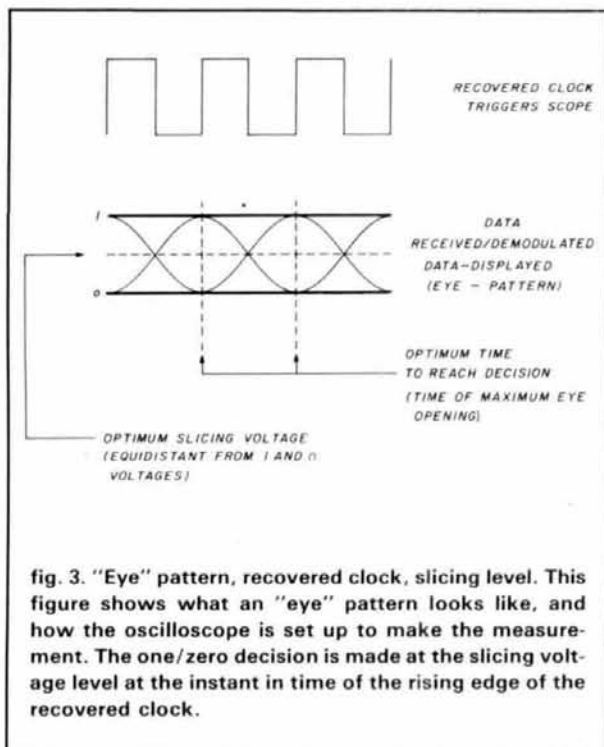


fig. 3. "Eye" pattern, recovered clock, slicing level. This figure shows what an "eye" pattern looks like, and how the oscilloscope is set up to make the measurement. The one/zero decision is made at the slicing voltage level at the instant in time of the rising edge of the recovered clock.

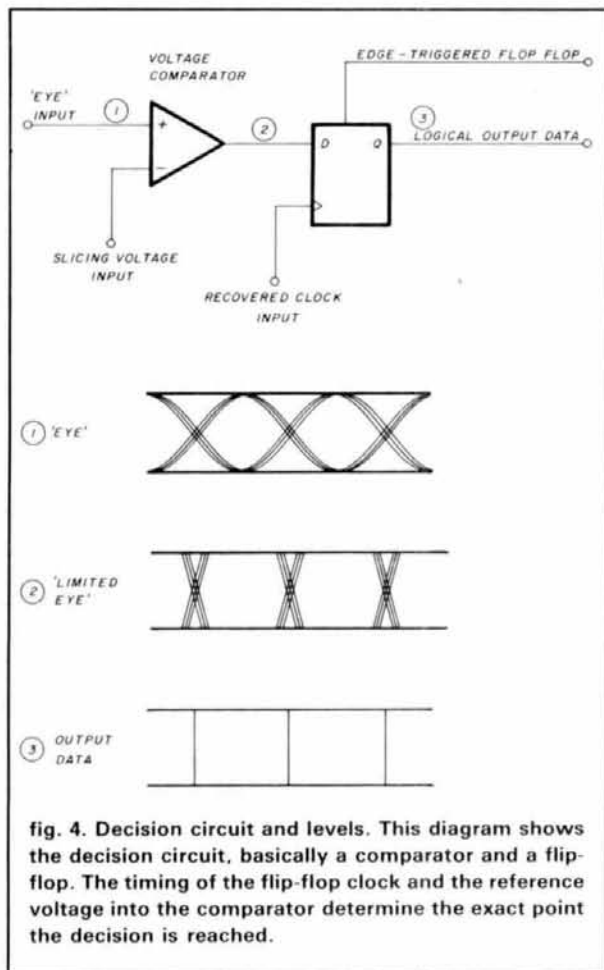


fig. 4. Decision circuit and levels. This diagram shows the decision circuit, basically a comparator and a flip-flop. The timing of the flip-flop clock and the reference voltage into the comparator determine the exact point the decision is reached.

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they don't, the average voltage will drift as the low-frequency components from the scrambler charge and discharge the coupling capacitors.

Several modifications of the K9NG modem are

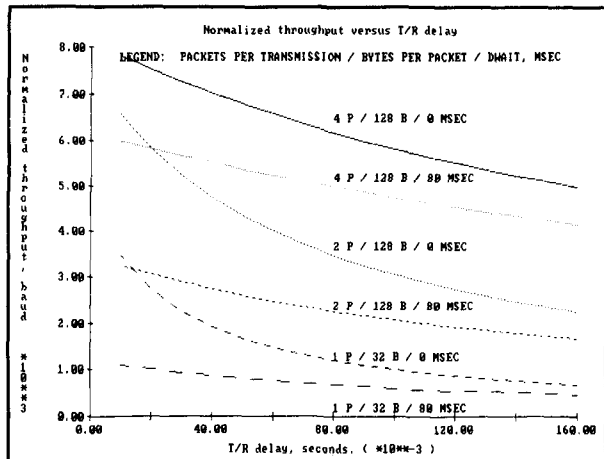


fig. 5. Effective channel capacity vs. T/R delay. This graph shows the effective capacity (baud rate) of a channel for different transmit/receive switching delays, with and without an additional delay, called DWAIT (digipeater wait time). TEXNET does not use digipeaters, so DWAIT = 0.

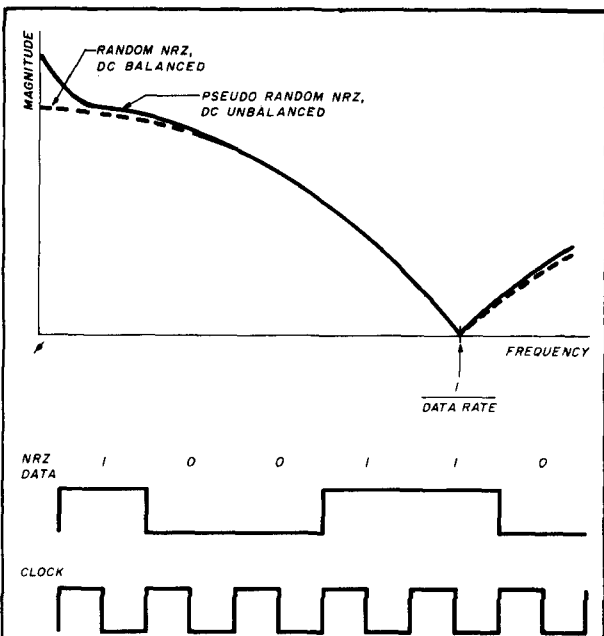


fig. 6. NRZ signal diagram. An NRZ signal has a value of logic zero or logic one for the entire period of the bit. The spectrum of a truly random signal is from dc, diminishing to zero at (1/bit time). A pseudo-random signal has a diminished spectrum near dc, theoretically vanishing.

associated with increasing the time constant of these circuits, where they're not critical to the T/R switching delay. A list of the modifications is included in fig. 8. The most significant improvement came with addition of a group-delay equalizer in the receive section of the modem. In order to appreciate the requirements for this, let's look at how filters work both in the frequency domain and the time domain. A typical i-f filter used in voice or CW work has very sharp "skirts." That is, the amplitude response decreases sharply from the filter center frequency. In addition, the response is relatively flat within the passband. This filter has a great deal of time delay distortion (see fig. 9). The time delay is minimum at the filter center frequency and rises sharply at both the upper and lower filter cutoff points. This causes no particular problems with voice, where the distortion of the waveform isn't important. But with wideband data signals, where we need to distinguish the value of data bits that are adjacent in time, the spectral energy nearer to the filter cutoff points will undergo a greater delay through the filter than the energy near to the center frequency. When this signal is converted to baseband (i.e., fm demodulated), it can be shown that frequencies near the fil-

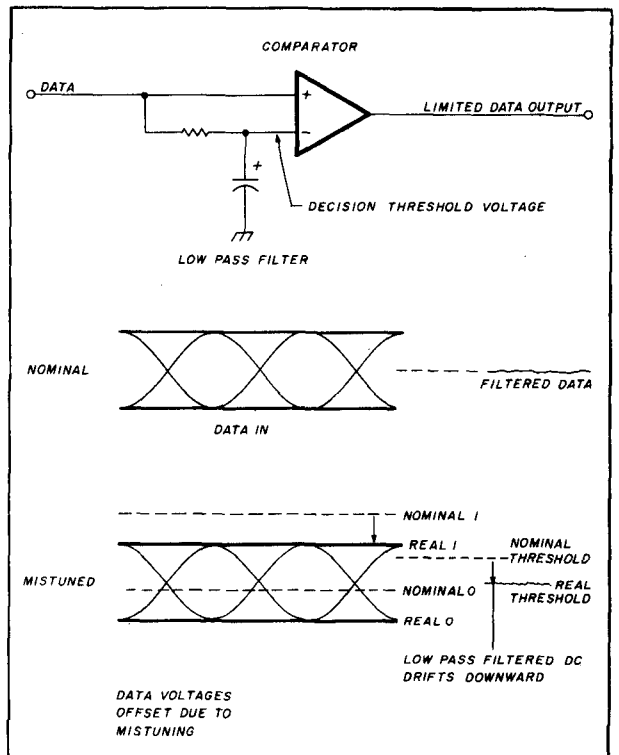


fig. 7. NRZ filtered, slicing level, low-pass filter to generate slicing level vs. frequency drift. If the NRZ is scrambled, so that the code is dc-balanced, then low-pass filtering the data will generate the correct logic-slicing voltage. As the one/zero voltages drift with receiver or transmitter mistuning, the slicing voltage will track automatically.

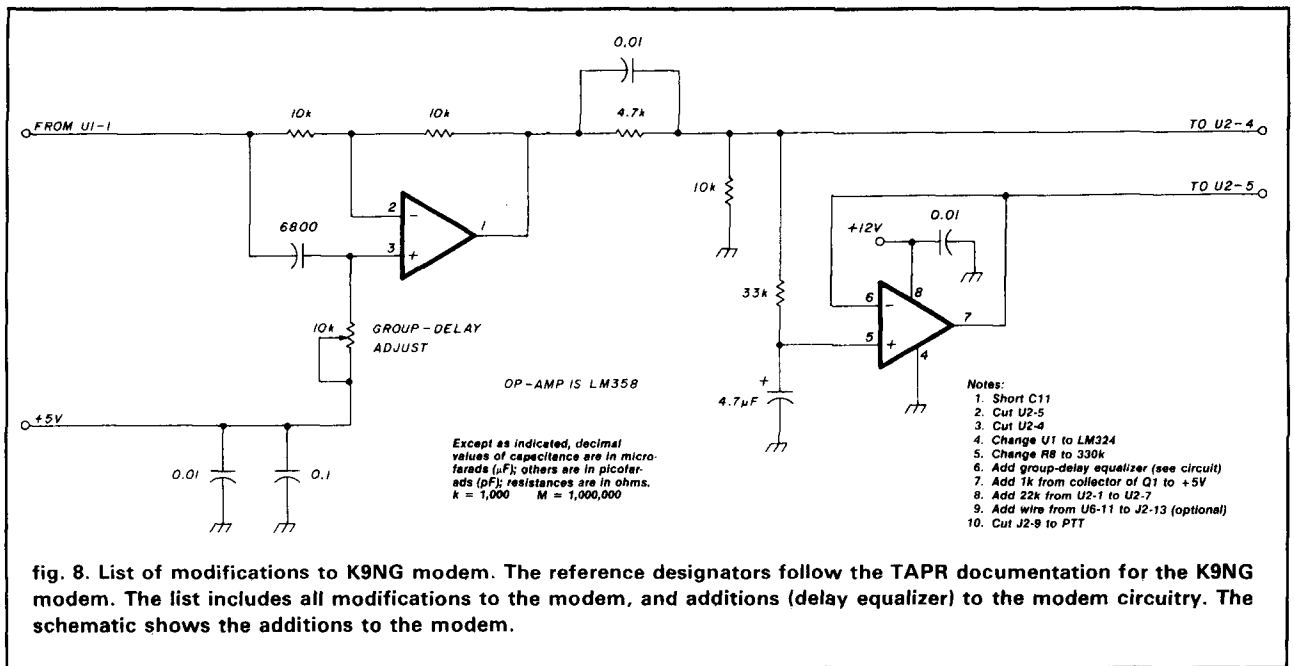
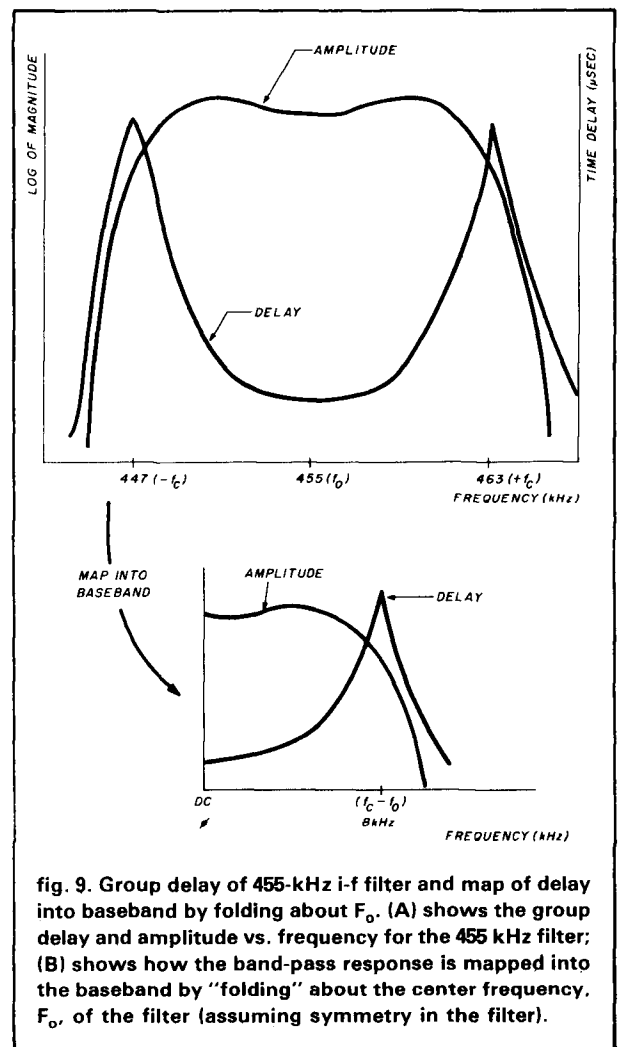


fig. 8. List of modifications to K9NG modem. The reference designators follow the TAPR documentation for the K9NG modem. The list includes all modifications to the modem, and additions (delay equalizer) to the modem circuitry. The schematic shows the additions to the modem.

ter center frequency correspond to the low-frequency spectrum of the NRZ (baseband) signal, while the frequencies near the filter cutoff (away from the filter center) correspond to the higher frequency components of the baseband signal. Thus we can "map" the time delay of the filter into baseband by "folding" the time response about the center frequency, which becomes the zero frequency of the baseband. Thus the filter produces small delay at low baseband frequencies, and larger delay at higher baseband frequencies.

The 455-kHz i-f filter in the FM-5 radio is a sealed ceramic unit; no adjustment is possible. Instead of designing a new filter with desirable time-delay characteristics, we instead built an active filter circuit, with flat amplitude response, but with an adjustable time-delay response. This network has maximum delay at dc, and decreasing delay at higher frequencies. The circuit is adjustable, so that we could construct an approximate inverse time delay to that caused by the i-f filter. Figure 10 shows the amplitude, phase, and time response of this baseband group-delay equalizer (active filter), and the baseband eye pattern with and without the delay being equalized. After delay equalization, it can be seen that the eye is much more open near the center of the bit, when the 1/0 decision is reached. Our measurements indicated a 7-dB improvement. We also provided about 1 dB of peaking of the frequency response, which opened the eye about 1 dB more, yielding an 8 dB improvement in the receiver. Actual tests with the radios indicated that this made the difference between usable and nonusable performance. Without the equalizer, the radios



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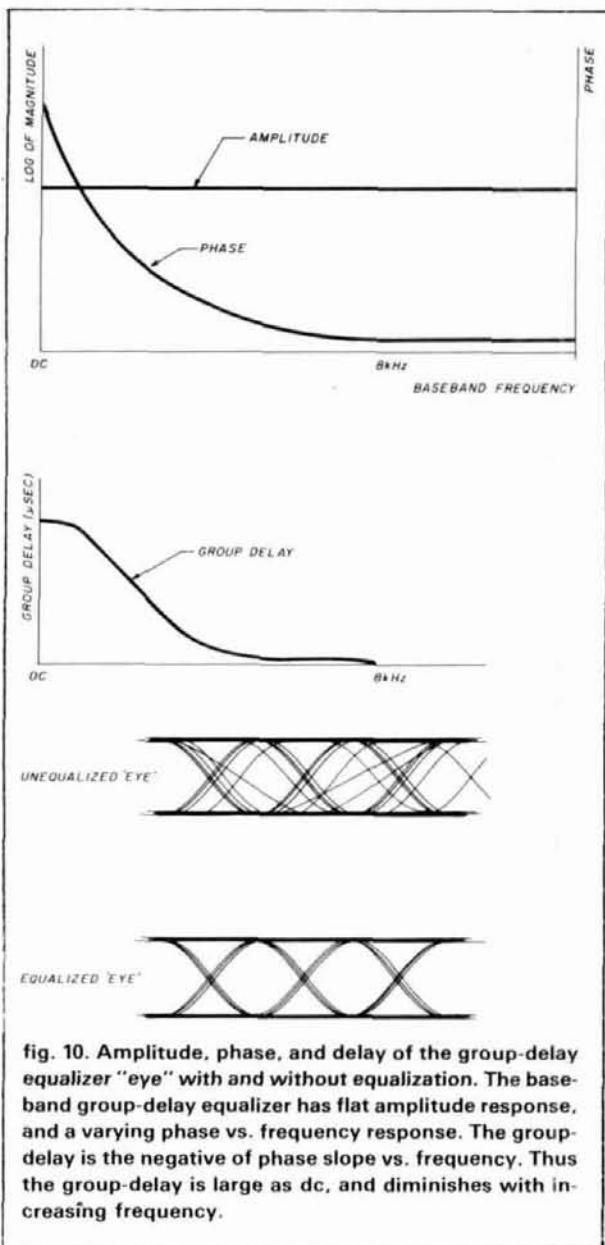


fig. 10. Amplitude, phase, and delay of the group-delay equalizer "eye" with and without equalization. The base-band group-delay equalizer has flat amplitude response, and a varying phase vs. frequency response. The group-delay is the negative of phase slope vs. frequency. Thus the group-delay is large as dc, and diminishes with increasing frequency.

"dribbled" (i.e., had a background error rate regardless of the strength of the received signal), which caused many packets to be lost.

One additional problem that had to be overcome on our "real" path test hop was insufficient image rejection in the FM-5. The test path is in an area where television channels 11 and 13 are very strong. The channel 11 video carrier is near the image frequency of our desired channels (the radio i-f is 10.7 MHz, so the image is 21.4 MHz lower than the signal frequency). **Figure 11** shows a simple filter; **fig. 12** shows its response (S11 and S12). This filter was extremely effective in eliminating the image response.

With these improvements, we've run a 12-mile path on 220.55 MHz at 9600 baud, at better than 98 percent

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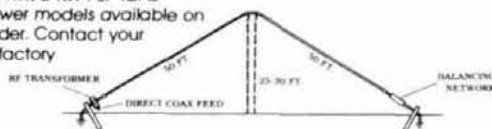
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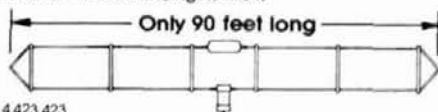
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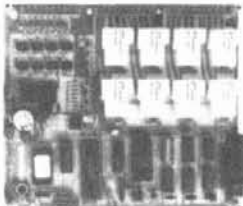
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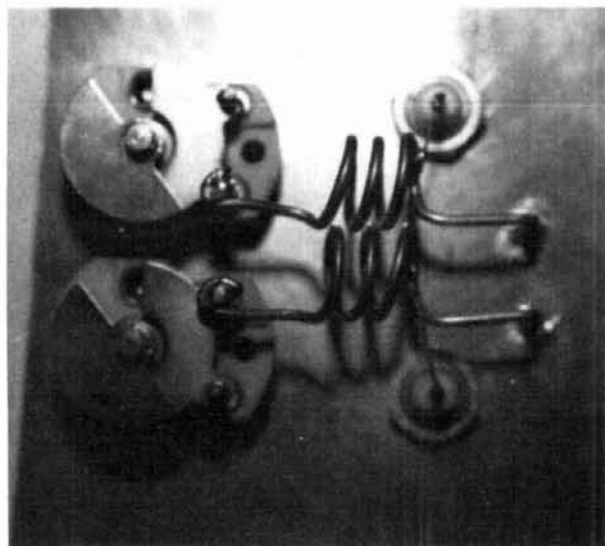


fig. 11. The 220-MHz front-end filter. Air-variable capacitors are necessary for adequate filter Q . The two resonators are closely coupled. The high tap position sets the loaded- Q of the resonators and prevents over-coupling, while maintaining low insertion loss. This filter is exceptionally difficult to tune properly.

reliability — 98 percent of the packets are received without error. This hop is such that 4-dB additional attenuation at one radio caused the packet reliability to be approximately 5 percent, and so was a stringent test of the radios and modems in that the radios were operated near the minimum acceptable received signal level.

The FM-5/K9NG modem experiments gave us valuable insight into proper modem/radio design. We

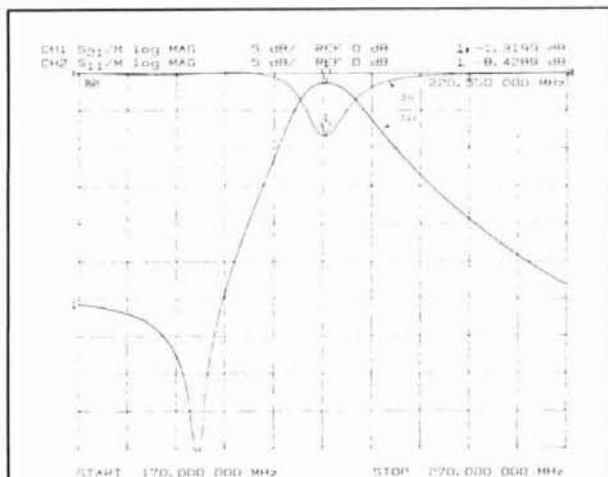


fig. 12. S11 and S12 of the filter. The forward transfer (insertion loss) of the filter is S12, while the return loss (related to input VSWR) is S11. This filter has about 1.3 dB insertion loss, and about 8 dB return loss. The notch is near the image frequency of the 220-MHz receiver.

181



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182

plan to revise the modem circuits, utilizing surplus commercial 440-MHz equipment for the actual network.

NCP design

The next element in the system is a microprocessor performing as the network switching node. The node control processor (NCP, **figs. 13 and 14**) is an original design, though several modifications have been made since the original artwork was done. The design of the unit is conventional, with a few points emphasized for reliability.

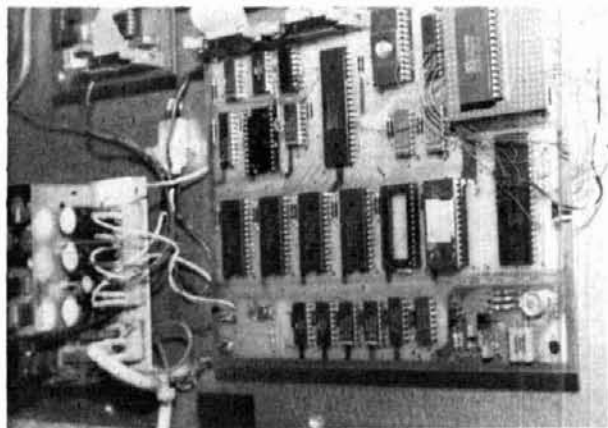


fig. 13. The Node Control Processor (NCP), version 1, with several modifications. A new version 2, which is a multi-channel super-set of the popular TNC-2 unit, is being designed.

The unit contains a Z80 operating at 4 MHz, 32K of EPROM, 32K of RAM, an SIO/2 serial communications IC for the serial HDLC ports, a counter/timer IC (labeled CTC and situated on the modification "ledge" overhanging the main board) for providing the interrupt clock time slices (of 8 ms), and two special circuits. Careful design of the NCP hardware and interrupt daisy chain to match the software was required for the computer to operate reliably at 9600 baud and support multiple HDLC channels simultaneously. All of the I/O devices utilize vectored interrupts through the Z80 interrupt daisy chain. **Figure 15** shows the entire TEXNET node prototype (both radios and modems, ac power supply, and the NCP).

We decided to develop our own board in order to keep costs down and allow the inclusion of two special circuits.

The first circuit is a reliable crystal oscillator. Many logic-gate type of crystal oscillators aren't reliable enough for use in remote, unattended computers. Numerous tests have shown that under certain voltage transients, gate-type oscillators won't start reliably. It's a nuisance to have to climb a tower to recycle the power just to restart a crystal oscillator, and inconvenient to users to have the network out of service during this time. The circuit chosen is a conventional Pierce type, with an additional transistor buffer amplifier. It was tested extensively and found to be robust. (One test to try on an oscillator is to feed the circuit from an adjustable voltage supply. Set the

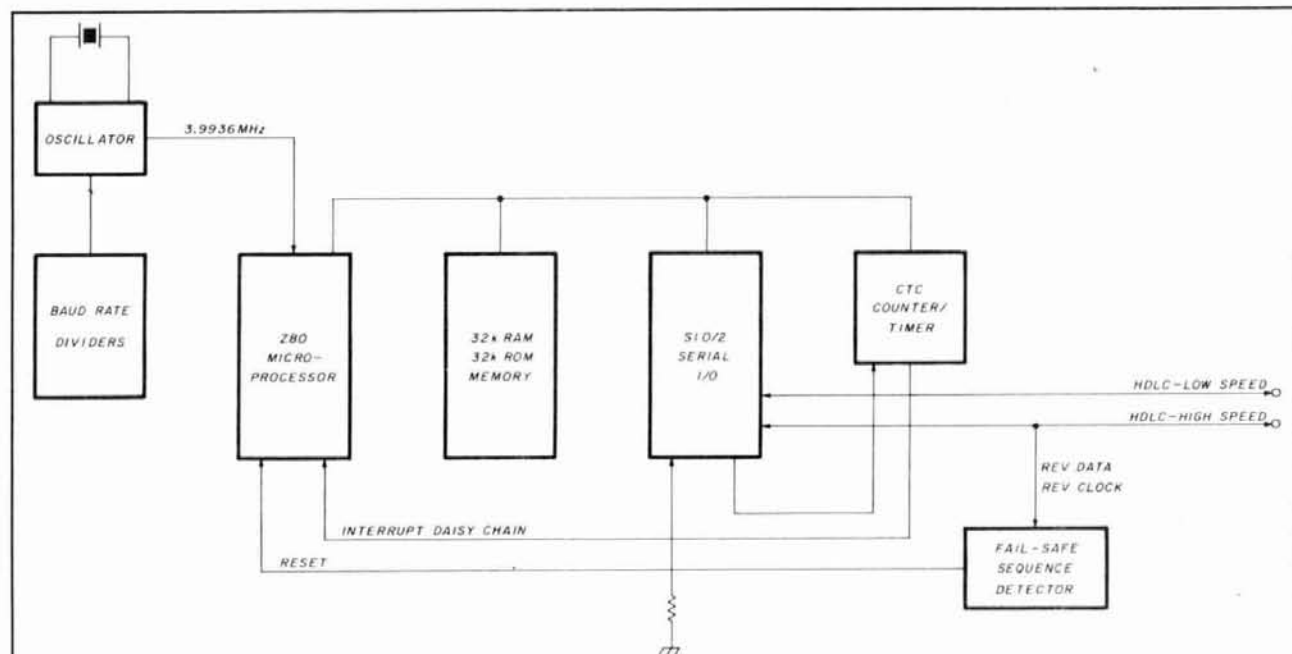


fig. 14. The NCP, showing the crystal oscillator, baud generator, microprocessor, 32 kbytes of EPROM, 32 kbytes of RAM, one HDLC synchronous communication IC, the counter-timer IC, and the fail-safe state machine.

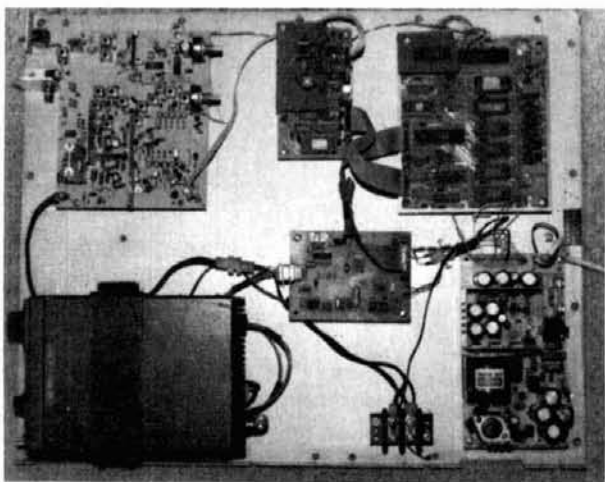


fig. 15. The complete TEXNET node prototype. This is the first prototype of the TEXNET node, which includes the NCP, high-speed modems, the 2-meter radio, the 220-MHz radio, and the ac power supply.

voltage to 0, turn on the power, and very slowly increase the supply potential to the nominal voltage — for example, over a period of 30 seconds. If the oscillator refuses to start, starts on the 3rd harmonic, or starts at the wrong frequency, the circuit should be rejected as unsuitable for unattended operation. Many logic-gate oscillators will fail this test.)

The second special circuit is a fail-safe state machine. This is an EPROM-based logic circuit that monitors the IP data and clock lines (from the high-speed trunk radio), completely independently of the processor or communication ICs. It searches for the presence of a very long (72 bit) sequence. If this sequence is ever detected, the state-machine activates the reset line on the microprocessor, thus restarting the node software. An EPROM contains both the value of the 72-bit sequence and the state-machine code necessary to operate the circuit. Each node is programmed with a different 72-bit sequence, which we've termed the "fire code." Any user of the network can cause the generation of a message with the fire code of a suspected node to be embedded in the message and sent via the network to the suspected node. Thus any node in the network can be rebooted (reset) from any other point within the network. There's a very small chance that ordinary user traffic through the network could resemble the fire code, but since the sequence is so long (72 bits), the mean time between false activation is calculated to be considerably more than 1 million years.

A new version of the NCP will include strappable options, and the circuitry for optional addition of the packet message system (PMS), a 5-megabit hard disk drive-based bulletin board that allows up to ten users

to be connected simultaneously. Automatically accessible from anywhere in the network, it's compatible with the WORLI command set (the most popular packet-radio bulletin board system). **Figure 16** shows the PMS prototype — the world's first turbo-charged TNC-2 clone! This test box contains an ac power supply (vertical), an MFJ-1270 (a TNC-2 clone) with WB5PUC ROM, a disk controller, and a 5-megabit hard drive. The Z80 microprocessor is removed from the TNC-2, and an adapter is plugged in its place. The Z80 plugs into the adapter. The ROM software for both the NCP and the TNC-2 are very similar, except that the NCP supports two radios (or more). The NCP is designed so that it is a superset of the TNC-2 hardware. (Further details will be found Part 3 of this series, which will address software design.)

This concept could be tested as a satellite gateway, perhaps with UOSAT-11, as a store-and-forward message system.

power supply

The power supply for the network node is extremely important in determining the reliability of the network. If the network is to be useful for handling emergency communications, it should be able to survive temporary power outages. Consequently, the TEXNET power supply utilizes a gelled-electrolyte (gel-cel) lead-acid battery, which can provide power for several hours and has a reasonably long life if properly charged and maintained.³ The power supply for the node utilizes +17 through +24 VDC as the input power source, unregulated (but filtered). The supply/charger (see **fig. 17**) regulates the input to +13.8 VDC through

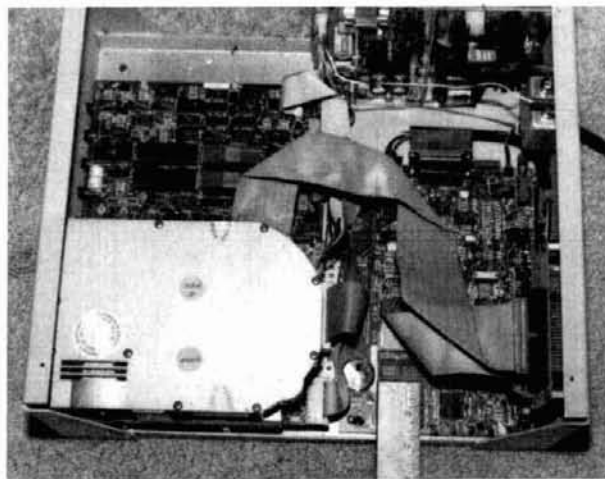


fig. 16. The Packet Message System (PMS), a 5-megabyte hard disk drive-based bulletin board, supports 10 simultaneous users. This prototype includes a TNC-2 clone, the hard drive controller boards, the 5-megabyte hard disk drive, and the ac power supply. A second version is connected to the node prototype, and is accessible from anywhere in the network.

a three-terminal, 10-amp regulator (LM396). The battery charge is controlled by a three-state temperature-compensated charger IC (a Unitrode UC3906). When the ac fails, a relay connects the battery to the load. A capacitor is used to hold up the +5 VDC load voltage for the switching time of the relay. Should the battery completely discharge (i.e., produce less than 12.0 volts dc), the relay protects the battery by disconnecting it from all loads, and the power fails.

The life of a battery is maintained only by very careful charging. This is the reason for the three-state charger. When the battery is initially depleted, it is charged at a constant current of C/10 (at 1/10 the ampere-hour capacity, that's a 10-AH battery charged

at a 1-amp current). When the battery voltage rises to nominal voltage (14.25 VDC), a controlled overcharge is initiated. Here the battery is charged at constant voltage (15.00 VDC) until the charge current decreases to C/100 (100 mA for a 10-AH battery). Failure to apply a controlled overcharge will result in the battery's receiving only 80 percent of its previous charge at 14.00 VDC. These voltages are for operation at +25 degrees C and for gel-cells. Liquid electrolyte batteries require different (i.e., lower) voltages. Variations in temperature require compensation in voltage; if compensation is not made, the battery will be severely undercharged at low temperatures and overcharged at high temperatures. The UC3906 contains most of the circuitry, and a temperature-dependent

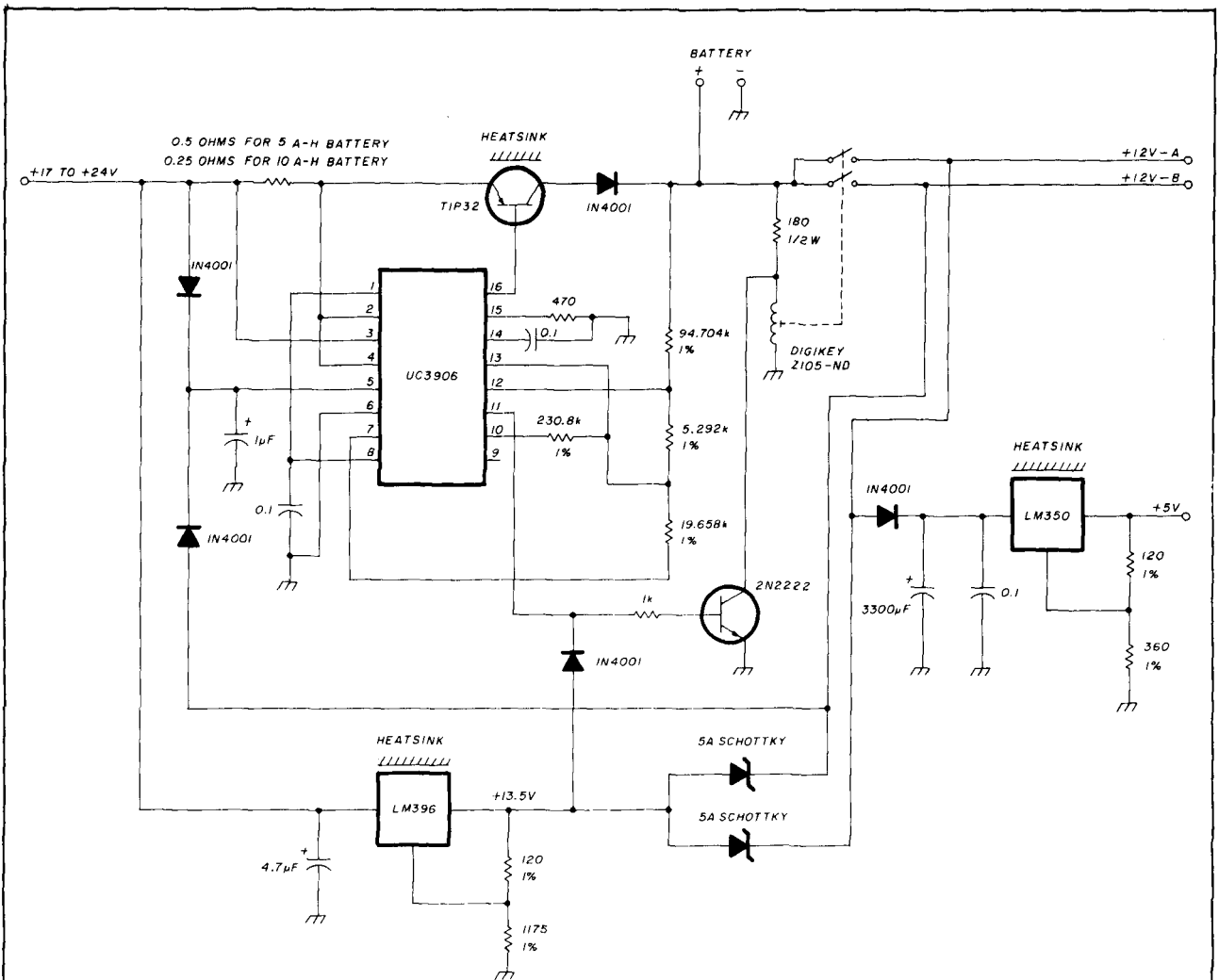


fig. 17. Schematic of battery charger/power supply. The battery charger/power supply is a three-state charger (see text) and power supply regulator. When ac power is present, the battery is charged and the three-terminal regulator supplies power to the node. When ac fails, the battery supplies power to the node. If the battery should completely discharge, the relay will disconnect all loads from the battery, thus protecting it. A large output capacitor on the +12VDC supply to the microprocessor voltage regulator supplies energy for the period of time it takes the relay to energize, preventing a momentary glitch in the +5VDC line.



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voltage reference to accomplish this. An external pass-transistor increases the current handling capability to 1 amp. The relay/transistor circuit connects the battery to the load upon ac failure and disconnects the load when the battery is depleted.

Allowing the battery to remain connected to the load after it's depleted will destroy it. New Gel-Cel batteries can normally be expected to operate for six years when properly used; automotive-type batteries, which are designed for high surge currents, will normally have only about a three-year lifetime in a standby power applications and require different voltages. **Warning: this circuitry is meant only for 12-volt lead-acid gel-cel batteries; its use is not recommended for any other type of batteries. Using Ni-Cad batteries with this charger could result in an explosion.**

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Performance, Reliability, and Customer Support: The Winning Team

While attractive front panels and impressive magazine advertisements may initially glamorize any amateur radio item, they can also reflect the classic proverb of beauty being only skin deep. The favorable returns from any unit and the success of its manufacturer, however, are directly influenced by **after-purchase reliability and factory-backed service**. Knowledge of such performance records and readily available customer support encourage the peace of mind to use and enjoy a new unit to its maximum potential.

ICOM considers the aspect of service from two interrelated standpoints: daily in-field use and possible "down the line" repairs if, and when, needed. This concept is pursued by first building **professional communications quality and reliability into every unit**, confidently backing it with a full warranty, then substantiating that dependability with **uncompromised factory authorized service and customer support**. All ICOM HF transceivers and shortwave receivers reflect that philosophy with their **full one-year warranties**...and service centers that are not bottlenecked with backlogs (stout performers simply give less trouble). ICOM isn't playing down customer support, but building a positive long-term reputation on it!

Today's era of advanced technology and seemingly endless consumers tends to replace old-

time "concerned treatment" with attitudes of "being one of a vast number in line." Returning a unit for adjustment or repair and later attempting to check its status sometimes proves to be a frustrating experience. While no one is infallible, ICOM honestly strives to avoid an attitude of "too many customers to provide congenial service." ICOM's customer service **hotline** at (206) 454-7619, for example, will put you directly in touch with the main service department. The only prerequisite is **mutual understanding** in sharing this resource so everyone can have queries answered and radios repaired. If a problem can't be alleviated via telephone, ICOM strives for a service center "turnaround time" of three to five days.

Continuing that customer support, **ICOM is the only amateur radio company with four factory-owned service centers in North America**. The centers are located in Atlanta, Georgia; Dallas, Texas; Bellevue, Washington; and Vancouver, British Columbia. Most ICOM service centers are also situated near major airports to further minimize transportation problems.

The amateur radio industry is ICOM's major interest; it's not a sideline or spin-off of other pursuits. ICOM doesn't manufacture stereos, VCRs, or televisions. ICOM is communications industry oriented with secondary involvement in top quality marine, land mobile, and avionics equipment.

The stouthearted reliability of ICOM equipment is continuously praised in testimonial letters from proud owners. A few samples from those "believe it or not" files include stories of transceivers literally drowned in salt water two or three hours, yet continuing to operate flawlessly...of no failures to date in the IC-735 and IC-751 power amplifiers...of handheld transceivers dropped from towers, and one was even run over by a truck(!), yet continued to operate after outer case repairs (fortunately, ICOM handhelds include a separate metal frame to protect PC boards and a high impact plastic "outer case").

The next time you switch on a deluxe HF transceiver, compact VHF mobile rig or handheld FM unit, pause a couple of seconds and think about its less apparent aspect of customer support and service. Who would you call if a problem arose, what would be their attitude, and approximately how long might you anticipate being off the air? If you're a proud ICOM owner, those answers are reassuring rather than aggravating.

Again, ICOM's dedication to top performance, exceptional reliability and unsurpassed customer support may not be visible on a front panel or in a colorful ad, but they're **included in every ICOM item**. ICOM equipment is simple to use and the best in quality. It's "Simply the Best" and an increasing number of amateurs are proving that statement in their setups every day. Isn't it time you, too, joined the ICOM winning team?



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grounding, shielding, and isolating: part 1

Most users of electronic equipment have to face problems of grounding, shielding, and isolating in order to prevent EMI (electromagnetic interference). Amateur Radio operators have an especially difficult problem because their equipment often must operate on the same desk as their high power radio transmitters. EMI is also caused by local electrical disturbances such as arcs, lightning discharges, and electrical motors.

The flip side of the EMI problem is prevention of interference to other equipment — such as televisions, radios, stereo systems, and other entertainment equipment — caused by Amateur transmitters. We must operate our stations in a manner that won't interfere with properly designed, properly operated equipment that's in good repair. Electronic equipment must do two things: respond to desired signals and reject undesired signals. There's not much we can do if the equipment fails to do the latter; if it meets both tests, however, our efforts can help profoundly.

In this installment, we'll discuss methods of preventing interference from outside and cross-interference between circuits or equipment in our own stations.

preventing EMI

Shielding and filtering of signal lines is the key to preventing EMI. Figure 1 shows a "generic" electronic device with several of the possible EMI protection methods used. This type of circuit could be a microphone input preamplifier, a receiver accessory, or some other piece of equipment. First, note that the entire instrument is built

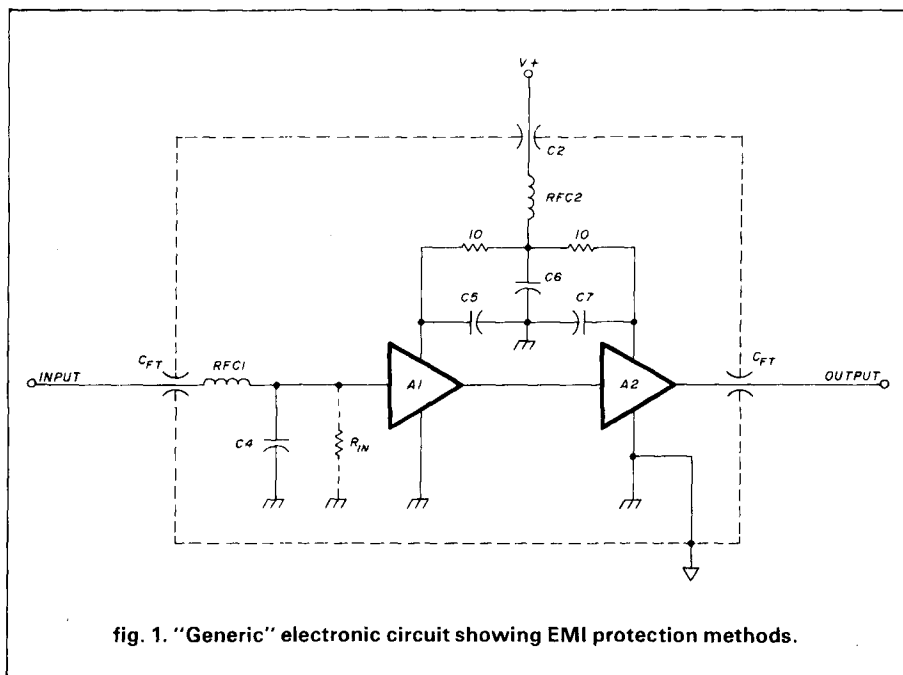


fig. 1. "Generic" electronic circuit showing EMI protection methods.

inside a shielded metal box, and the box is grounded. Points of entry and exit incorporate feedthrough "EMI filter" capacitors (C_{FT}) (500 pF to 0.002 μ F). Each stage is isolated from other stages by a resistor, and has its own decoupling capacitor ($C5$ and $C7$). The main power bus is decoupled ($C6$), and has a series rf choke ($RFC2$) to filter rf that gets past $C2$ and prevents it from interfering with the operation of the circuit. The input leads are similarly filtered with $RFC1$ and $C4$. The input resistance R_{IN} of the amplifier and the reactance of capacitor $C4$ also form a low-pass filter with a frequency response that rolls off at a 6 dB/octave rate from the -3 dB point defined by:

$$f(\text{Hz}) = \frac{10^6}{6.28 \cdot (C4) (R_{IN})}$$

where: f is the frequency in Hz, $C4$ is in μ F, and R_{IN} is in ohms.

One potential source of interference is noise and other EMI signals on power lines. I can recall troubleshooting digital instrumentation and computers in a medical school building. We found that the ac power lines were the source of the problem. Where sensitive electronic instruments are used, one might want to consider designing the electrical system to be either isolated from the building system or have separate neutral and ground conductors all the way back to the building's electrical service entrance ground.

Figures 2 and 3 show methods for dealing with power line noise. In fig. 2, we see an LC power line filter. Although the version shown here uses just one LC section, others are available (at higher cost) with two or more sections. These devices are shielded low-pass filters, and are mounted inside equipment as close as possible to the point where ac enters the cabinet.

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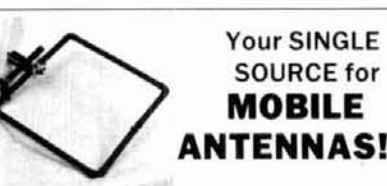
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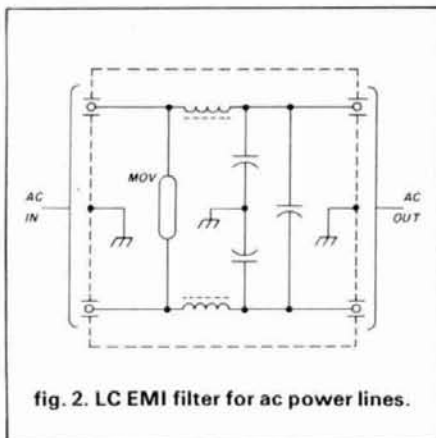
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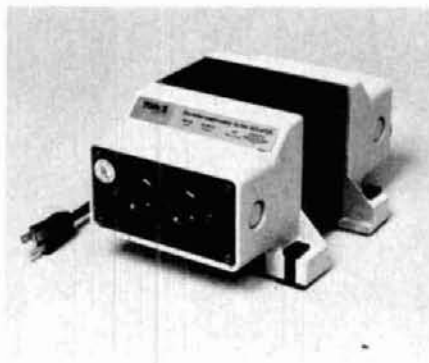


Some filters are available molded into the ac chassis connector.

Also in the circuit is a Metal Oxide Varistor (MOV) device used to suppress ac line transients above about 155 volts peak (some can reach 2000 volts for 30 microseconds). Power line transients can have surprisingly sharp rise times, and thus are capable of producing interference across a wide portion of the spectrum. Because of this some designs have these MOV devices ahead of the filter in order to reduce the intensity of peaks that contain high frequency components.

There is an issue regarding EMI filters that some people overlook: fusing. Ideally, the fuse should be as close as possible to the point where the ac line enters the cabinet. If the fuse is placed downstream from the filter, the equipment won't be protected if components of the EMI filter short to ground. If the ac power line is routed through the fuse holder before passing through the EMI filter, on the other hand, there's enough prefiltered wiring to form an antenna and radiate into the other circuits. As a result, I prefer to build the fuse either into the EMI filter (or an extra shielded compartment that also houses the EMI filter) or into a fused ac plug at the wall outlet.

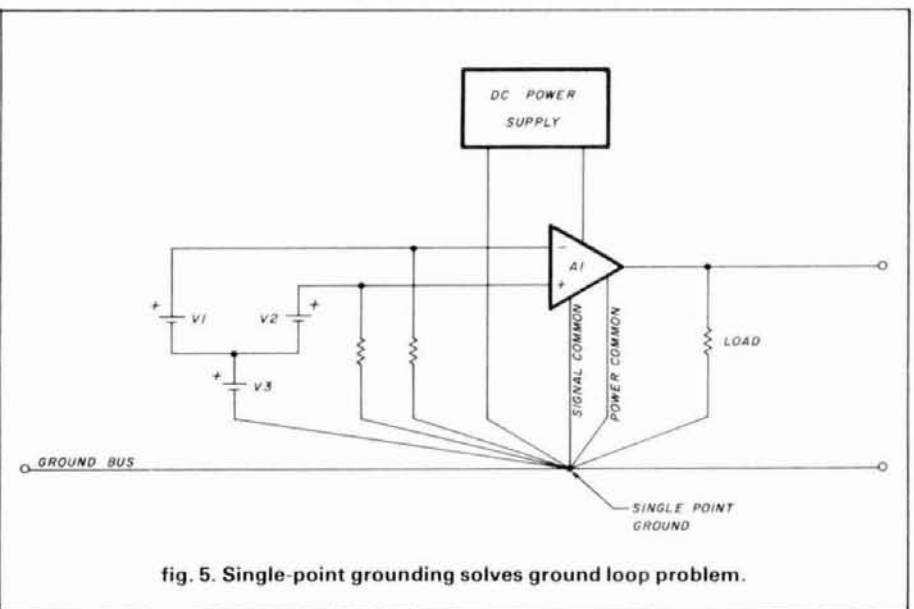
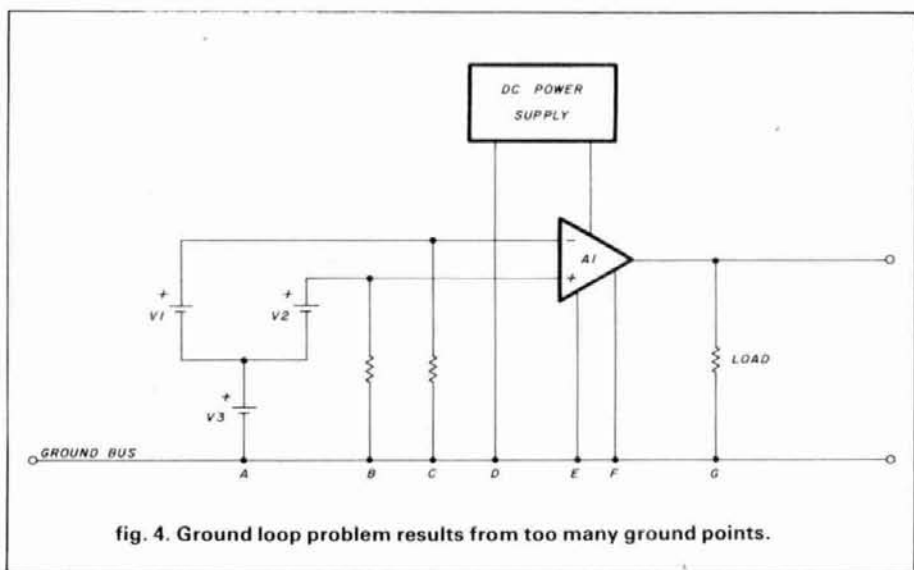
The Topaz transformer shown in fig. 3 performs three functions. First, it isolates the equipment electrical system from the ac electrical system. Second, it frequency-limits the system to prevent high frequency transients



Power line isolation transformer (Courtesy Topaz Electronics, San Diego, CA).

and pulses from passing into the equipment. Third, it provides a locally isolated ac power mains system that isn't ground referenced, making death or serious injury less likely in the event of accidental contact with the 110 or 220 volt ac lines. *All workbenches should have isolated power.*

My opinion, which is shared by many others, is that no computerized or other digital equipment — or many types of analog and rf equipment — should be operated in a noisy power line environment without one of these transformers. This is especially true if



(continued from page 4)

220 AT RISK — ACT NOW

with radiolocation through 1990, when that use will be largely phased out). Not a bad tradeoff, to go from 5 MHz (shared with two other hungry services and thus a potential roosting place for every channel-seeking schemer who's come along in the last decade) to a 3-MHz band that's as firmly ours as any Amateur band can be, and thus likely to stay that way for the foreseeable future!

But this is not 10 years ago, and that "lightly occupied" 220-225 MHz band is indeed *fully* occupied, particularly in most major urban areas. Even without Novice enhancement, the proposed reallocation will cause severe disruption to important, established Amateur activities. With the addition of Novice operators to the already crowded — not "lightly loaded" — 220 MHz band, perhaps the pundits who thought this one up will then decide that it's really 2 meters that's "lightly loaded" and declare that 2 meters is the one to be shared with others!

Can this proposal be stopped? Very possibly not — but every effort must be made and made *quickly*. **The deadline for comments on General Docket 87-14 is April 6**, though several petitions to have this date extended are being filed. A minimum of one original and five copies of your comments are required; 11 copies are necessary if you wish each Commissioner to receive one. To be effective, cite occupancy facts and support use projections. "You can't do this to us!" isn't a very valid argument in Washington!

The FCC does indeed giveth . . . and may well taketh away, too!

— Joe Schroeder, W9JUV

short circuit

W6SAI

The inset table shown in **fig. 1** of W6SAI's February, 1987, column ("Ham Radio Techniques," page 45) should be corrected to read as follows: Tuned input network to match 420 ohms ($Q=4$)

f(MHz)	C ₁ (pF)	C ₂ (pF)	L ₁ (μH)
3.5	605	361	6.30
7.0	302	180	3.15
14.0	151	90	1.57
21.0	100	60	1.05
28.0	75	45	0.79

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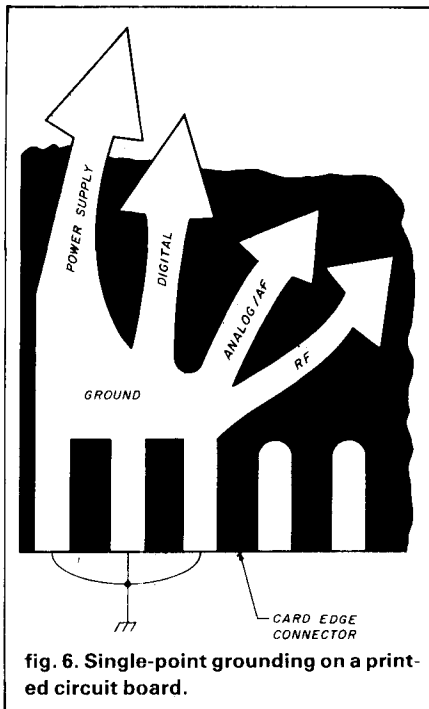


fig. 6. Single-point grounding on a printed circuit board.

the operation of the equipment is critical.

Although most Amateur equipment doesn't fall into this category, there are some cases where it might. For example, suppose you have a computer-controlled repeater transmitter. A power line transient could easily upset the program, causing it to "bomb" with the transmitter on . . . with no way for a control operator to turn it off without a personal visit to the site. While this problem is easily overcome by good design, I was told of one such situation in the Midwest, so I have to assume that others may also encounter it.

common-mode rejection

Other electrical devices nearby, as well as the ubiquitous 60-Hz field from building wiring, can induce signals into audio and low-frequency amplifier in-

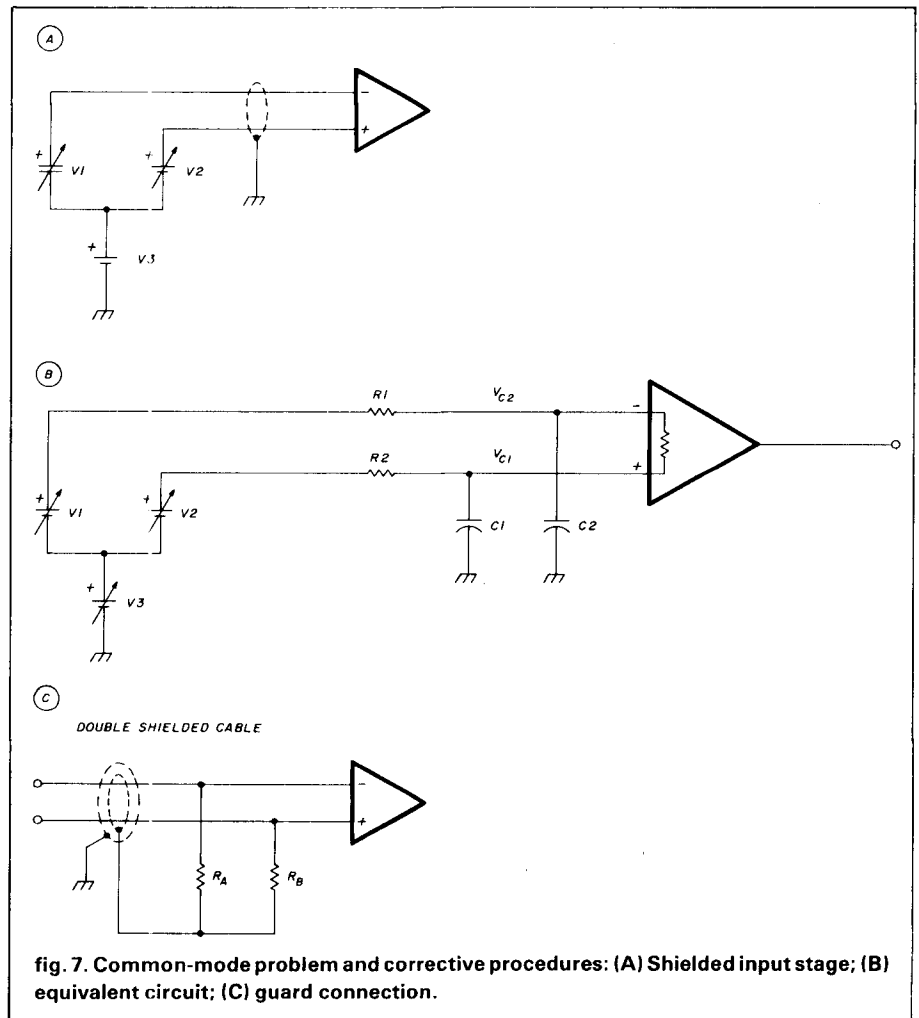


fig. 7. Common-mode problem and corrective procedures: (A) Shielded input stage; (B) equivalent circuit; (C) guard connection.

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puts. It's wise to use differential amplifiers (where possible) for these applications because of their high common-mode rejection ratio (CMRR). Signals from desired sources are connected as differential signals (V1 and V2 in **fig. 4**) across the two inputs, while interference from the 60-Hz lines tends to be common mode because it affects both inputs equally.

It's sometimes possible, however, to manufacture a differential signal from a common-mode signal. This can occur in two ways; both involve the improper use of shields. One source of the problem is the ground loop, as shown in **fig. 4**. This problem arises from the use of too many grounds. In this example the shielded source, shielded input lines, the amplifier, and the power supply are all grounded to different points on the ground plane. Power supply dc currents flow from the power supply at point "D" to the amplifier power supply common at point "F," forming a voltage drop along the way. Similarly, other sources

also cause ground plane voltage drops. Known as ground loop signals, these signals form valid input signals from the amplifier's "point of view."

The cure for ground loop signals is shown in **fig. 5**. In this example we see that all of the ground connections within the equipment are routed to a single common grounding point. This effectively eliminates the ground loop voltage drops. Also, note that rather than allowing each line to have its own shield, a single shield around both lines is used.

When you're designing a new system — and have the opportunity to design the printed circuit boards — make sure that single-point grounding is used. **Figure 6** shows a method for minimizing the noise problems in a circuit board. Note that four different grounds are used: one each for the power supply, the digital signal, the analog signal, and rf. All four grounds are joined together at a single point on the card edge connector and then spread out to their respective circuits.

common-mode signals

Figure 7 shows the causes and cures for another form of signal error: common-mode signals *manufacturing* differential signals. The circuit in **fig. 7A** uses standard single shielding, but the equivalent circuit shown in **fig. 7B** reveals the problem. The shield produces a capacitance to ground with the input wires (C1 and C2). In addition, there are cable and source resistances in the circuit, represented in **fig. 7B** as R1 and R2. The system works well if $R1/XC1 = R2/XC2$, but even small imbalances in the RC networks will allow common-mode voltages to manufacture a differential signal. In that case, it's found that V_{C1} does not equal V_{C2} so the amplifier sees what it accepts as a valid input signal.

A "guard shield" (**fig. 7C**) circuit can be used to overcome this problem. The guard shield is driven by signals from the two input lines summed together through high-value resistances, R_A and R_B and, in many

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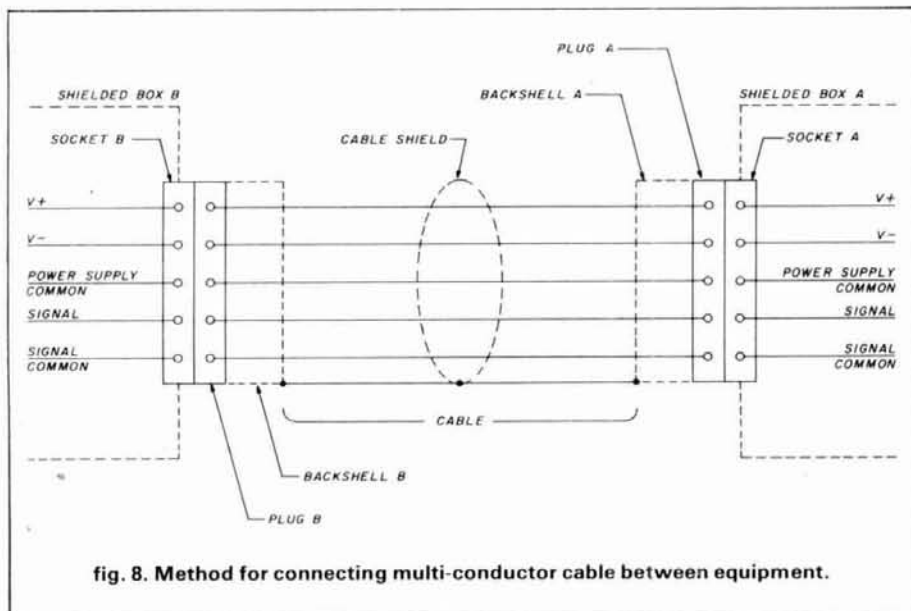


fig. 8. Method for connecting multi-conductor cable between equipment.

cases, a unity gain common-mode amplifier. The method has the effect of placing both sides of the cable capacitances at the same potential, so $V_{C1} - V_{C2} = 0$. The outer shield isn't strictly necessary, but is highly recommended in rf-rich environments.

don't run signals through shields

Another shielding scheme for connecting two pieces of related equipment is shown in fig. 8. This situation involves two pieces of equipment

(identified here as "A" and "B") that pass multiple connections back and forth, including signals and power. While traditionally not common in Amateur equipment, this arrangement is becoming more so with the increasing use of interfaces between computers and Amateur Radio equipment (as in packet radio). A common error made in constructing the multiconductor cables is to pass signal or power common return paths through the shield itself. All such common paths should have their own separate wire in the cable bundle. Some internal conductors may be individually shielded. The shield is connected to the shielded backshells of each connector, but doesn't carry signals. It's sometimes permissible to ground the shield to the chassis through a pin in each connector, provided that it too does not carry signals or power.

Next month, we'll take a look at station grounding and other methods of preventing TVI that we might cause.

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At last! I'd stuck the final red pin in the map of the world. Covered with red pins, one for each DX country confirmed, and hung proudly on the wall of the operating room, it instantly showed visitors my DX prowess.

I had some good ones. Andaman Islands? Yes. Burma? Yes. Franz-Joseph Land? Yes. Tibet? Yes. Hours of DX operation had covered the map except for a gaping hole near the center of Asia. A no-man's-land of zero Amateur activity! I'd worked many stations around the perimeter of this mysterious "black hole," which measured about 600 miles in diameter.

This blank zone encompassed what was once known as "Chinese Turkestan," with the city of Urumchi — now known as Urumqi — as the focal point of interest. Located in western China, nestled between India and the USSR, Xinjiang Province (as it's now known) has never, as far as I know, been on the ham bands. (There was a rumor, circa 1948, of a certain C5YY who was supposed to be active from Urumchi, but nothing ever came of this story.)

Rumor has it that a new station will soon be on the air in the western section of the People's Republic of China

using the call BY0AA. Perhaps this will be the station to represent the last frontier of DX — Xinjiang Province!

intermediates and prefixes

What seems simple now was rather complicated in the early days of radio. For example, you hear G6ZO working KH6BZF. You know it's a QSO be-

Table 1. Post-World War I international Amateur call numbers.

Number	Country
1	Italy, Luxembourg
2	England
3	Poland
4	Germany, Belgium
5	England
6	England
7	Denmark
8	France
9	Switzerland
0	Netherlands

tween England and Hawaii by the prefix letters of the calls. Easy. But return, in your mind's eye, to the early days. When Amateur Radio began back before World War I, the infant hobby had no assigned call letters. Hams used their initials or an abbreviation of their town or city. SNJ was in Hartford and HU was in Honolulu.

Just before the war, licensing was

Table 2. Original Amateur international intermediates (reprinted from QST, March, 1923).

A—Australia
 AU—Alaska*
 B—Belgium
 BE—Bermuda
 BZ—Brazil
 C—Canada and Newfoundland
 CH—Chile
 CR—Costa Rica
 D—Denmark
 E—Spain
 F—France
 FI—French Indo-China*
 G—Great Britain
 GI—Ireland*
 H—Helvetia (Switzerland)
 HU—Hawaiian Islands
 I—Italy
 IC—Iceland*
 J—Japan
 K—Germany
 L—Luxembourg
 LA—Norway*
 M—Mexico
 N—Netherlands
 O—South Africa
 P—Portugal
 PE—Palestine*
 Q—Cuba
 R—Argentina
 S—Scandinavia (Denmark, Finland, Sweden)
 SR—Republic of Salvador
 U—United States
 Y—Uruguay
 Z—New Zealand

*These intermediates have been self-assigned and are unofficial. They are in more or less general use, however.

instituted in the United States and call areas were initiated. About the same time, this happened in other countries. After the end of World War I, when Amateurs returned to the air, the situation was much clearer. In Europe, an agreement between countries assigned numbers to various countries

Table 3. "New" international intermediates (reprinted from QST, January, 1927).

NEW INTERNATIONAL INTERMEDIATES, EFFECTIVE 0000 G.M.T., FEB. 1, 1927.

EUROPE

- EA—Austria
- EB—Belgium
- EC—Czechoslovakia
- ED—Denmark and Faroe Ids.
- EE—Spain and Andorra
- EF—France and Monaco
- EG—Great Britain and Northern Ireland
- EH—Switzerland
- EI—Italy
- EJ—Jugo-Slavia
- EK—Germany
- EL—Norway, Spitzbergen and Franz Josef Land
- EM—Sweden
- EN—The Netherlands
- EO—Irish Free State
- EP—Portugal, Madeira Ids., and the Azores
- EQ—Bulgaria
- ER—Rumania
- ES—Suomi (Finland)
- ET—Poland, Esthonia, Latvia, Courland and Lithuania
- EU—U. S. S. R. ("Russia"), including Ukraine
- EV—Albania
- EW—Hungary
- EX—Luxemburg
- EY—Greece
- EZ—Zone of the Straits

ASIA

- AA—Arabia
- AB—Afghanistan
- AC—China (including Treaty Ports), including Manchuria, Mongolia, and Tibet.
- AD—Aden
- AE—Siam
- AF—French Indo-China
- AG—Georgia, Armenia and Azerbaijan
- AH—Hedjaz
- AI—India (and Baluchistan) and Goa
- AJ—Japan and Chosen (Korea)
- AK—(Unassigned)
- AL—(Unassigned)
- AM—Federated Malay States (with Straits Settlements)
- AN—Nepal
- AO—Oman
- AP—Palestine
- AQ—Iraq (Mesopotamia)
- AR—Syria
- AS—Siberia, including "Central Asia"
- AT—Turkey
- AU—(Unassigned)
- AV—(Unassigned)
- AW—(Unassigned)
- AX—(Unassigned)
- AY—Cyprus
- AZ—Persia

NORTH AMERICA

- NA—Alaska
- NE—Bermuda Id.
- NC—Canada, Newfoundland and Labrador
- ND—Dominican Republic
- NE—(Unassigned)
- NF—Bahama Ids.
- NG—Guatemala
- NH—Honduras
- NI—Iceland
- NJ—Jamaica
- NK—(Unassigned)
- NL—Lesser Antilles
- NM—Mexico
- NN—Nicaragua
- NO—British Honduras
- NP—Porto Rico and Virgin Ids.
- NQ—Cuba and Isle of Pines
- NR—Costa Rica
- NS—Salvador
- NT—Haiti
- NU—United States of America
- NV—(Unassigned)
- NW—(Unassigned)
- NX—Greenland

- NY—Panama
- NZ—Canal Zone

SOUTH AMERICA

- SA—Argentina
- SB—Brazil, Trinidad Id., and St. Paul Id.
- SC—Chile
- SD—Dutch Guiana
- SE—Ecuador and Galapagos Archipelago
- SF—French Guiana
- SG—Paraguay
- SH—British Guiana
- SI—(Unassigned)
- SJ—(Unassigned)
- SK—Falkland Ids. and Falkland Dependencies
- SL—Colombia
- SM—(Unassigned)
- SN—Ascension Id.
- SO—Bolivia
- SP—Peru
- SQ—(Unassigned)
- SR—(Unassigned)
- SS—(Unassigned)
- ST—(Unassigned)
- SU—Uruguay
- SV—Venezuela and Trinidad
- SW—(Unassigned)
- SX—(Unassigned)
- SY—(Unassigned)
- SZ—(Unassigned)

AFRICA

- FA—Abyssinia
- FB—Madagascar, Reunion Id., Comoro Id., etc.
- FC—Belgian Congo, Ruanda, Urundi
- FD—Angola and Kabinda
- FE—Egypt
- FF—French West Africa, including French Sudan, Mauritania, Senegal, French Guinea, Ivory Coast, Upper Volta, Dahomey, Civil Ter. of the Niger, French Togoland, etc.
- FG—Gambia
- FH—Italian Somaliland
- FI—Italian Libya (Tripolitania and Cyrenaica)
- FJ—Somaliland Protectorate and Socotra
- FK—Kenya, Zanzibar Protectorate, Uganda, Anglo-Egyptian Sudan, and Tanganyika Territory.
- FL—Liberia
- FM—Tunisia, Algeria, Morocco (including the Spanish Zone), Tangier
- FN—Nigeria
- FO—Union of South Africa, Northern and Southern Rhodesia, Bechuanaland Protectorate, and Southwest Africa
- FP—Portuguese Guinea and Cape Verde Ids.
- FQ—French Equatorial Africa and Cameroons
- FR—Rio de Oro and adjacent Spanish Zones, Ifni, and Canary Ids.
- FS—Sierra Leone
- FT—Eritrea
- FU—Rio Muni (Spanish Guinea) and Fernando Po
- FV—French Somaliland
- FW—Gold Coast Colony, Ashanti, Northern Territories and British Togoland
- FX—Seychelle Dependencies
- FY—(Unassigned)
- FZ—Mozambique

OCEANIA

- OA—Australia (and Tasmania)
 - OD—Dutch East Indies*
 - OE—Melanesia*
 - OH—Hawaiian Ids.
 - OI—Micronesia*
 - OO—Polynesia*
 - OP—Philippine Ids.
 - OZ—New Zealand
- * To be further partitioned when activity warrants.

SHIP STATIONS

Ship stations with amateur calls will place an X before their usual intermediate. E. g., Australian 3AA at sea, calling U. S. 1AW, would send "1AW NUXOA 3AA" The reply would be "3AA XOANU 1AW".

(see **table 1.**) By 1923, this system had been replaced by letters, and calls began to resemble those with which we're familiar today. An interesting identification scheme was adopted; these new "intermediates," as they were called, are listed in **table 2.**

Viewed from today's perspective, the use of the intermediate, written as a lower case letter, seems strange. English 6ZO would have had an intermediate call of g6ZO. But the intermediate wasn't sent *with* the call, but instead was combined with the intermediate of the station being called. For example, assume English 2SZ was calling New Zealand 4AA. The call sequence would have gone like this: 4AA 4AA 4AA zg 2SZ 2SZ 2SZ. When 4AA replied to 2SZ, the sequence was: 2SZ 2SZ zg 4AA 4AA.

This didn't lead to rapid, simple exchanges of QSOs, but it was a step in the right direction.

By 1927 the International Amateur Radio Union started a serious program to straighten out the international identification of Amateur calls. A new list was formed (**table 3**). The first letter of the prefix denoted the continent and the second the country. The cumbersome intermediate scheme still was used, but some hams were signing their call and intermediate together and using the French word "de" (from) between the call of the sending station and the station being called, much as is done today.

The United States intermediate "U" was replaced with "NU." Shortly thereafter, the Government started issuing full calls with the prefix letter "W" denoting the United States, as determined by the International Telecommunications Union. The old intermediate system faded into oblivion: government assigned prefixes were in use, and the intermediate "de" was accepted by Amateurs worldwide.

It wasn't an easy task to arrive at internationally accepted call letters. As I said, what seems so simple now was very complex in the early days.

Some old timers have lived through all the call letter changes. The famous call AC4YN was in use until the post-

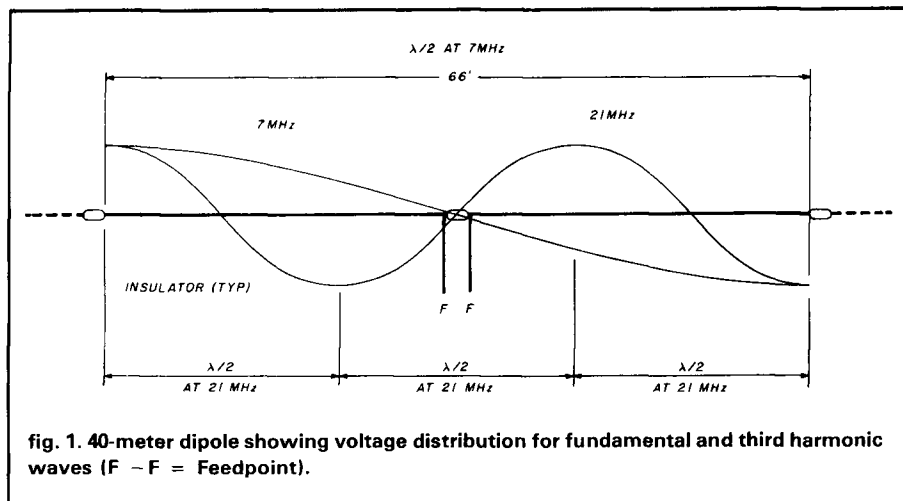


fig. 1. 40-meter dipole showing voltage distribution for fundamental and third harmonic waves (F - F = Feedpoint).

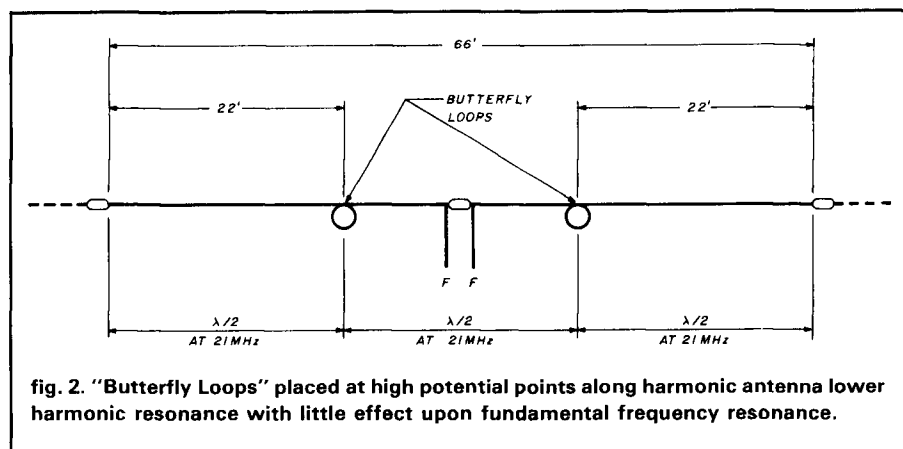


fig. 2. "Butterfly Loops" placed at high potential points along harmonic antenna lower harmonic resonance with little effect upon fundamental frequency resonance.

war period when China took over control of Tibet. The old prefix "FB," for Madagascar, was in use until that country achieved independence from France.

So the next time you hear an unusual prefix on the band, you can appreciate the years of trial-and-error effort that went into establishing a workable call sign system.

wire multiband antennas

It's well known among Amateurs that a center-fed dipole antenna will work (almost) on its third harmonic. Some hams have had good success using a 7-MHz dipole on the 21-MHz ham band. Unfortunately, because of end effects, the dipole isn't exactly resonant at the third harmonic, but at some frequency slightly above it. Thus a dipole cut for 7.1 MHz will be resonant at a frequency higher than the 21

MHz band. For resonance on the third harmonic, the wire should be 68.46, not 66 feet long (**fig. 1**). The general formula for a harmonic antenna is:

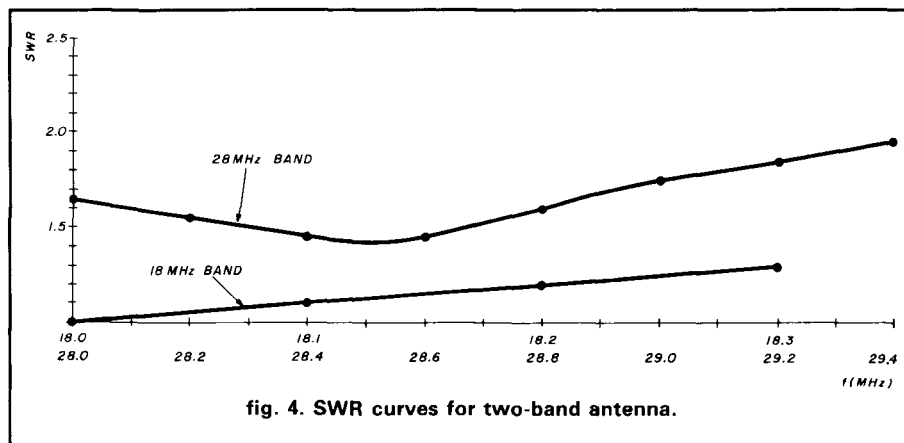
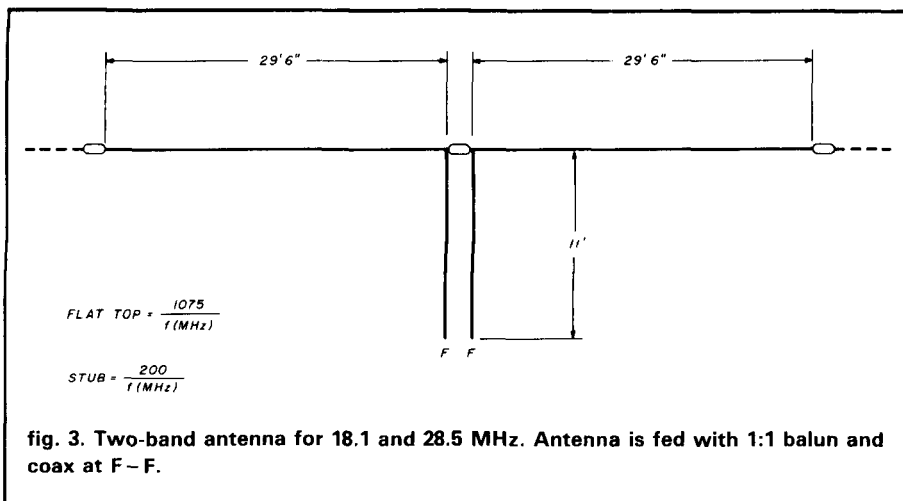
$$\text{Length (feet)} = \frac{492 (N - 0.05)}{f \text{ (MHz)}}$$

where N is the number of half waves in the antenna.

This formula holds true when the antenna is a straight wire. But what happens when the wire is bent back upon itself? Or, what happens when the wire is artificially loaded for harmonic resonance?

harmonic "loading"

Let's take the case of the 7-MHz dipole (**fig. 2**). As is, its fundamental resonance is 7.1 MHz. Its third harmonic resonance is 21.95 MHz. The problem is to load the antenna to a lower frequency in the 21-MHz region without



disturbing its resonance in the 40-meter band. This can be done by attaching "butterfly loops" at the high voltage points in the antenna corresponding to 21 MHz operation. The loops will have little, if any, effect on 7-MHz operation. The size of the loops must be determined experimentally, and they should be equal in size. The loop need not be round — almost any shape seems to work, and resonance is adjusted by changing the shape and size of the loop. I made my loops out of No. 12 enamel wire, with a diameter of about 1 foot. The loops were attached to the flat-top temporarily by means of small copper battery clips. After loop placement was adjusted for resonance at 21.2 MHz, the loops were soldered permanently to the antenna wire. I varied the loop diameter several times before I finally hit resonance (as determined by an SWR meter) at

my chosen frequency in the 15-meter band.

Another situation where this idea would work is the case of a 10-MHz dipole, whose third harmonic resonance is well above 30 MHz. It should be possible to place loading loops at the 10-meter high voltage points in the flat-top to bring antenna resonance within the 10-meter band. Someday I'm going to try this interesting antenna. If you hear me, you'll know the scheme works!

another two-band antenna

I have an experimental license (KM2XDW) for conducting tests on 18 MHz. Although most of the time I've used a dipole antenna, I've recently experimented with half-waves in phase, as shown in fig. 3. The old reliable "two half-waves in phase" design has been used for many years; un-

fortunately such an antenna cut for 18 MHz just wouldn't fit into the restricted space in my backyard. The solution was to make the flat-top shorter and increase the length of the folded center section. This section acts somewhat in the manner of a matching transformer, allowing the antenna to be fed with a 1-to-1 balun and a coax line.

The antenna, cut to fit the space, and its dimensions are shown in the drawing. Interestingly, it was found that the antenna also exhibited resonance in the 10-meter band! By luck, the total wire length in the antenna was just about 1-1/2 wavelengths on 10 meters. The VSWR curves for 18 and 28 MHz are given in fig. 4. Using the formulas given in fig. 3, the antenna may be cut for any two frequencies that have the ratio of 1.57 to 1. Thus, an antenna cut for 14 MHz will also present a second resonance at $14 \times 1.57 = 21.98$ MHz. That's a little too high in frequency to be of practical use, but by adding butterfly loops at the high voltage points on the wire for the harmonic frequency, that resonance can be lowered to the 21-MHz band with little, if any, effect on 14-MHz operation.

general case

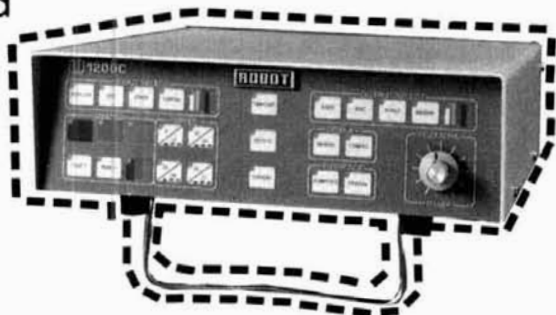
These antenna examples show that by changing the shape of a long wire, and by adding capacitance at the high voltage points at a harmonic frequency, the higher resonance point may be moved about. The idea that resonances in a long wire fall only at approximate multiples of the fundamental frequency applies only when the wire lies in a straight line. In addition, the harmonic resonant frequency determined by the configuration of the wire can be further manipulated by proper application of capacitance loops. If you want a simple multiband antenna, center-fed with a coax line, you can run your own experiments along the lines of those shown here. Armed with an SWR meter and a notebook for keeping records of your experiments, the sky's the limit!

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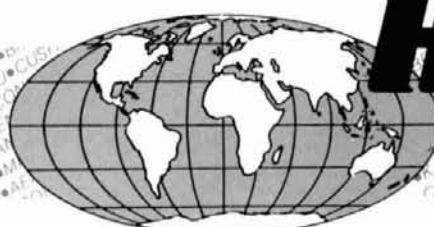
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a three-tube 4CX250B linear amplifier

Who says
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I had considered building a 2-meter linear amplifier using three 4CX250B tubes for some time. Initial calculations indicated that I could achieve good performance if I could simply translate a design on paper into an actual working unit.

To simplify the design effort I wrote a computer program that would perform precise calculations and simulations of plate and grid line operation for various operating conditions such as different plate voltages and currents, tight or loose coupling with the antenna, or reactive antenna impedances. Doing this yielded optimum Qs and efficient line values.

Many hams and clubs in Yugoslavia were interested in the design. Sketches were copied and distributed, and amplifiers built and tried. The initial design worked as planned. Modifications have since been incorporated; this article describes the best and final design. Several other published designs were checked against the computer program and found not to be optimum.

three tubes in parallel

This amplifier uses three tubes in parallel in which the screen is at both DC and RF ground potential. Many hams who have tried to build around the 4CX250B have given up because of the unavailability or cost of the sockets. Grounded screen operation offers a viable yet inexpensive alternative, provides stable operation, and requires no sockets.

Though the mechanical layout may seem primitive to the sophisticated VHF builder, it's really simple and efficient. Tube replacement is somewhat more difficult than with sockets, but it can be done quickly once you get used to it.

voltages

A schematic of the completed amplifier is shown in fig. 1.

The cathode is at -350 volts and the grid at -400 volts with respect to the screen (remember, it's grounded). To simplify the power supply design, voltage-dropping resistors (68k, 100W) and zener diodes are used. Additional protection is provided by slow-blow fuses. The plate voltage should not exceed 2.5 kilovolts unloaded and no less than 2.1 kilovolts loaded.

It's useful to include as many meters as possible — but if you choose not to, make sure that at least the plate and screen current are independently monitored.

Considering the relatively high voltage on the cathode, the filament transformer secondary should be well insulated from ground. One of the secondary leads is soldered directly to the tube pins and the other via a necessary RF choke.

RF section

The plate line is a quarter-wave section of stripline that is directly grounded at one end and connected to the tubes via a plate blocking capacitor, C_0 (see fig. 2 for details). The quality, values, and placement of these capacitors are critical and should be carefully considered. The following conditions must be satisfied:

- high capacitance value resulting in less than one ohm of capacitive reactance
- high Q
- placement close to the tubes

The blocking capacitors used in this design consist of two metal plates with a teflon (PTFE) dielectric. The lower plate is grounded (plate strip L_a) while the other is connected to the tube plates. I used teflon foil which measured between 0.3 and 0.5 mm in thickness and was perfectly smooth, with neither cracks nor slots. The upper metal sheet is extended up to the grounded side of the plate line, thereby increasing capacitance and aid-

By Dragoslav Dobričić, YU1AW, St. Supljikca 105/8. 26000 Pancevo, Yugoslavia.

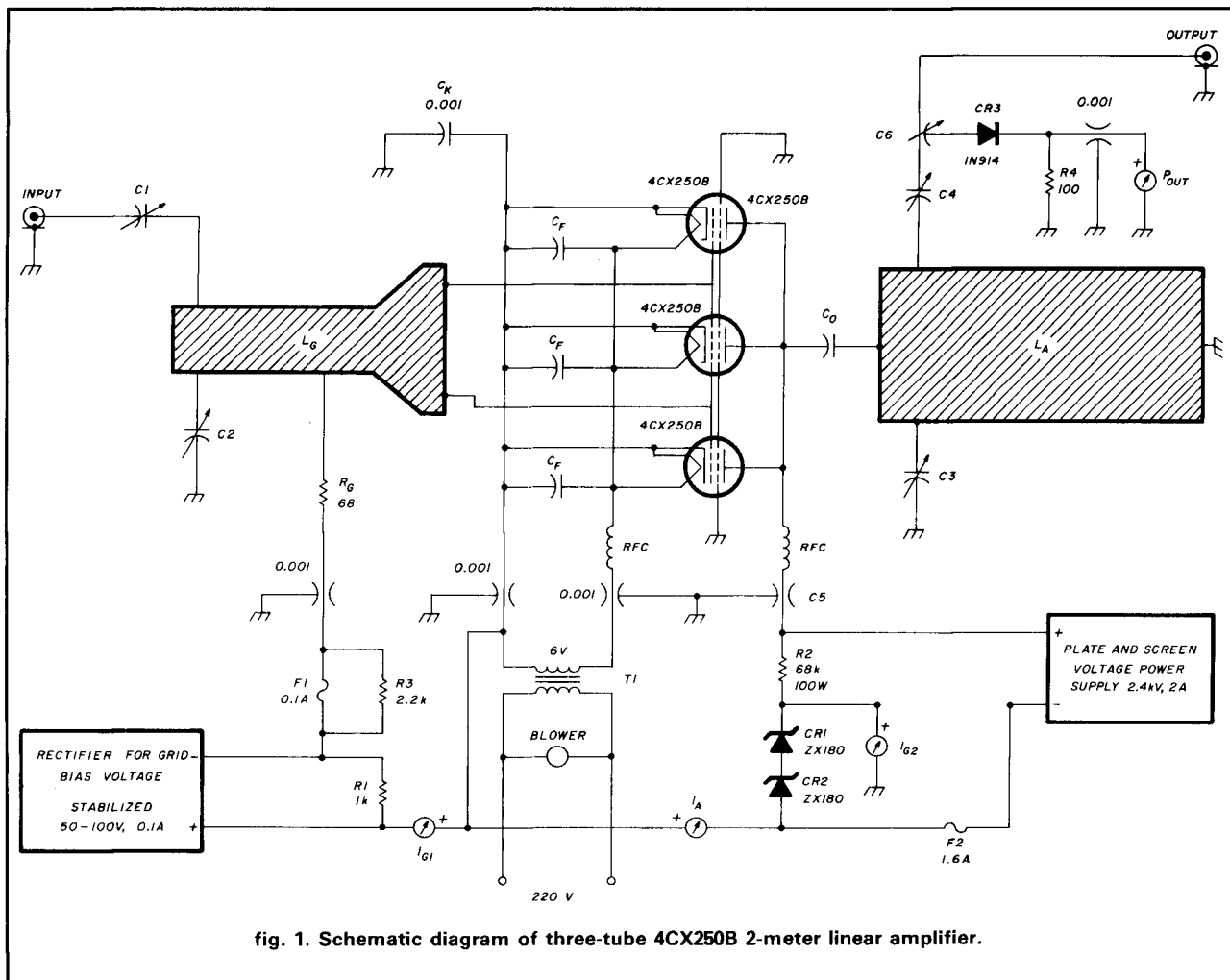


fig. 1. Schematic diagram of three-tube 4CX250B 2-meter linear amplifier.

ing in rf/dc isolation. An enlarged view is shown in **fig. 3**. Plate voltage is supplied through an RF choke to the location of lowest RF potential (almost zero). This improves amplifier stability while maintaining good RF/DC isolation.

Capacitor C5 (a section of single-sided glass-fiber PC board) is glued to the plate compartment wall and acts as a feedthrough capacitor. High voltage is introduced by means of a BNC connector. No additional steps are necessary to prevent feedback of the RF signal into the power supply.

As shown in **fig. 4** and **photo A**, there's an "ear" on both sides of the resonator, i.e., C4 and C5 stator plates. Rotor plates are soldered to the central pin of a female N connector (for C4) and grounded (for C5). Plate separation — and consequently capacitance — can be changed mechanically by adjusting a length of fishing line wrapped around a shaft made of brass or some insulated material.

The grid line is a half-wave 70-ohm impedance strip-line used instead of the quarter-wave line because input capacitances are rather high (see **fig. 5** and **photo B**).

An air-variable capacitor, C2, at the end of the line provides the resonant tuning control. Drive power is applied through C1 and bias is fed via a 68-ohm resistor at the point of minimum RF voltage. Using a resistor instead of a choke reduces the amplifier's tendency to oscillate. A resistor should always be used when high gain tetrodes are employed.

Grid connections are all soldered to the input strip through three small holes. C1 is connected to the input BNC connector with a length of coaxial cable with its plastic jacket removed (see **fig. 6**).

Cathode pins on the 4CX250Bs are 2, 4, 6, and 8. They should be connected together with a piece of copper strip the same width as the pin length. All number 3 pins should also be connected together and soldered to the cathodes. This places all cathodes and one side of all filaments at the same potential. Pin 1 is connected to the screen ring and can be used as a ground contact. Capacitor Ck should be connected between pins 1 and 2 using the shortest possible leads. Capacitor Cf should be connected between the remaining filament contact (pin 7) and pin 6 or 8. Pin 5 should not be used at all.

construction

Dimensions and details for the fabrication of the upper and lower plate, teflon spacer, grid line, and capacitors C3 and C4 are shown in **fig. 7**.

The amplifier cabinet is made from aluminum and has separate grid and plate compartments (**photo C**). Ventilation holes in the plate between the two compartments

permit air flow; air enters the grid compartment, is brought up to cool the plates and then exhausted through homemade teflon chimneys.

Though the air flow path is rather long, there is a negligible drop in air pressure along the way compared to that experienced at the plate's cooling fins. The 4CX250B tubes require a good-sized squirrel cage blower to overcome this pressure drop (**photo D**). In addi-

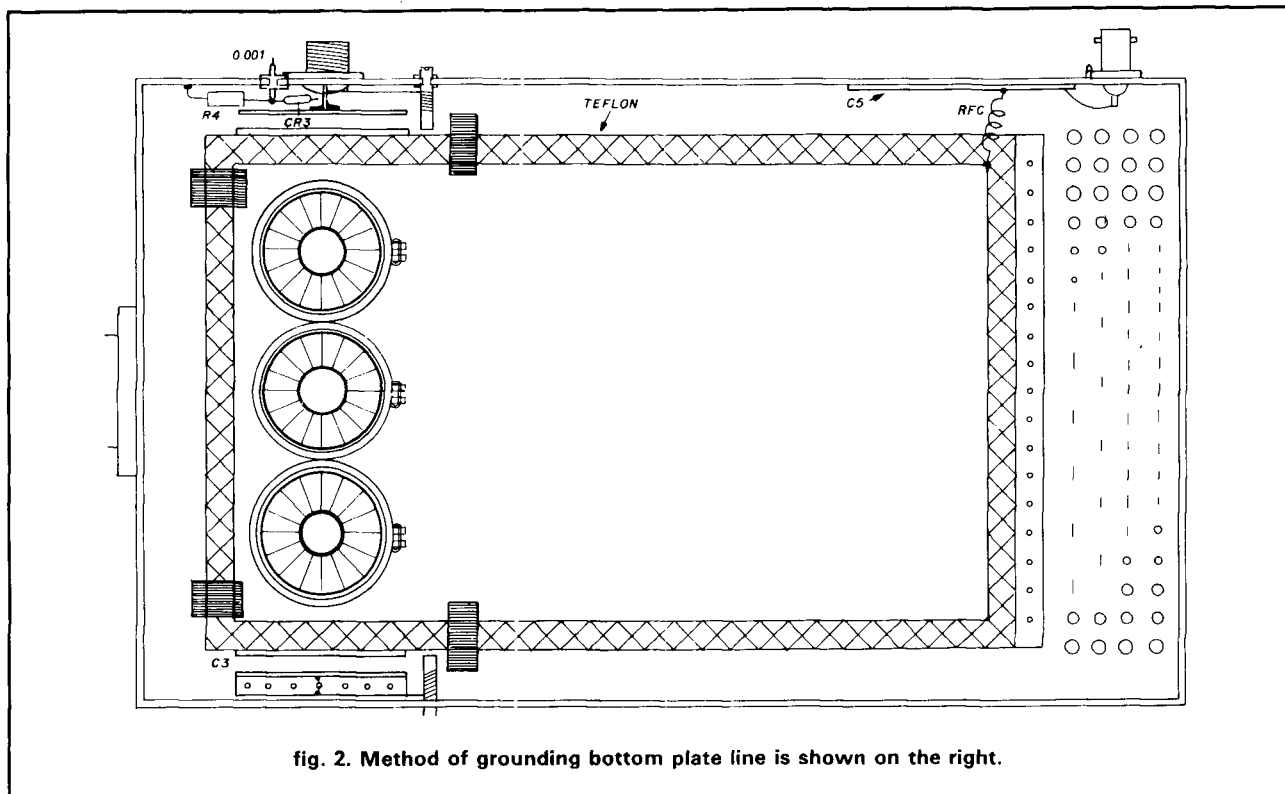


fig. 2. Method of grounding bottom plate line is shown on the right.

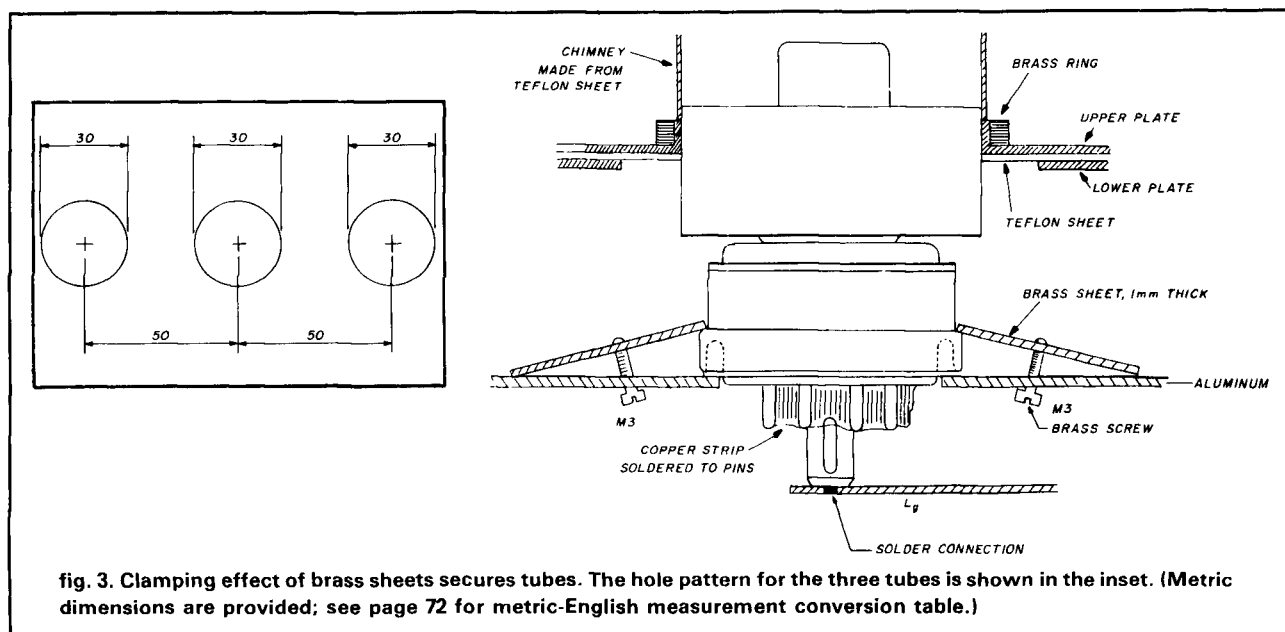
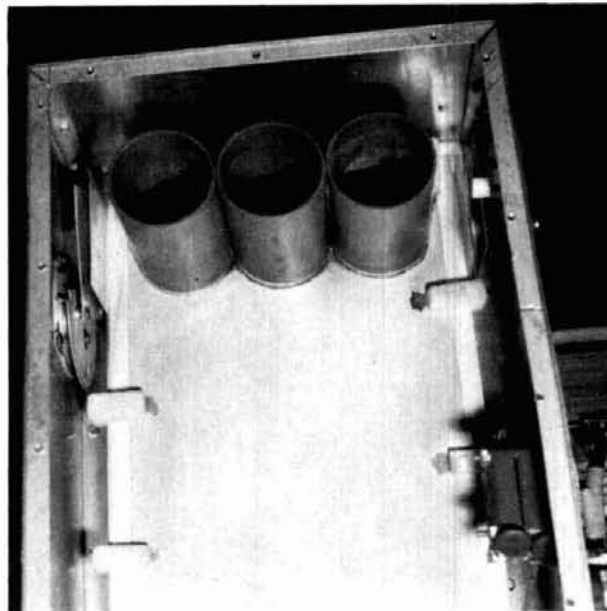
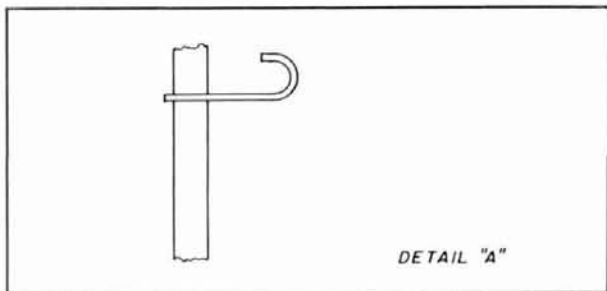
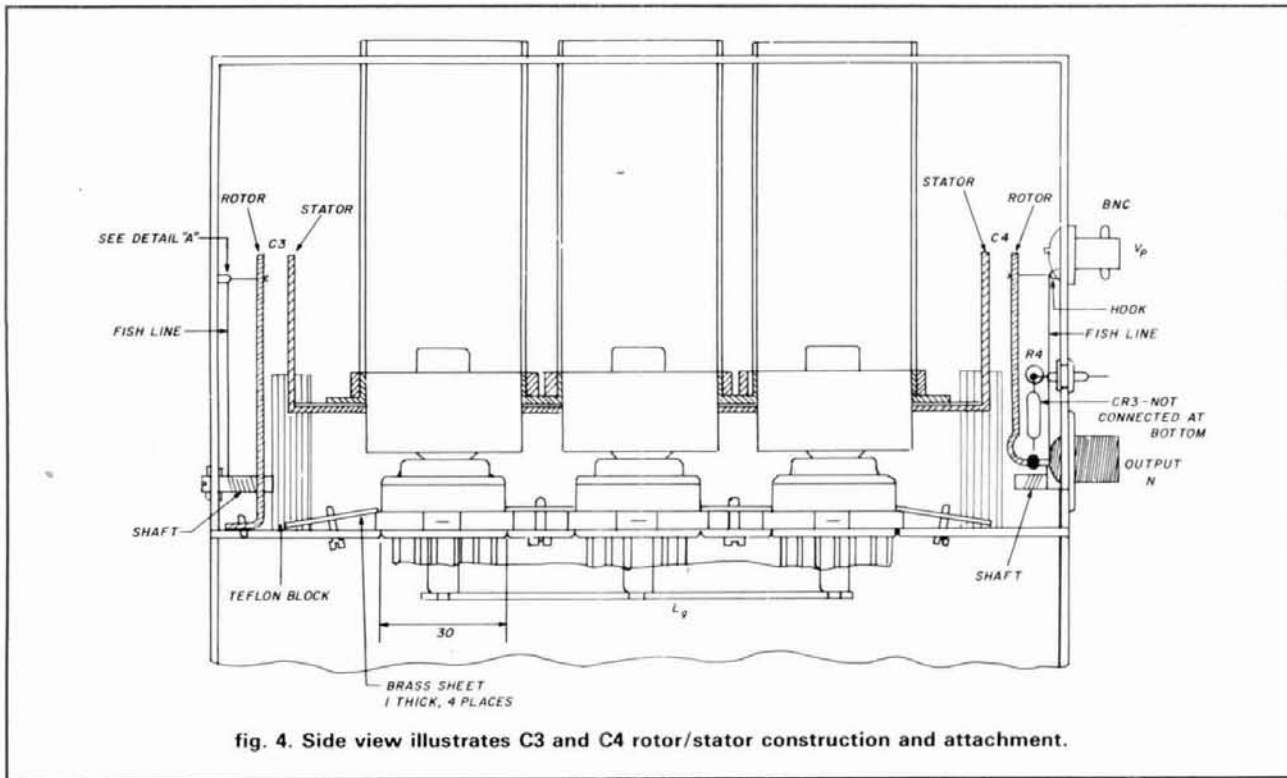


fig. 3. Clamping effect of brass sheets secures tubes. The hole pattern for the three tubes is shown in the inset. (Metric dimensions are provided; see page 72 for metric-English measurement conversion table.)



tion, good sealing between the aluminum plates is needed to prevent leaks and subsequent loss in air pressure. Holes between compartments should be round rather than slotted because elongated holes could become slot antennas, causing oscillation as a result of plate/grid coupling.

The plate line resonator (lower plate) is joined to the shield with brass screws. Good bonding is essential because of high currents at this location, so be sure to use as many screws as specified. The upper and lower parts of the plate line are held in place with slotted tefflon blocks that fit it securely to the resonator.

The tubes are mounted in the pre-drilled holes in the shield with the diameter chosen according to the screen ring dimensions. The screen ring is slightly larger than the ceramic body and good RF and DC contact is made. Each tube is secured at four locations and a single piece of brass is used to clamp adjacent tubes. To remove the

Photo A. Capacitance variation is achieved by mechanically adjusting length of fishing line.

tubes, reverse the procedure by loosening the clamps and rotating them 90 degrees. Do not apply too much pressure on the tubes because the screen ring could be damaged.

Plate chimneys are fabricated from thin tefflon sheet rolled and clamped to fit the plate radiator diameter (see

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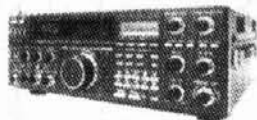
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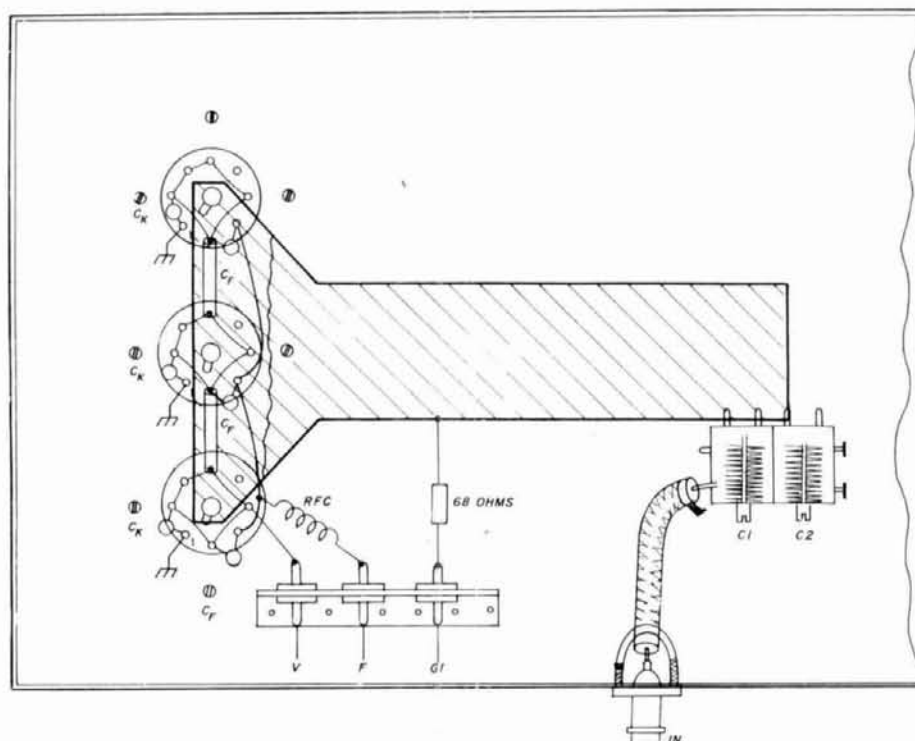


fig. 5. Half-wavelength grid line is utilized to handle high input impedances.

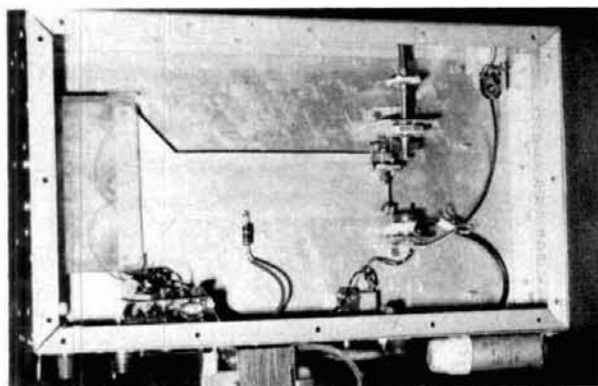


Photo B. Bottom view of grid compartment illustrates major component placement. Tube sockets are on the left.

photo E). An even less expensive solution would be to use paper heavy enough to withstand the expected temperatures. The plates can be connected to the plate resonator using finger stock or other material. The most important requirement is that good electrical contact is made.

adjustments

Switch on the amplifier in the following sequence. Turn on the filament voltage, bias voltage, and blower.

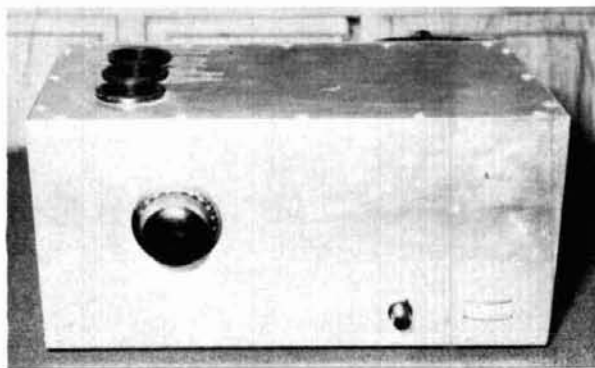
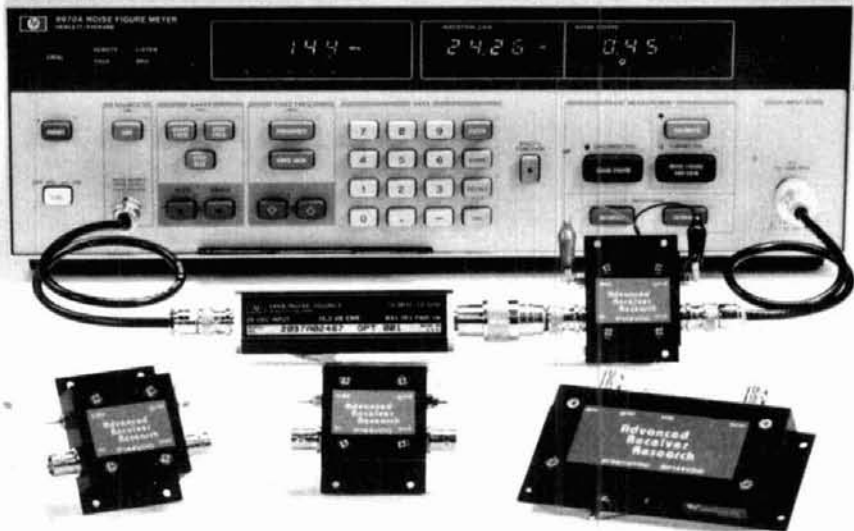


Photo C. Front view of 2-meter linear shows essentials are all there: input, tuning, and exhaust.

After a 1-minute warm-up, apply screen and plate voltage. If hash is heard in the receiver with the amplifier in standby mode (all voltages on), increase the bias voltage to -90 volts or more (negative). The bias voltage is normally adjusted for an idling plate current of 50 mA per tube.

Carefully increase the drive power while adjusting C1 and C2 for a plate current peak. Adjust C3 for a screen grid current peak, then repeat the first two steps with increased drive power. Adjust C4 for a final screen grid current of 8 mA per tube.

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P50VDG	50-54	<0.5	24	+12	GaAsFET	\$79.95
P144VD	144-148	<1.5	15	0	DGFET	\$29.95
P144VDA	144-148	<1.0	15	0	DGFET	\$37.95
P144VDG	144-148	<0.5	24	+12	GaAsFET	\$79.95
P220VD	220-225	<1.8	15	0	DGFET	\$29.95
P220VDA	220-225	<1.2	15	0	DGFET	\$37.95
P220VDG	220-225	<0.5	20	+12	GaAsFET	\$79.95
P432VD	420-450	<1.8	15	-20	Bipolar	\$32.95
P432VDA	420-450	<1.1	17	-20	Bipolar	\$49.95
P432VDG	420-450	<0.5	16	+12	GaAsFET	\$79.95

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SP28VD	28-30	<1.2	15	0	DGFET	\$59.95
SP50VD	50-54	<1.4	15	0	DGFET	\$59.95
SP50VDG	50-54	<0.55	24	+12	GaAsFET	\$109.95
SP144VD	144-148	<1.6	15	0	DGFET	\$59.95
SP144VDA	144-148	<1.1	15	0	DGFET	\$67.95
SP144VDG	144-148	<0.55	24	+12	GaAsFET	\$109.95
SP220VD	220-225	<1.9	15	0	DGFET	\$59.95
SP220VDA	220-225	<1.3	15	0	DGFET	\$67.95
SP220VDG	220-225	<0.55	20	+12	GaAsFET	\$109.95
SP432VD	420-450	<1.9	15	-20	Bipolar	\$62.95
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SP432VDG	420-450	<0.55	16	+12	GaAsFET	\$109.95

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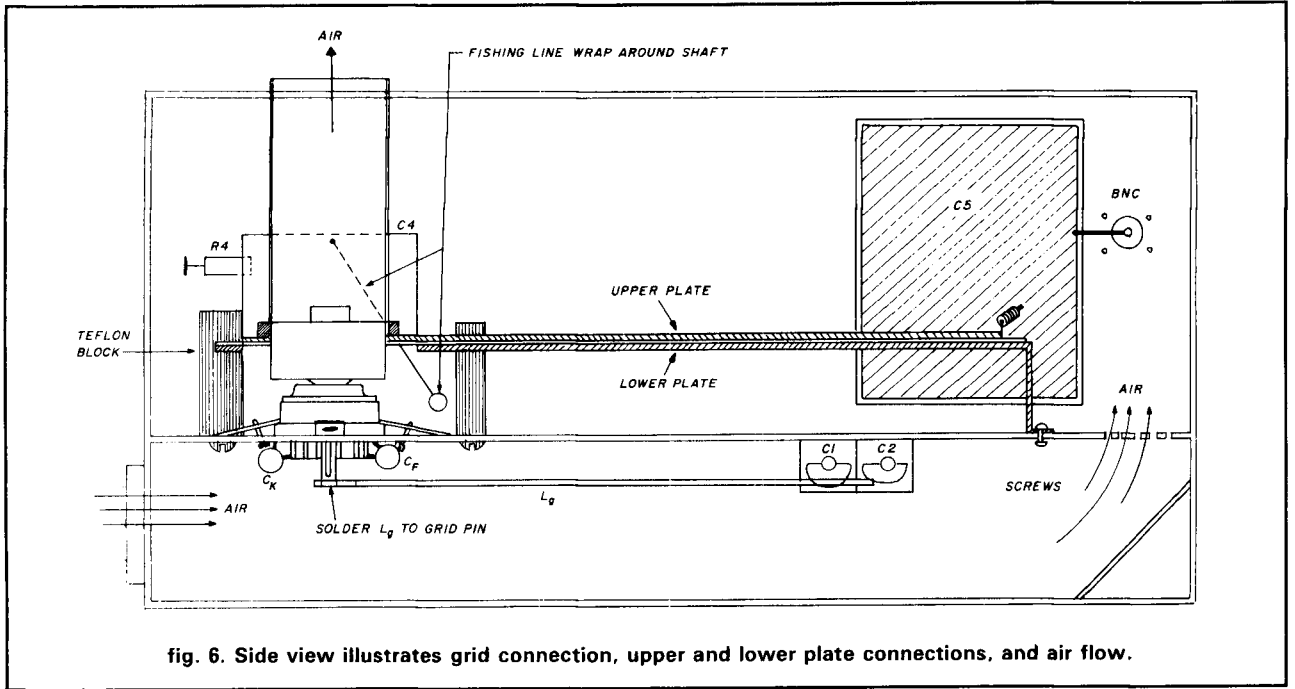


fig. 6. Side view illustrates grid connection, upper and lower plate connections, and air flow.

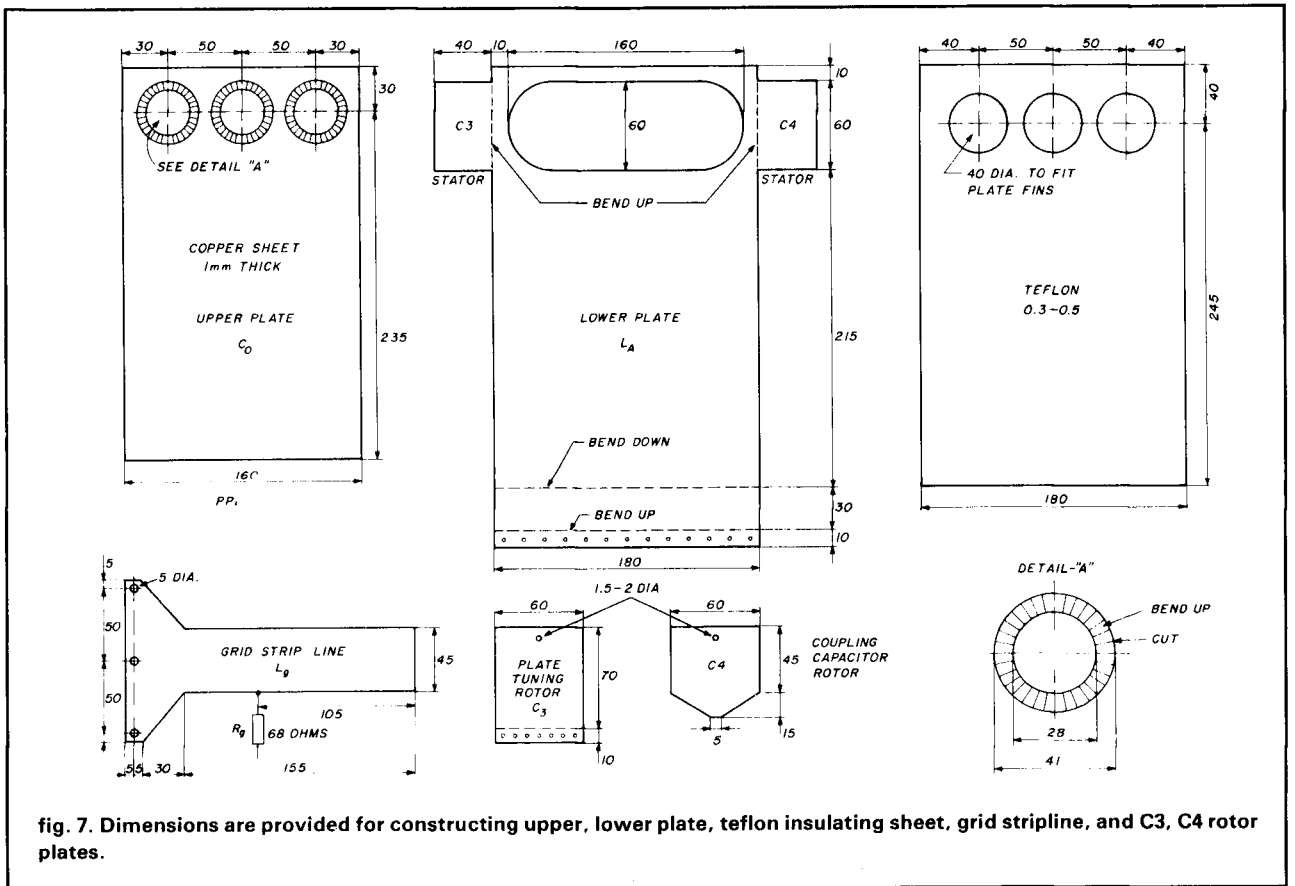


fig. 7. Dimensions are provided for constructing upper, lower plate, teflon insulating sheet, grid stripline, and C3, C4 rotor plates.

The final adjustments can be done by keying the amplifier with a series of dots (using an electronic keyer). This allows maximum drive voltages to be applied with-

out exceeding maximum power dissipation levels. All meters should indicate approximately 35 percent of maximum value. Negative screen grid current means that the

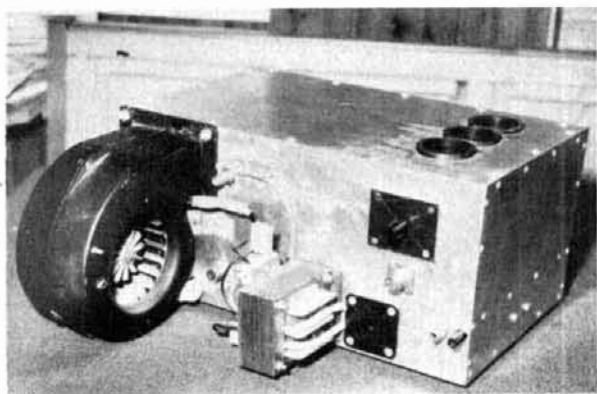


Photo D. Rear view of amplifier reveals same basic simplicity and effectiveness of construction.

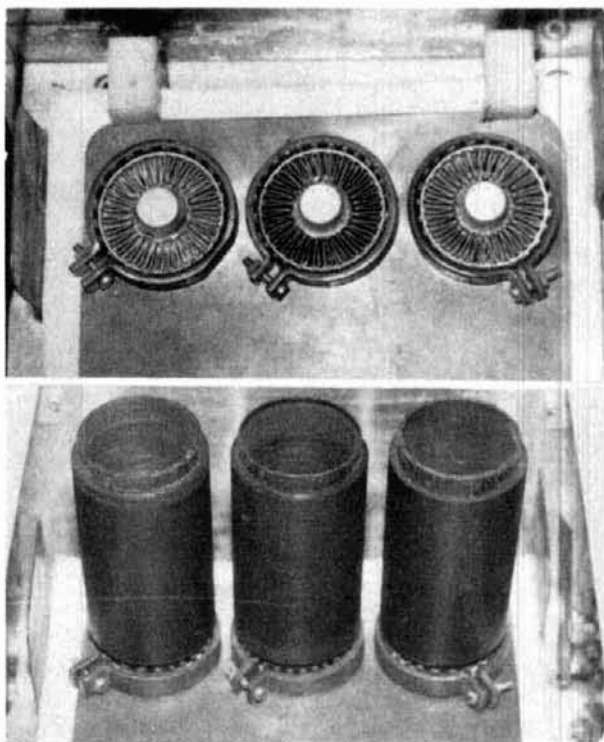


Photo E. "Before" and "after" views of chimney placement for the 4CX250B's.

drive power is too low and/or that the amplifier is too tightly coupled to the antenna.

single-tube SSB and CW ratings

At the ratings shown below, fm, a-m, RTTY, and SSTV operation are *not* recommended.

V_{g1}	= -65 volts
V_{g2}	= 350 volts
V_p	= 2.4 kV
I_{g1}	= 15 mA (max)

I_{g2}	= 8mA (max)
I_p	= 50mA (idle)
I_p	= 400mA (max)
P_{out}	= 650 watts (max)
P_{in}	= 960 watts (max)
P_{diss}	= 300 watts (max)
R_L	= 3.1 kilohms
Efficiency	= 68 percent

If all three tubes are closely matched, expect a tripling of output power and all currents. Equal power sharing (output) can be verified by measuring the temperature of air leaving the chimneys. If there appears to be an imbalance, increase idling current; the inequalities should become even more obvious.

determining component values

Computer simulation determined that the values of the following capacitors at resonance should be as follows:

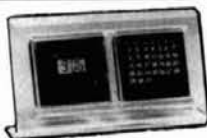
$C1 = 35 \text{ pF}$	$C3 = 3 \text{ pF}$
$C2 = 10 \text{ pF}$	$C4 = 5 \text{ pF}$

Low losses, broad bandwidth, small temperature detuning, low RF voltages on the tube (resulting in longer life and little possibility of flashovers), and broad plate tuning make this a practical design. Note that all calculations were based on Eimac tube data and differences may occur if other manufacturers' tubes are employed.

Another useful feature of this amplifier is its low driving power requirement. Low cathode stray lead inductances help to reduce the drive requirements. It's possible to achieve between 1 and 2 dB of additional gain if a special effort is made to resonate (series) the value of C_k with the stray cathode lead inductances. It actually becomes necessary if you are drive-power limited.

additional hints

- Use as many screws as possible to join the cabinet together in order to reduce rf leakage.
- Use a type N — not an SO239 — for the output connector.
- Plate voltage may be supplied through a BNC connector and RG-58.
- Any departures from the dimensions called out on the drawings could result in degraded amplifier performance.
- The plate line should be cut from brass or copper sheet 1 mm thick and should be silver-plated, if possible.
- Both of the ZX180 zener diodes (180 volts, 12 watts) are mounted on heat dissipation devices.
- Other combinations of zener diodes or tube voltage regulators may be used to provide a regulated +360 volts.
- Resistor R3 (2.2k) acts like a fuse, delivering bias if the "slowblow" 0.1 A fuse is blown.



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millimeters	inches
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0.5	.019
1.0	.039
1.5	.059
2.0	.079
5	.197
10	.394
15	.591
28	1.102
30	1.181
40	1.575
41	1.614
45	1.772
50	1.969
60	2.362
70	2.756
105	4.134
155	6.102
160	6.299
180	7.087
215	8.465
235	9.252
245	9.646

acknowledgments

I wish to thank all those hams who believed in this unusual design and had the courage to try it.

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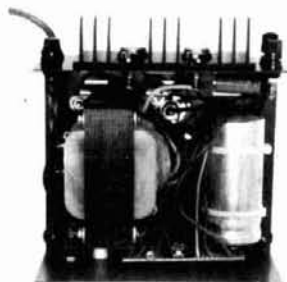


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RS-4A	3	4	3 3/4 x 6 1/2 x 9	5
RS-7A	5	7	3 3/4 x 6 1/2 x 9	9
RS-7B	5	7	4 x 7 1/2 x 10 3/4	10
RS-10A	7.5	10	4 x 7 1/2 x 10 3/4	11
RS-12A	9	12	4 1/2 x 8 x 9	13
RS-20A	16	20	5 x 9 x 10 1/2	18
RS-35A	25	35	5 x 11 x 11	27
RS-50A	37	50	6 x 13 3/4 x 11	46

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• Separate Volt and Amp Meters RS-20M	16	20	5 x 9 x 10 1/2	18
RS-35M	25	35	5 x 11 x 11	27
RS-50M	37	50	6 x 13 3/4 x 11	46

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VS-20M	16 9 4	20	5 x 9 x 10 1/2	20
VS-35M	25 15 7	35	5 x 11 x 11	29
VS-50M	37 22 10	50	6 x 13 3/4 x 11	46

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MODEL RS-12S

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MODEL	Continuous Duty (Amps)	ICS* Amps	Size (IN) H x W x D	Shipping Wt (lbs)
RS-7S	5	7	4 x 7 1/2 x 10 3/4	10
RS-10S	7.5	10	4 x 7 1/2 x 10 3/4	12
RS-10L(For LTR)	7.5	10	4 - 9 - 13	13
RS-12S	9	12	4 1/2 x 8 x 9	13
RS-20S	16	20	5 x 9 x 10 1/2	18

VHF/UHF WORLD

Joe Reiser
W1JK

33 cm: update

One year ago this column was devoted to our newest UHF band, 33 cm (902-928 MHz).¹ Since that time, much has happened on this band, which is now just over 18 months old. Activity is growing fast, but more activity could really help popularize the band.

Last year's column was intended to be an "entry level" article. In this month's column, I'll present further information that will upgrade some of the circuits and the antenna described in the earlier article. I'll also discuss some commercial gear that has recently become available. With luck, the new information will encourage increased activity by helping others to become active on one of our newest and most exciting UHF bands.

latest activity

Weak-signal operation on 33 cm has been reported in the W1, W2, W3, W4, W7, W9, W0, VE2, and VE3 call areas. (Did I miss anyone?) Most of the activity is reportedly on the calling frequency of 903.1 MHz, as recommended in last year's column. There are also several fm repeaters active in the upper portion of the band, two of which are in the W1 call area. A California and a Massachusetts ATV repeater have also been reported as active, so the band is obviously being used for purposes other than weak-signal operation.

As a sign of acceptance of 33-cm contacts, the ARRL now counts 33-cm contacts separately for new multipliers and points in VHF/UHF contests. 33 cm also has its own VUCC (VHF/UHF Century Club) award for those who've worked at least 25 grid squares on this

band. Furthermore, the ARRL has announced that the Spring Sprint contests this year will include a separate 33-cm contest night. The tentative date is Friday, May 8, 1987 (7 to 11 PM, local time).

Several stations have also been active on portable expeditions. They've done quite well using low power (10 watts) and single 12-foot loop Yagi antennas. I've been successful in working only a few grids that way, so my total number of grids worked so far is a puny ten. However, while writing this month's column I heard that at least one ham has sent out several active "rovers" and worked the necessary 25 grids to claim the first 33-cm VUCC award.

propagation

Although initially some of us weren't sure what kind of propagation we'd experience on 33 cm, we expected it to be a blend of 70 (432 MHz) and 23 cm (1296 MHz). I think most of us now active on the 33-cm band feel that the typical propagation is more like 70 cm for casual contacts. Stations using 10 watts and a single 12-foot boom loop Yagi can easily cover a 100-mile range on SSB, with signal strengths as strong or stronger than can be anticipated on 70 cm.

Extended tropospheric propagation hasn't been well used because of the poor distribution of equipment available for 33 cm during the few openings this past year. The aurora, though absent in recent months, should return within the next year or two as solar cycle 22 begins. This will afford the best opportunity for auroral QSOs above 450 MHz, presently the highest frequency at which two-way Amateur

QSOs have been reported in this mode.

Aircraft scatter propagation on 33 cm seems to be as good or better than on 70 cm and more like the conditions experienced on 23 cm.^{2,3} Several stations have reported QSOs showing all the signs of aircraft scatter — a sudden appearance of a signal, reasonable signal strength, some flutter, then a gradual decrease in signal strength. Aircraft scatter is possible to about 500 miles, but 200 to 350 miles is near optimum.

While I haven't received any reports of activity, this band is a natural for 33-cm EME, since small-diameter (12 foot) dish antennas should be sufficient for such communications. The lack of EME operation on 33 cm is probably attributable to the fact that few, if any, Amateurs are using high power (greater than 150 watts) amplifiers on the band. I'm sure this QRP situation won't last very long!

As you know, I like to keep track of DX records on the frequencies above 50 MHz. At the present time the best DX reported on 33 cm is a mere 377 miles (606 km).⁴ Let me know if you break this record. Once again, I believe this record will be broken before this column is in print. What a great challenge we have on a new band!

33-cm antennas

As pointed out in last year's column, 33 cm is a transitional band for antenna design. Consequently, the loop Yagi is the most popular antenna type; it's easy to construct and has proved fairly successful.

Reference 1 included a description of a 33-element loop Yagi design on a 12-foot boom. This particular design is

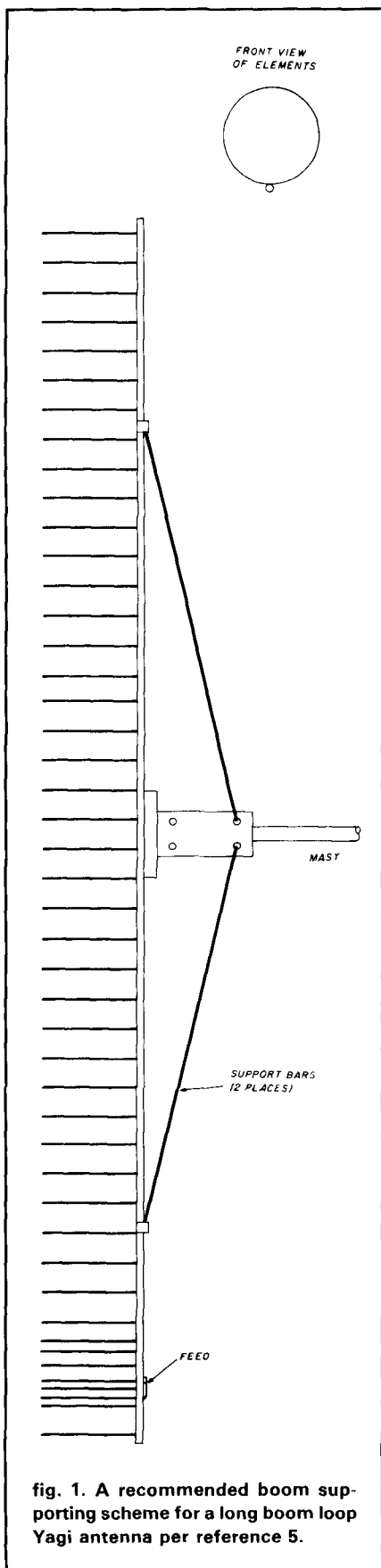


fig. 1. A recommended boom supporting scheme for a long boom loop Yagi antenna per reference 5.

probably not fully optimized, since it couldn't be scaled from any of the original designs shown in reference 5. However, it does have moderate gain (greater than 18 dBi) and a reasonably clean radiation pattern, as evidenced by the fact that most of the active 33-cm stations are using this or a similar design. Down East Microwave* sells a 33-element, 33-cm loop Yagi, but I don't know whether it's this or another design.

After a bit of head scratching on the subject of a simple approach for a higher gain antenna design, and a needle from Sam, W2PGC, I decided to build a 45-element loop Yagi scaled from the original 45-element, 23-cm optimized design described in reference 5. This would require a boom length of 17 feet, 2 inches.

After more thought about available materials and tubing sizes, I decided to use a 17-foot, 6-inch boom. The first part of the boom is the same length as the original design (12 feet) with the same diameter tubing (1 inch). A 6-foot length of 7/8-inch tubing is then inserted 6 inches into the director end of the boom for an overall length of 17 feet, 6 inches.

With this longer boom length and smaller diameter tubing on the front end, the overall strength of the boom will be decreased. A larger diameter boom could be used, but that would be more expensive and increase wind load, which is already high on a loop Yagi design. Therefore, I recommend a boom supporting structure similar to the one described in reference 5. A simplified sketch of it is shown in fig. 1.

Since there was a little extra boom available with a 5-foot, 6-inch extension, I added another director, making the new design one of 46, rather than 45, elements. The final design is shown in fig. 2; the driven element details are shown in fig. 3. This longer boom design uses the same element materials as the older 33-element design, with a slightly different element length taper schedule.

The gain of the 46-element loop Yagi should be about 20.5 dBi, a big

improvement over the 33-element design. The beamwidth is approximately 15 and 16 degrees in the E and H planes, respectively. Therefore, the recommended stacking distances per references 5 and 6 are 37 and 35 inches in the E and H planes, respectively.

Finally, the original 33-element loop Yagi described in reference 1 was designed for 902 MHz. As a purist, I thought that the new longer boom model should be moved slightly higher in frequency. Hence the spacings and element lengths on the 46-element loop Yagi differ slightly from those on the original 33-element design.

If anyone wants to extend the original 33-element design, I don't think it will matter much if the old spacing is maintained. All that will be necessary is to add the extra boom section and directors, placing them each 5.115 inches further out from the preceding director. Don't forget to always reference element spacings from reflector No. 2 in order to keep any tolerance buildup to a minimum. Also note that some of the director lengths of the original design should be changed as shown in fig. 2 if optimum performance is expected.

Several Amateurs are developing experimental long boom (12 feet or longer) Yagi designs, but they'll need to wrestle with the problems of boom corrections and impedance matching — a formidable problem on the UHF bands. I'll let you know how successful I am with mine!

Tonna Antennes (F9FT) has developed a short (8 foot, 4 inch) boom, 23-element Yagi. Marketed by The "PX" Shop,** it has a specified gain of 18.2 dBi with a 21- by 22-degree E and H plane beamwidth, respectively. Tonna Antennes circumvented the problem of boom corrections by mounting the elements above the boom on specially designed standoff insulators similar to those on their 23-cm designs. The feed system uses a folded dipole.

Needless to say, there will be some

*Down East Microwave (W3HQT), Box 1655A, RFD No. 1, Burnham, Maine 04922.

**The "PX" Shop (KC2PX), 52 Stonewyck Drive, Belle Meade, New Jersey 08502.

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The PK-232 uses an eight-pole bandpass filter followed by a limiter discriminator with automatic threshold correction. The internal modem automatically selects the filter parameters, CW Fc = 800 Hz, BW = 200 Hz; HF Fc = 2210 Hz, BW = 450 Hz; VHF Fc = 1700 Hz, BW = 2600 Hz.

The PK-64 uses on screen indicators to show status, mode, and DCD (Data Carrier Detect) while the PK-232 uses front panel indicators. Both units use discriminator style tuning for HF operation. And that's just the tip of the iceberg. Features like multiple connects on packet, hardware HDLC, CW speed tracking, and other standard AEA software features are included in both the PK-64 and PK-232.

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more churning going on in 33-cm antenna design in the foreseeable future. Older designs will be improved and new designs will be forthcoming. Only time will tell whether the present antenna approaches are optimum for the 33-cm band.

circuit update

The converter/transverter designs in reference 1 have worked out quite well for many of us on the 33-cm band. Note that a few errors appeared in some of the figure captions in reference 1: the conversion loss of the transmit type upconverter shown in **fig. 3** should be 16 dB (not 9 dB), and the gain of the medium power amplifier in **fig. 12** should be 20 dB (not 30 dB).

Some additional comment on the hybrid modules are in order. Since reference 1 was published, the USA distributor for Toshiba has discontinued importing the SAU11 and SAU15 hybrid modules. However, I've been advised by Hiro Shiozawa, JA0JCJ, that he can provide some of the modules — particularly the Toshiba units — directly from Japan. I'd suggest that you write to him to make further arrangements.*

There are several other sources of suitable hybrid modules. The NEC MC-5809 driver module mentioned in reference 1 works just as well as the Toshiba SAU15 as a 100-milliwatt linear amplifier. Although I haven't tried it, the NEC MC-5843 shown in **fig. 12** of reference 1 should work as well as the Toshiba SAU11, albeit at a higher price.

Other substitutes for the high power module with 7 to 8 watts minimum output power (such as the NEC MC-5828, MC-5829, and MC-5842) are also available. Each has a different gain and output power. Of particular interest is the MC-5828, which sports an output power of 8 watts *minimum*, with only 1-milliwatt input power — all for \$42.50 in single quantities! The NEC distributor in the USA is California Eastern Labs.**

Although I have no specific part numbers or additional information, I

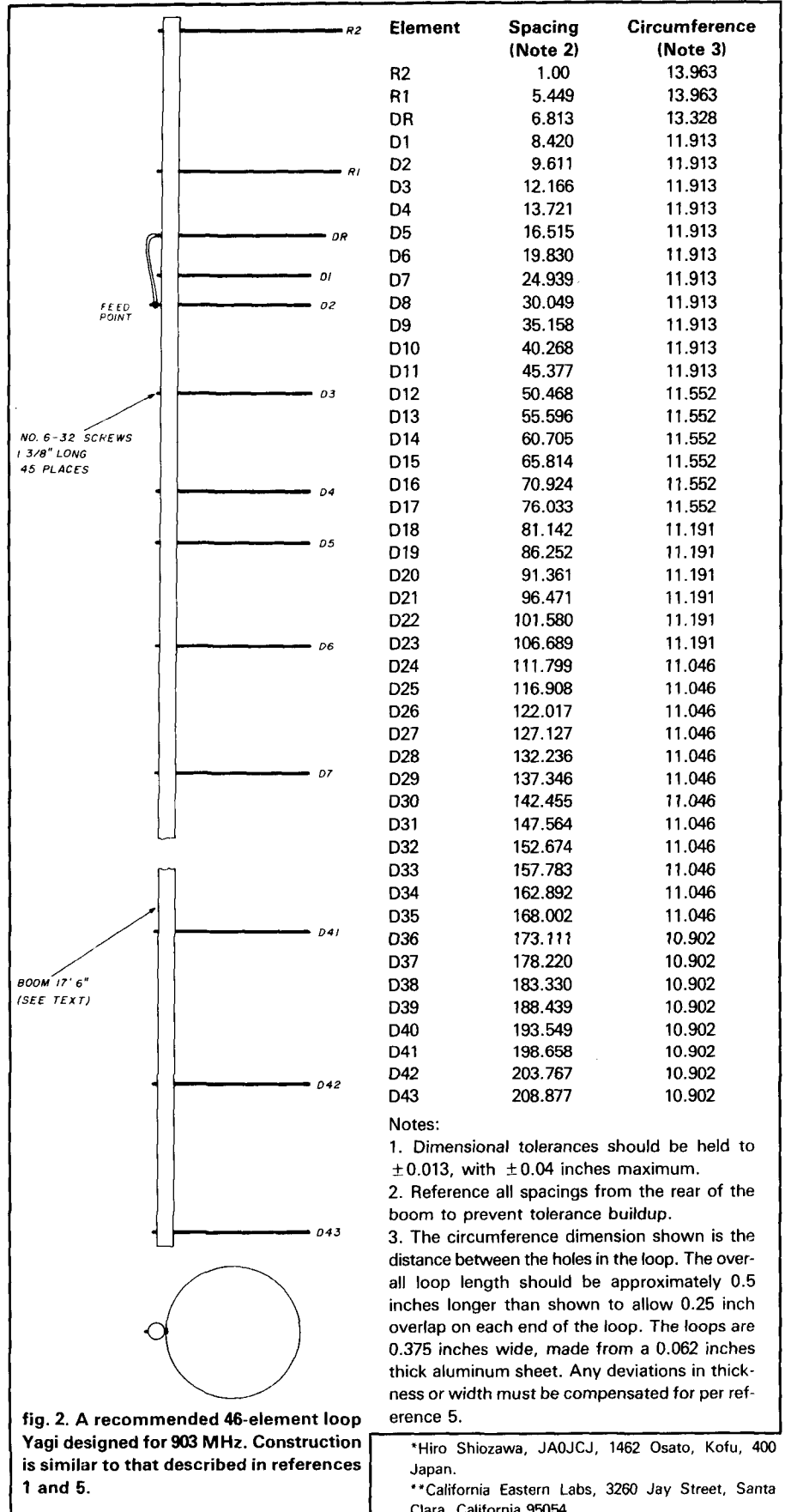


fig. 2. A recommended 46-element loop Yagi designed for 903 MHz. Construction is similar to that described in references 1 and 5.

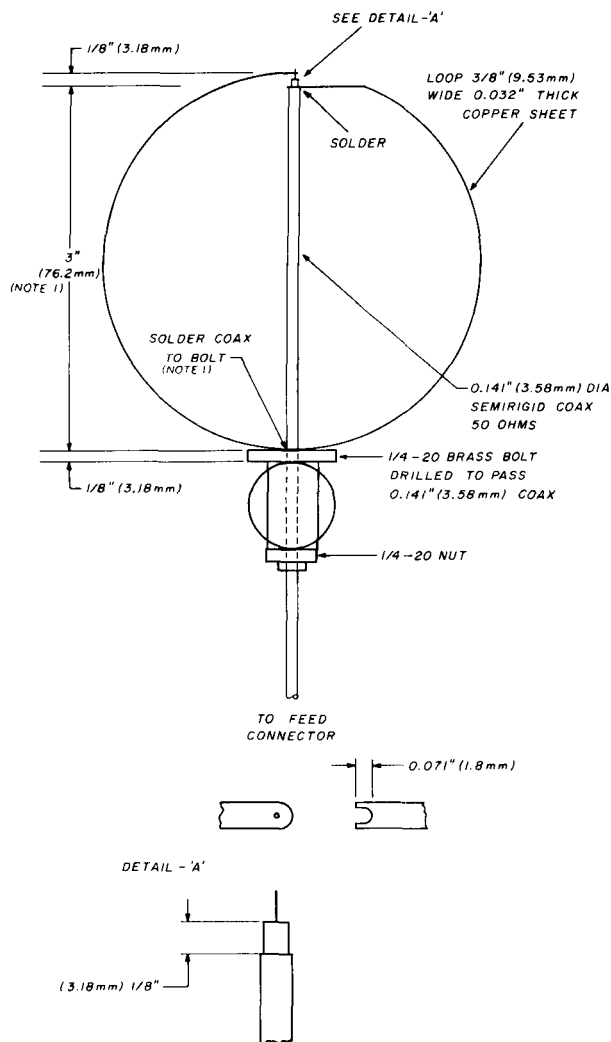


fig. 3. The feed system for the 46-element loop Yagi shown in fig. 2. Before soldering loop, adjust height (and length, if necessary) of loop for best VSWR.

know that Motorola also has a line of 33-cm hybrid power modules. The power levels are in the same class as the ones already mentioned, so if you're interested in using this type of module, I'd suggest that you also check the Motorola product line.

SSB with class-C amplifiers

The hybrid modules just described all operate in class C. In the past I speculated that you could run SSB through a class C amplifier with an acceptable IMD if certain parameters were taken into consideration. This is particularly advantageous when using solid-state "bricks" or hybrid modules,

because previous attempts to rework them for class B bias usually resulted in thermal runaway.⁷

Since then I've verified with on-the-air tests that this is indeed possible. Basically the trick I use is to rf-bias rather than dc-bias the class C module by applying a small amount of carrier to an SSB signal. The optimum seems to be about 1 watt of output from the 10-watt class C module with no modulation applied.

I have an older phasing type of SSB exciter, so I just unbalance the carrier until the output of the module is about 1 watt without modulation. I then operate SSB normally. This procedure will probably work with other class C

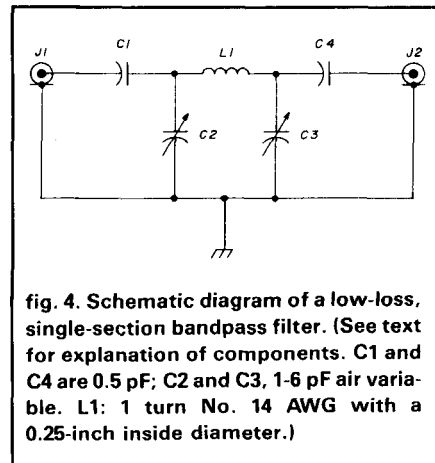


fig. 4. Schematic diagram of a low-loss, single-section bandpass filter. (See text for explanation of components. C1 and C4 are 0.5 pF; C2 and C3, 1-6 pF air variable. L1: 1 turn No. 14 AWG with a 0.25-inch inside diameter.)

circuits and devices if the rf bias is held to about 10 percent of the maximum output power. One precaution: don't transmit on SSB using this technique for long periods of time unless you have an adequate heat sink attached to the hybrid module. You could destroy it!

transverters

The circuits described in reference 1 can be easily assembled for transverter operation with a common local oscillator and power splitter as described. As proposed, the modular approach allows quick insertion of new or improved designs as they become available. This approach has worked out well for homebrewers like me.

Recently SSB Electronics introduced the model LT33S transverter. Available from Transverters Unlimited,* it operates at 903 MHz with a 144-MHz i-f and sports a low-noise GaAsFET preamplifier and 6 to 10 watts of transmitter output power. This particular unit is quite popular with 33-cm enthusiasts.

linear amplifiers

As suggested in reference 1, there are many choices of bipolar power transistors for solid-state linear power up to about 20 watts. Unfortunately, time and space won't permit me to discuss new solid-state linears too deeply at this time, but I will make a few suggestions.

*Transverters Unlimited (VE3CRU), Box 6286 Station A, Toronto, Ontario, Canada M5W 1P3.

The NEC NE0804 (5 watts), the NE0810 (11 watts), and the Thompson-CSF (formerly Solid State Microwave) SD1418 common emitter power transistors are recommended. They should work well using a circuit similar to that used on 23 cm by WB5LUA.⁸ All that's required for 33-cm operation is to lengthen the input and output lines and possibly change the value of the chip capacitor that shunts the base and collector of the transistor to ground.

Some medium power (40 to 100 watt) solid-state amplifiers are now being used. Typically they run class C with grounded base transistors. WA3JUF has proposed such a circuit using the Thompson CSF SD1414.⁹

high power amplifiers

For high power, tubes are strongly recommended. For power levels up to 100 watts, the ubiquitous 2C39/7289 is highly recommended. These tubes are plentiful and inexpensive.

W1RIL and others have modified the 70-cm 2C39/YD1050 amplifier described in *The UHF Compendium* to work on 33 cm.¹⁰ They shortened the half-wave plate line to 78 mm (as opposed to 175 mm), shortened the output coupling link, modified the input matching network as required, and obtained 50 to 100 watts output with reasonable gain.

A 3CU400/800 flat cavity amplifier design is recommended for medium power levels (300 to 500 watts).¹¹ Quarter- and half-wave cavity amplifiers using the 7650/7651 or the larger 7213/7214 tubes are highly recommended. They work well on these frequencies, can deliver high power, and are often seen at flea markets. I'm sure that many surplus or UHF TV "pulls" are also available. Stripline amplifiers will probably still work adequately at these frequencies.

filters

The filters shown in reference 1 work well. However, they are entry level filters with fairly wide bandwidth. The input filter is particularly wideband but more than sufficient to protect the input of the MRF901 preamplifier

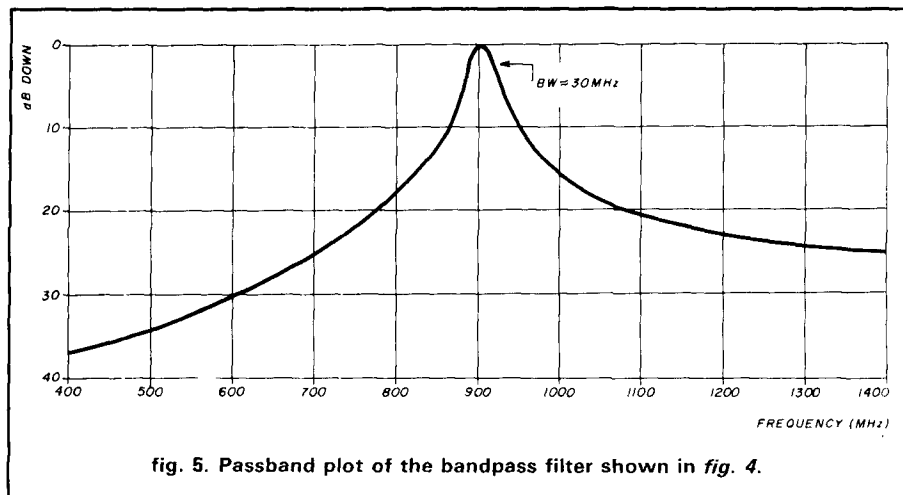


fig. 5. Passband plot of the bandpass filter shown in fig. 4.

design (shown in reference 1) if there are no local high power rf emitters such as UHF TV stations nearby.

Recently I developed an improved input filter with narrower half-power bandwidth — 30 versus 100 MHz — and less than 0.5 dB insertion loss. In addition, the 20-dB down points are only 300 MHz apart, as contrasted with the earlier design of 1000 MHz. It's similar to the design approach used in my 2-meter high dynamic range converter.¹²

A schematic of this single-section filter is shown in fig 4. C1 and C4 can be realized by connecting two 1.0-pF capacitors in series. C2 and C3 should be the piston or air-variable type if minimum insertion loss is desired. For lowest loss and best isolation, this filter should be built in its own shielded box (such as a Pomona Electronics 2417, 2428, or equivalent).

This filter is best tuned by adjusting C2 and C3 for minimum VSWR using a low power signal source. These capacitors will interact with each other, so alternate tuning until the VSWR is at a minimum. A typical passband plot of this filter is shown in fig. 5.

Likewise, the two-section bandpass filter shown in reference 1 is also an entry level filter. It has a half-power bandwidth of 50 MHz, which is a little too wide if a 28-MHz i-f is used. A slightly more complex three-section filter with a narrower half-power bandwidth, 36 MHz, and better skirt selec-

tivity of 100 MHz, rather than 350 MHz (30-dB down points), has been developed. A schematic is shown in fig. 6.

This filter is of the combline type with the input and output connections made directly to the resonators. The coupling is set by the spacing between resonators, the height of the filter enclosure, and the input/output tap position.

As with the single-section filter just described, the tuning capacitors should be of the low-loss piston or air-variable type. The resonators should be spaced 1 inch center-to-center and mounted midway between the top and bottom of the enclosure. The filter is built in a small shielded box (with 4 x 2 x 1-inch outside dimensions) such as the Hammond 1590L or equivalent. The most important parameter of the shielded box is the height inside the enclosure, which is about 0.8 inches. This filter can be tuned for minimum loss, but for best performance a sweep-tuned setup is required. A typical passband plot is shown in fig. 7.

preamplifiers

The MRF901 preamplifier in reference 1 has performed quite well. Until recently it has been my only preamplifier. Though its noise figure is too high (typically 3 dB) for serious weak-signal operation it makes an excellent second-stage postamplifier.

For a really low noise figure, use a GaAsFET preamplifier. I've recently

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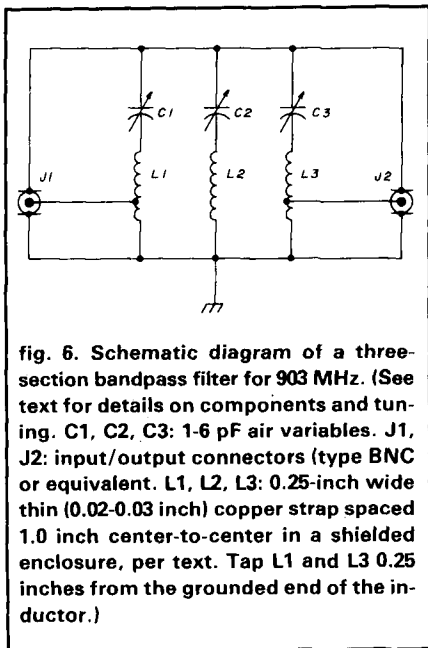
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4:1 L-network transformer that worked well for me on the VHF bands and is relatively broadband with low loss. As a result, the output compression point is very high; this is a performance parameter some of the other matching schemes I've used didn't have.

To align the preamplifier properly, a good noise figure generator is recommended. However, adequate performance can be obtained by using a weak-signal source for tuning. After tuning C1 and C2 for best sensitivity, spread or compress the turns on L2 for maximum gain, typically 13 to 15 dB. Overall noise figure is a function of the construction and GaAsFET used, but 0.5 to 1 dB is typical. The device type isn't critical; the MGF1402, MGF1301, NE72019, or equivalent GaAsFETs should all work well.

to cover from 25 to 1000 MHz and from 1030 to above 1300 MHz, offers excellent frequency stability and weak-signal characteristics on the 33- and 23-cm bands.

commercial fm gear

As mentioned in reference 1, a portion of the 33-cm band is a citizens band in Japan. As a result, many low power (5 watt) fm transceivers are available in Japan, but not yet in the USA, although that situation could change any day now.

Table 1 shows a few of these transceivers. They're all citizens band type with typically 75 to 100 fixed channels between 903 and 905 MHz. These transceivers have F3E emission and a nominal 5 watts output power. They're very small and excellent for portable operation.

One channel on these transceivers is at 903.1125 MHz. W1XX and others have been using this channel at portable locations because it's close to the weak-signal calling frequency. However, if these transceivers become available in the USA, I hope that they'll be used only above 904 MHz so as not to interfere with weak-signal operations near 903.1 MHz.

summary

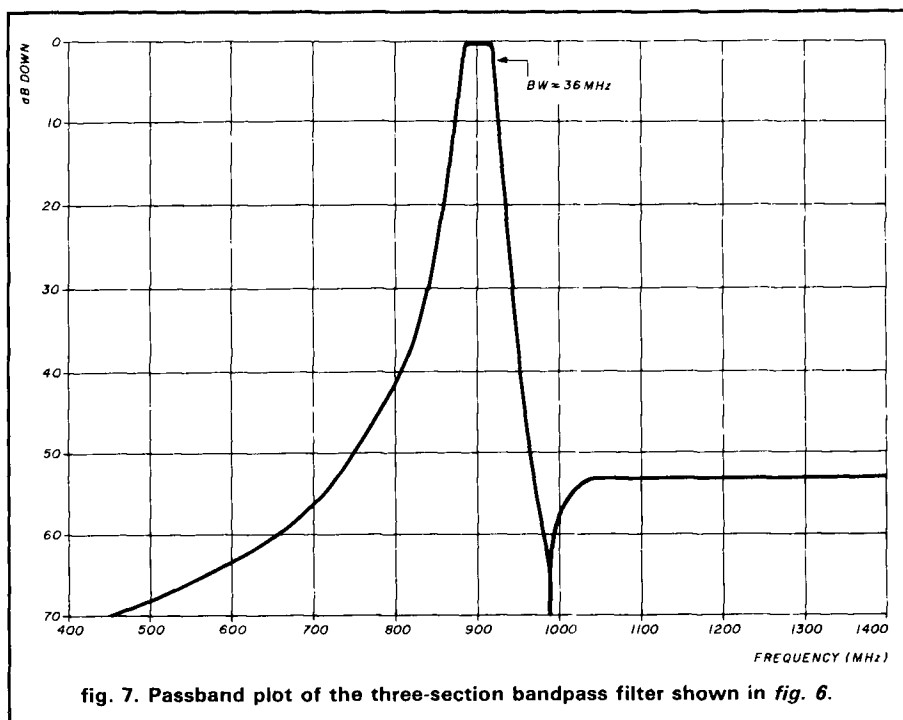
This month's column reviewed the current status of operation on the 33-cm band. New higher performance circuits and some equipment, both homebrewed and commercial, were recommended. Improvements to the circuits recommended in reference 1 were also discussed. I hope this information will stimulate increased activity. . . see you on 903.1 MHz, especially on Friday evenings — a recommended activity night — at 9 PM (local time)!

acknowledgments

I'd like to thank Bart Jahnke, KB9NM, of the ARRL for his help in locating the information on the Japanese fm citizens band gear.

SMIRK lives

Ray Clark, K5ZMS, president of SMIRK (Six Meter International Radio Klub) has recently informed me that



built one similar to the design shown in reference 13. A schematic is provided in fig. 8. The input circuit, a tuned tank, is easy to construct and provides a reasonable amount of selectivity, reducing the susceptibility to overload from high power emitters in the VHF range and below.

The output matching circuit uses a

receivers

Wide-frequency coverage receivers (such as Yaesu's FRG9600) are now available. Not necessarily weak-signal types, they have noise figures typically in the 6 to 10 dB range; a low-noise preamplifier ahead of the receiver is recommended for weak-signal operation. The new ICOM R-7000, designed

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The 1987 Callbook Supplement is a new idea in Callbook updates; it lists the activity in both the North American and International Callbooks. Published June 1, 1987, this Supplement will include all the new licenses, address changes, and call sign changes for the preceding 6 months.

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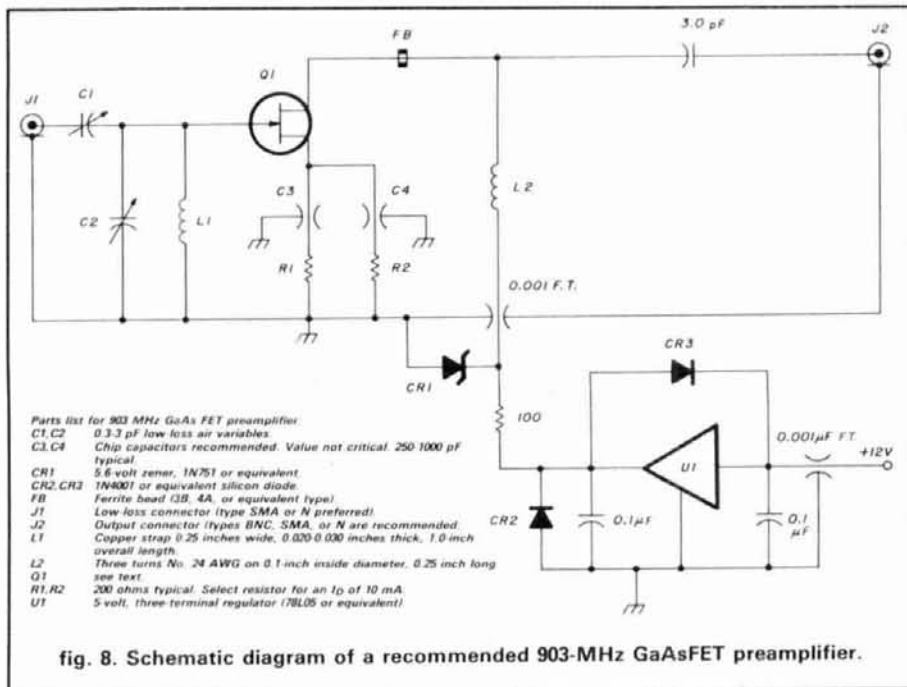


fig. 8. Schematic diagram of a recommended 903-MHz GaAsFET preamplifier.

"SMIRK is still alive and kicking." However, because of budgetary constraints, copies of the quarterly newsletter won't be sent to anyone who isn't a paid-up member. If you're a SMIRK member and haven't renewed your dues, send \$3 to Ray at his callbook QTH.

new award program

In order to stimulate more weak-signal operation on 70 cm (432 MHz), Art Holmes, WA2TIF, has instituted a new awards program. Basically, there's a monthly award for making at least 50 QSOs, a monthly award for 1000 points by multiplying number of QSOs times grids worked, and an endorseable award for working 100 or more different stations on the band. The only stipulation is that contacts made on nets and during contests don't count. Write Art at 11 Kerr Road, Rhinebeck, New York 12572, for further information.

new records

There have been some unusually late and long DX tropo openings this year. The one during the U.S.A. Thanksgiving weekend will be long remembered because so many North American records were broken. Al-

though this represents preliminary information, I'll list those new records that I've been able to document.

The 70-cm tropo record is now held by Ray, WB3CZG, FN10AX, for a QSO on November 29, 1986, with Kent, WA5VJB, EM12LQ, of approximately 1318 miles (2120 km). On the same date, a new 23-cm record of approximately 1287 miles (2070 km) was set between Ray, WB3CZG, FN10AX, and Dave, KD5RO, EM13PA. Finally on November 28 and 29, 1986, the 13-cm (2304 MHz) record was broken several times by WB5LUA, KD5RO, K9HMB, and W8YIO. When it was all over, Dave, KD5RO, EM13PA, and Lew, W8YIO, EN82BE, emerged as the new record holders for a distance of approximately 940 miles (1516 km).

Just before this great weekend, Tony, K5PJR, EM26OP, and Jim, WA5ICW/5, EMO4HX, set a new 6-cm (5760 MHz) tropo record of approximately 285 miles (459 km). November certainly was an exciting month for UHF and microwave records! Congratulations should go not only to the new record holders, but to all those who — if only for a few minutes — held the record. Great going! Keep it up!

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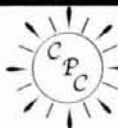
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WA6P operated this solar powered packet station during 1986 Field Day as part of the McDonnell Douglas Amateur Radio Club and Southern California Amateur Club entry. The photographer was WA6AUF.



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grid square index

Recently Folke Rosvall, SM5AGM, created *The Radio Amateur's World Atlas*, a 24-page book of maps and tables which show the geographical coordinates of all 32,400 grid squares in the world, along with an extensive index of cities worldwide. I've been using this book for some time and find it quite helpful in locating grid squares of stations in the USA as well as DX stations anywhere in the world. Since most of my acquaintances seem to be unaware of this book, I thought I'd mention it because it's so helpful with my record keeping. It's available from Ham Radio's Bookstore for \$3.95 plus \$3.50 shipping and handling.

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- April 21: ARRL 220-MHz Spring Sprint Contest (evening)
- April 22: Predicted peak of the Lyrids meteor shower at 1100 UTC
- April 24-26: Dayton Hamvention
- April 29: ARRL 432-MHz Spring Sprint Contest (evening)
- May 2-3: West Coast VHF Conference (contact WB6GFJ)
- May 5: Predicted peak of the Eta Aquarids meteor shower at 1300 UTC
- May 8: ARRL 902-MHz Spring Sprint Contest (evening)
- May 14: ARRL 1296-MHz Spring Sprint Contest (evening)
- May 15: EME perigee
- May 15-17: 13th Annual Eastern VHF/UHF Conference, Nashua, New Hampshire (contact W1EJ)
- May 23-24: ARRL 50-MHz Spring Sprint Contest (evenings)

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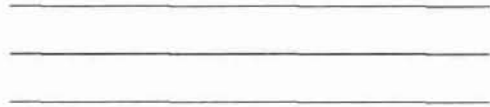
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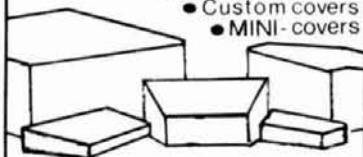
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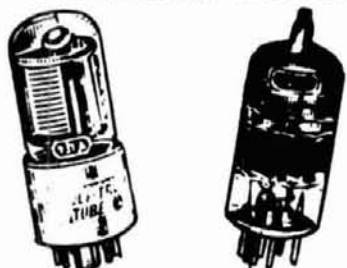
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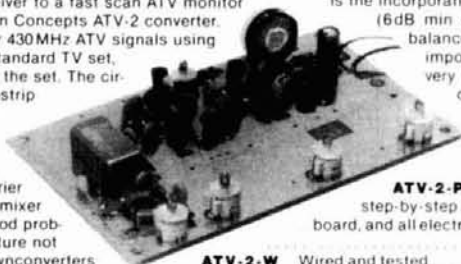
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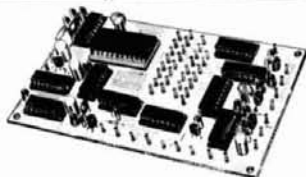
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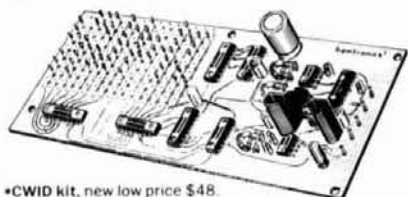
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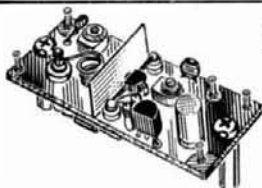
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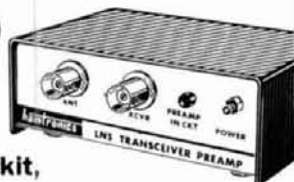
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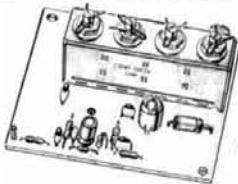
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This can result in a system with noise performance far less than ideal. For example, it might seem that an improvement in noise figure of a low-noise system would naturally result in lower total output noise and better output signal-to-noise ratio. However, this isn't always the case. It's quite easy to improve the noise figure to its optimum minimum value while causing an *increase* in output noise and a degradation in signal-to-noise ratio for any given signal. John Maxwell gives a good review of this problem in reference 1.

Noise figure is a "figure-of-merit" (a measure of "idealness") that demonstrates the amount of noise a system such as an amplifier adds to a signal-processing task such as amplification. If an amplifier were perfectly noiseless, it would add no additional noise and the noise figure would be zero. Similarly, a noise figure of 3 dB implies that the amplifier adds as much noise to the system as was initially present prior to amplifying (remember, a 3-dB increase in power is a doubling in power.) Also, the noise figure

is a measure of the amplifier's performance, not the circuit performance. That's why we may take a specific circuit and optimize the noise figure but actually degrade the output signal-to-noise ratio.

Noise figure, NF, is defined as 10 times the common logarithm of the "Noise Factor," F , and the noise factor is defined as the ratio of the total output noise power of a system (an amplifier, for example) to the output noise power due to the source alone.²

$$F = \frac{\text{Total output noise power}}{\text{Output noise power due to the source alone}} \quad (1)$$

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

The noise factor in **eqn. 1** is a ratio of the actual output noise of a system to the output noise that would be present due to the source alone if the system were perfectly noiseless. Two things are important to notice about **eqn. 1**: the first is that the noise factor is defined in terms of noise powers, and second is that it is defined at the output of the system. Examining **eqns. 1 and 2**, something is conspicuous by its absence — a signal. The noise figure apparently has no relation to any signal or to the S/N. In fact, you don't even need a signal to find the noise figure, and without a signal, you have an S/N of zero! Well, the noise factor (and noise figure in turn) may be loosely considered a noise-to-signal ratio where the source noise is both the signal and a component of the total noise.

However, the noise figure may be related to the S/N relatively simply. Consider an amplifier with the following parameters.

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N_S = Source noise power delivered to the amplifier input

N_I = Amplifier equivalent input noise power

G_P = Amplifier power gain

P_S = Amplifier input signal power

Now, the numerator of **eqn. 1**, total output noise power, is actually the sum of the noise contributed by the amplifier and that contributed by the source. Also, a specific output power is given by the corresponding input power multiplied by the amplifier power gain. We can then write the noise factor as **eqn. 3**.

$$F = \frac{(N_S G_P + N_I G_P)}{N_S G_P} \quad (3)$$

Now, if we divide both the numerator and denominator by the output signal power ($P_S G_P$) and rearrange the terms, the results of **eqn. 4** are found.

$$F = \frac{P_S/N_S}{(P_S G_P) / [(N_S + N_I) G_P]} \quad (4)$$

In examining **eqn. 4**, we see that the numerator is simply the signal-to-noise ratio of the input source. Similarly, the denominator is the signal-to-noise ratio at the amplifier output. So the noise factor is equal to the input S/N divided by the output S/N.

$$F = \frac{S/N_{input}}{S/N_{output}} \quad (5)$$

This is commonly known as the "Friis Equation," in honor of H.T. Friis,³ who originally developed it in 1944. A subtle but important point to observe about the Friis equation is that the term S/N input is the S/N of the *input signal, not the S/N at the amplifier input*. Also, it should be understood that this is *not* the definition of noise factor, but rather a derived expression computed from the definition given in **eqn. 1**.

optimum source resistance

For any amplifier system, a value of optimum source resistance exists which will produce an optimum minimum noise figure. The problem lies in the fact that the value of optimum source resistance is rarely even close to the resistance of the source we wish to use. The NF of the source-amplifier combination is then less than optimum. It is often thought that a resistor could be added in series or shunt with the source to modify its *apparent* value as seen by the amplifier to become

the optimum source resistance. This would indeed improve the NF with respect to the apparent source resistance, but not with respect to the original source.

The addition of this resistor has two effects. First, it loads the source dissipating some of the signal power that would normally be available as amplifier input, causing the signal term of the S/N expression to be reduced. Second, it adds the resistor's thermal noise to the system, causing an increase in the noise term of the S/N. The net effect of adding the modifying resistor is that the output signal is reduced, and the output noise is increased, which results in an overall reduction in the S/N. However, the NF is reduced to its optimum minimum value! Is this contradictory? Can we truly improve the NF of a system **and** actually degrade the S/N? If this is true, is the NF a useful parameter at all?

Perhaps a practical example or two using the parameters of real components will clarify this problem. (For more detailed analyses, check references 1 through 4.) The basic difficulty lies in the fact that the addition of a series or shunt resistor to a given source resistance only modifies the apparent resistance of the source. The actual value remains unchanged. The source contributes a specific signal and a noise commensurate with its resistance. For some specific signal, this provides some source S/N. The modified source will exhibit a lower available signal for the same actual signal level due to the loading effect of the added modifying resistor. The thermal noise of the modified source may be higher or lower than the original source, depending on whether the added resistor was added in series or shunt. However, in all cases the source S/N for any given signal level of the original source will be reduced. The addition of a simple resistor to modify the source then in effect makes the source "more noisy," which in turn renders any noise of the amplifier system less significant. That can cause an improvement in the NF. However, since the signal has been made more noisy, the S/N is degraded.

examples

Consider the circuit shown in **fig. 1**. This is a classic model of a generalized amplifier showing the noise parameters.⁴ The source E_{NS} is the RMS noise potential of the source resistor R_S . The amplifier noise is referred to the amplifier input as two *equivalent* input noise sources: an input noise potential E_{NI} , and an input noise current I_{NI} . If you check the specifications of various high-performance amplifiers, you'll find these parameters. It should also be emphasized that as with almost all noise parameters, the total noise available from the sources is a function of the noise bandwidth, BW_N , and the frequency of measurement. Generally, the spot noise values (noise per Hertz of bandwidth) are specified as a function of frequency.

In the model, the amplifier input resistance is R_I and the voltage gain is A_V . Both of these are defined as noiseless because all noise contributions are included in the two equivalent input noise sources. It can be shown that the optimum source resistance that yields the best NF is given by eqn. 6.²

$$R_S (\text{optimum}) = E_{NI}/I_{NI} \quad (6)$$

With that optimum source resistance, the optimum NF is given by eqn. 7.²

$$NF (\text{optimum}) = 10 \log \left(1 + \frac{E_{NI} \cdot I_{NI}}{2kT} \right) \quad (7)$$

where: k = Boltzmann's constant = 1.38×10^{-23} watts-seconds/ $^{\circ}k$

T = Absolute temperature in $^{\circ}k$

As an example, consider the source as a magnetic phonograph cartridge with a source resistance of 1000 ohms and a noise bandwidth of 20 kHz. Also, consider an LF356 FET amplifier configured for a voltage gain of 100. The LF356 parameters from the data sheets are given below.

- R_I = 10^{12} ohms
- E_{NI} = 12 nV/Hz @ 1 kHz
- I_{NI} = 0.01 pA/ Hz @ 1 kHz (shot noise of input current)
- A_V = 100 (configured gain)

To simplify the example, the noise contributions of any circuit resistors are ignored. (If we were actually trying to optimize noise performance, those noises would be included.) Also, consider that the 1 kHz spot-noise values given are constant over the frequencies of interest. From eqn. 6, the optimum source resistance is found to be 1.2 megohms; from eqn. 7, the optimum NF at that source resistance is found to be 0.063 dB! This very low optimum NF is typical of FET-input devices, but the very high optimum R_S often results in poorer practical noise performance (S/N) than that provided by higher noise bipolar devices. This occurs because the optimum source resistance for typical bipolar elements is often much nearer the value of typical sources (such as our phonograph cartridge) than that of FET devices. This will be demonstrated below. The NF for this amplifier with the 1-k source is 9.9 dB. The output noise is 179 μ V RMS. (Note: noise signals must be added as powers or by adding their mean-square values.) Adding a 1,199,000-ohm resistor in series with the 1-k source provides an optimum 1.2-megohm source resistance to the amplifier. The output noise is then 2 mV RMS and the NF is indeed 0.063 dB. The optimum NF configuration will have an output noise eleven times higher than the less optimum unmatched case and in both cases the gain will be 100 (the 10^{12} -ohm amplifier input resistance does not load either

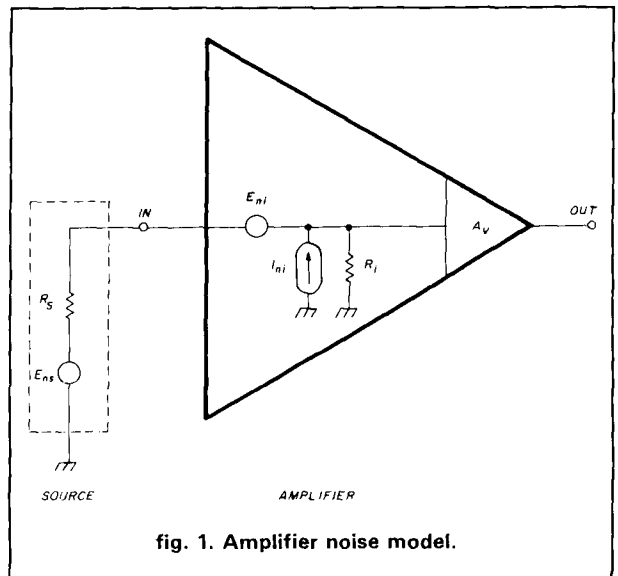


fig. 1. Amplifier noise model.

source configuration significantly). Then, for any given signal, the optimum NF configuration will exhibit an output S/N eleven times poorer than the unmatched configuration. So, even though the NF is poorer for the unmatched case, the noise performance is better: 21 dB better!

Now we'll try matching in a more optimum manner. Let the 1-k source be transformed to the optimum 1.2-megohm resistance with an ideal transformer having a 1:34.6 turns ratio. Note that this is not an optimum power match. We'll define the transformer as noiseless, but in a very accurate analysis its winding resistances would contribute some thermal noise that would have to be considered. Also, we'll adjust the amplifier gain to provide the same overall gain of 100 of the previous example to allow simple comparison. This does not affect the S/N or NF since both of these are independent of system gain and the transformer was defined as noiseless. Now the total output noise is 57.97 μ V and the noise due to the source alone is 57.55 μ V, resulting in the expected 0.063 dB NF. With the selected amplifier, this configuration will provide the best possible S/N for any given signal. It is 9.8 dB better than the simple unmatched case ($20 \log 170 \mu\text{V}/57.97 \mu\text{V}$) and 30.8 dB better than the resistive matched case.

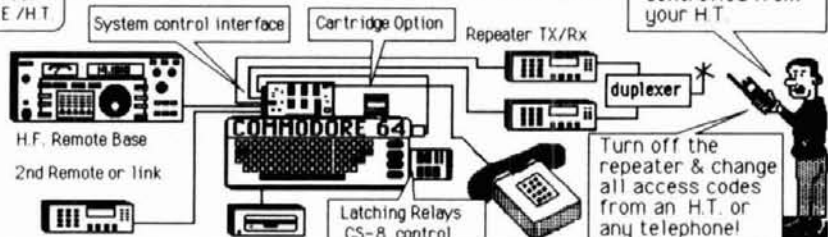
This may still be a little confusing, and you may be convinced that NF is a useless parameter. Let's examine what was done in the examples above. First we started with a device with a 1-k source resistance and an amplifier with a 1.2-megohm optimum source resistance and gain of 100. Direct application of the source to the amplifier gave us a signal gain of 100 with a 179 μ V output noise and a 9.9 dB NF. Adding a 1,199,000-ohm series resistor gave us an optimum noise match with respect to the amplifier, with an out-

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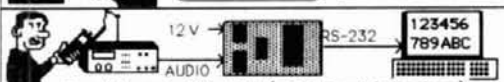
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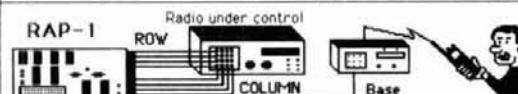
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put noise of 2 mV and a 0.063 dB NF with the signal gain still 100. Since the gain was the same in both cases, the output signal would be the same for both for a given input signal. Then, since the output noise was higher in the second case than in the first, the S/N would be poorer, even though the NF was better.

What did we do wrong to get the lower NF case to give us a poorer noise performance? It's actually very simple. When we added the series resistance, the effective source seen by the amplifier was optimized, but the *actual* source resistance was unchanged. We inadvertently defined the source resistance in two different ways: as the real source resistance in one case and as the source plus a series resistance combination in the other. The 1-k source resistance is part of a circuit element that contributes both signal and noise. Adding a series resistance adds a circuit element that contributes only additional noise and no additional signal, so it should be expected that the S/N should be made poorer.

Looking back at eqn. 1, we see that the numerator is related to the source noise. If we define the source as the 1000-ohm resistance of our actual phonograph cartridge, we'll compute one value of output source noise, but if we use the 1.2-megohm resistance of the combined resistances, we'll find another (higher) value. As far as our amplifier is concerned, all components attached to its input constitute the source. However, to our actual source, only its elements constitute the source. When we add resistors to modify the source resistance, which value of source resistance should we use for the NF calculations? That depends upon what we're trying to find. If we wish to find the NF of an amplifier, everything tied to the input is then the source. If, however, we are trying to determine the NF of an amplifier with a specific source, only the actual source should be considered as the "source." In general, the actual source is the true source as far as the total circuit is concerned. Adding extraneous components to the source doesn't change its actual value, only its apparent value seen by the amplifier. Since it's the signal from a specific source that we're generally interested in processing, we should always compute the "output noise due to the source" in eqn. 1 from the actual source of interest and not that "seen" by the amplifier input.

With this in mind, let's go back to our original phonograph cartridge source and LF356 amplifier. We found that the NF computed from eqn. 2 with the source directly applied to the amplifier was 9.9 dB. When we added the 1,199,000-ohm resistor and defined the 1.2-Megohm source plus series resistance combination as the "source," we found an NF of 0.063 dB. Now, what would the NF be if we added the series resistor but still used the 1000-ohm value as the source resistance? It would be 30.9 dB! This clearly

shows that with respect to our actual source, the addition of a series resistor only makes the NF poorer.

If we could in some manner make the actual source resistance and the source resistance seen by the amplifier equal the optimum source resistance, we would achieve the best possible noise performance. For example, if we could place in series 1200 of our phonograph cartridge sources (with each delivering the same signal), we would achieve the optimum source resistance with the actual source. That would provide the optimum NF and a very good S/N because the available signal power would be 1200 times that of a single source. However, that solution is obviously very impractical. We could possibly redesign the source to exhibit the 1.2 megohm resistance, but that too is generally impractical. The transformer offers us a very practical means of *transforming* the source resistance to the optimum needed, and well-designed transformers can do this almost noiselessly. With a transformer, both the signal level and source impedance are scaled together: potentials and currents by the turns ratio, and resistances by the turns ratio squared. This maintains the source S/N, as seen by the amplifier, constant. So, in our example, the entire equivalent 1.2-megohm source resistance at the transformer output contributes signal, 34.6 times more than the 1k-ohm source with a 1,199,000-ohm series resistor, with the same noise contribution of the optimum 1.2-megohm resistance. That obviously results in a much better S/N than the simple addition of a series resistor because the signal has been increased along with the increase in the source resistance seen by the amplifier.

In some applications we have control of the design of the source and can tailor its resistance to optimize noise performance with a specific amplifier. However, in the majority of cases the source resistance is fixed and we must select, or design, an amplifier whose optimum R_S is close to the specified source, or use a transformer to provide noise matching. For example, the LM1897 amplifier has an optimum NF of 1.82 dB at an optimum source resistance of 5.33 k. That optimum NF is much poorer than the 0.063-dB value of the LF356. However, this amplifier provides a 3.87-dB NF with the 1-k source resistance, which is much better than the 9.9-dB value provided by the lower noise LF356 with the 1-k source resistance. The higher noise amplifier, with the specified source resistance, will actually provide a 6-dB higher S/N than the lower noise unit for any given signal. This is because the LM1897 has an optimum source resistance value that more closely matches the 1-k source than the LF356. This shows that just because one amplifier has lower optimum noise figure than another, it may not provide better noise performance in a specific configuration or application.



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Finally, there's the question of how the optimum NF source resistance relates to the optimum power match source resistance. That's very simple; they're totally unrelated. To demonstrate this, we can use a transformer to match the 1-k source resistance to the 10^{12} -ohm input resistance of the LF356, a turns ratio of 1:31,620. This will provide an optimum power match, but the NF will be 37.8 dB! That's considerably poorer than the 0.063 dB optimum NF provided at the optimum noise match. This is because the optimum noise source resistance is related to the equivalent input noise sources, and the optimum power match source resistance is related to the input resistance. In general, the equivalent input noise sources and the input resistance are unrelated, rendering the optimum matching resistances similarly unrelated.

The principal purpose in optimizing the power matching is to maximize power gain. In the case of optimizing noise performance, gain is not a particularly important consideration since additional gain stages may be added to obtain the total desired gain once the optimum noise performance is achieved. However, the process cannot be approached in reverse; you can't match for optimum gain and then optimize noise performance.

When using NF parameters specified for some specific device with which you are designing, you must carefully examine those specifications to be sure you understand what's presented. This is particularly true in the case of rf components. Quite often curves of optimum noise figure are presented showing the optimum NF as a function of frequency or bias current, but rarely is the actual value of the optimum source impedance given. Only at the optimum source impedance can the optimum NF be achieved. Since most rf measurements are made in systems of standard impedance such as 50 ohms, it might be assumed that the NF curves are given for that source impedance. If you make that assumption, you'll often find that you can't achieve the NF values specified. In other cases, only the NF curves at the measurement system impedance are given, providing no information for optimizing the source impedance for best NF. Even worse, you often can't tell from the provided information which source impedance was used. There are, of course, exceptions such as the NE388 series MESFET, but these are all too rare.

The situation is much better in the area of the lower frequency components. For those components, the "Contours of Constant Noise Figure" are quite often given. These show numerous curves of NF as a function of source resistance, operating point, frequency, etc. Using these data, one can quite effectively design low-noise audio, i-f and low-frequency rf amplifiers that provide the performance predicted by the mathematical design. In fact, performance somewhat

better than that specified for the part can often be achieved because the manufacturer must use conservative specifications to allow for a range of variation in the component line. In those cases where sufficient data isn't provided, you must make your own measurements of the various needed parameters. At lower frequencies that's relatively simple, but at the higher frequencies (i.e., greater than 100 MHz), that can be quite difficult without a collection of precision equipment.

In conclusion, noise figure can be an excellent figure-of-merit for assessment of the potential noise performance of amplifiers and other systems, but a thorough understanding of the parameter is necessary if it's to be successfully applied. The performance suggested merely by the good noise figure of a device may not be reflected in the actual signal-to-noise ratio achieved from the device in a practical application. In an application where the source characteristics are predefined, it's generally more important to use an amplifier with an optimum source resistance that closely matches the specified source resistance than to merely choose an amplifier with the lowest possible noise figure. In no case can the signal-to-noise ratio of a system be improved by modifying the source resistance with a series or shunt resistor for making the combined resistance equal to the optimum source resistance for the system. This process can indeed result in the optimum noise figure with respect to the apparent source seen by the amplifier, but will always degrade the output signal-to-noise ratio for any given signal. If the "source" element used for determining the noise figure is always chosen as the actual source alone, meaningful results will be obtained.

It should be understood that the parameter noise figure is simply a figure-of-merit. In practical systems, it's the actual signal-to-noise ratio that's of primary interest. However, the optimum noise figure for an amplifier (or other system) does represent the best possible performance that can be achieved with that amplifier when the source resistance is properly matched to the optimum resistance for the amplifier. If you always compute noise figure using the actual physical source as the source referred to in eqn. 1, you'll always arrive at a correct result.

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 FL-63 250 Hz CW filter (1st IF) 54.50
 FL-52A 500 Hz CW filter (2nd IF) 108.00 99⁹⁵
 FL-53A 250 Hz CW filter (2nd IF) 108.00 99⁹⁵
 FL-33 AM filter 35.25
 FL-70 2.8 kHz wide SSB filter 52.00
 RC-10 External frequency controller 39.25

Other Accessories:
 IC-2KL 160-15m solid state amp w/ps 1999.00 1699
 PS-15 20A external power supply 169.00 154⁹⁵
 PS-30 Systems p/s w/cord, 6-pin plug 299.00 269⁹⁵
 OPC Opt. cord, specify 2, 4 or 6-pin 10.00
 MB Mobile mount, 735/745/751A 24.50
 SP-3 External speaker 61.00
 SP-7 Small external speaker 49.00
 CR-64 High stab. ref. xtal (745/751) 63.00
 PP-1 Speaker/patch 159.25 149⁹⁵
 SM-6 Desk microphone 44.95
 SM-8 Desk mic - two cables, Scan. 78.50
 SM-10 Compressor/graph EQ, 8 pin mic 136.25 124⁹⁵
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 AT-500 500W 9-band auto. antenna tuner 559.00 489⁹⁵
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 EX-248 FM unit 55.50
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VHF/UHF base multi-modes
 IC-551D 80W 6-meter SSB/CW Regular SALE 799.00 719⁹⁵
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 AG-1* Mast mounted preamplifier ... 99.50
 IC-471H* 75W 430-450 ... CLOSEOUT 1399.00 999⁹⁵
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*Preamp \$99⁹⁵ with 271A/471A/471H Purchase

Accessories common to 271A/H and 471A/H
 PS-25 Internal power supply for (A) ... 115.00 104⁹⁵
 PS-35 Internal power supply for (H) ... 199.00 179⁹⁵
 SM-6 Desk microphone 44.95
 EX-310 Voice synthesizer 46.00
 TS-32 CommSpec encode/decoder ... 59.95
 UT-15 Encoder/decoder interface ... 14.00
 UT-15S UT-15S w/TS-32 installed ... 92.00

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 IC-490A 10W 430-440 SSB/FM/CW 699.00 599⁹⁵

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 IC-47A Compact 25W 440 FM, TTP mic 549.00 479⁹⁵
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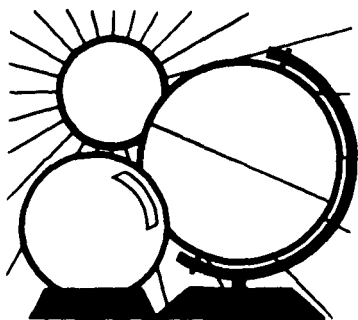
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DX FORECASTER

Garth Stonehocker, KØRYW

a spring dx solution

Atmospheric noise from thunderstorms can be a problem when you're chasing DX during this time of the year. Noise is propagated via the ionosphere just like DX.

In March and April, spring storms occur in the northern hemisphere. Fronts of warm and cold air generate the first major thunderstorms of the year, with fast-moving cold fronts producing particularly potent thunderstorms. As a storm front approaches your area, you'll begin to hear a significant increase in the noise level. You'll first notice this increase at a one-hop distance (about 600 to 1200 miles) when the storm front is about one day west of your location. You can reduce the received noise a few dB by using a directional antenna such as a rotating Yagi or a phased vertical array. Determine the noise direction and work DX in the opposite direction, or do your best to null it out using a directional trade-off between signal and noise strengths. Antennas with a low take-off-angle (TOA) at the operating frequency are best because noise normally arrives at angles greater than 30 degrees.

As the front draws nearer, the noise level will usually decrease until it's within a ground-wave's distance (about 50 miles). Now, loud individual discharges will be heard. A horizontally polarized antenna is the best radiator to use to lower the noise as much as possible. As the storm approaches, its sounds

become part of the "local noise"; as it moves away, its noise decreases, then increases again as the front reaches the one-hop distance point a day or so later. The directional-low TOA antenna again becomes helpful. (Correlate your observations with storm progress reports on the local television weather program.)

In looking for rare DX, you can save time by tracking storms in order to pinpoint when and where the most favorable listening conditions are likely to occur.

Over the years of gathering information for and writing this column, I've kept an ear open for data that would verify the events and propagation conditions presented in previous *last-minute forecasts* and *highest band available charts* in this column. To track the accuracy of these predictions, I plot daily values of solar flux and geomagnetic A (index) on a 27-day epoch graph and find it's hard to be right much over 60 percent of the time.

To verify published data, I monitor the ham bands and keep data collection in mind while DXing. What I usually look for is the highest band with signals, then note where they're coming from and the time the band is changing (out or in) towards that direction. During the month, I get a better sense of how good that month's chart was.

What do you think? Could the chart information be better? Let me know how you think it's working.

last-minute forecast

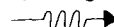
DX conditions on the higher frequency bands, 10-30 meters, are expected to be best through the first week and a half. The probability of transequatorial openings should be greater during this period of higher-than-normal solar flux, especially if geomagnetic disturbances materialize as expected on the down slope of this short-term flux peak. This month is still within the spring equinox disturbance window, so be aware and take advantage of this opportunity to work some southern stations.

The lower frequency bands will probably be affected by the disturbances expected around April 8-10 and 15-20. However, these will affect east, north, and west paths with lower MUFs and lower signal strengths with QSB. Look for DX from unusual locations if the disturbance isn't too strong — K's greater than 5 or 6. Otherwise, the second and third weeks should be good for DX, with low atmospheric noise except when spring weather frontal thunderstorms pass your QTH.

The perigee of the moon's orbit (for moonbounce DX) is on the 18th, with the moon showing full phase on the 14th. 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.

On March 29, expect a total eclipse of the sun in the southern part of South America and Antarctica, going up to Southeast Europe and Asia.

Ten, twelve and fifteen meters, the day-only DX bands, will be open mid-day to early evening almost every day to southern areas of the world. The openings on the higher of these bands will be shorter (if they occur at all), closer to local noon, and provide a possibility of transequatorial openings.



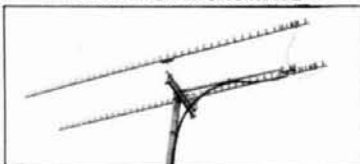
GMT	WESTERN USA									
	N	NE	E	SE	S	SW	W	NW		
0000	20	30	20	12	15	10	10	15		
0100	15	30	20	12	15	10	10	15		
0200	15	30	20	15	20	10	10	15		
0300	15	30	20	15	20	10	10	15		
0400	15	30	20	15	20	10	10	15		
0500	20	30	30	20	20	10	10	15		
0600	20	20	20	20	30	10	10	15		
0700	20	20	15	20	30	12	12	20		
0800	20	20	15	20	30	12	12	20		
0900	20	30	20	20	30	15	20*	20		
1000	20	30	20	20	30	20	20*	20		
1100	20	20	20	20	30	20	20	30		
1200	20	20	15	20	40	20	20	30		
1300	20	20	12	20	40	20	20	30		
1400	20	15	12	15	40	20	20	20		
1500	20	15	10	12	40	20	30	30		
1600	20	15	10	12	40	20	30	30		
1700	20	15	10	12	20	20	30	30		
1800	20	15	12	10	20	15	30	20		
1900	30	15	12	10	20	12	20	20		
2000	20	20	15	10	15	12	12	20		
2100	20	20	15	10	15	10	10	20		
2200	20	20	20	10	15	10	10	20		
2300	20	20	20	10	15	10	10	15		

GMT	MID USA									
	N	NE	E	SE	S	SW	W	NW		
0000	20	20	20	12	15	10	10	15		
0100	20	30	20	12	15	10	10	15		
0200	20	30	20	15	20	10	10	15		
0300	20	30	20	15	20	10	10	15		
0400	20	30	20	15	20	10	10	15		
0500	20	30	20	15	20	10	10	15		
0600	20	20	20	15	20	10	10	15		
0700	20	20	12	15	40	20	20	20		
0800	20	15	12	12	40	20	30	30		
0900	20	15	10	12	40	20	30	30		
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1400	20	20	15	10	20	12	12	20		
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1700	20	20	20	10	15	10	10	15		
1800	20	20	20	10	15	10	10	15		
1900	20	20	20	10	15	10	10	15		
2000	20	20	20	10	15	10	10	15		
2100	20	20	20	10	15	10	10	15		
2200	20	20	20	10	15	10	10	15		
2300	20	20	20	10	15	10	10	15		

GMT	EASTERN USA									
	N	NE	E	SE	S	SW	W	NW		
0000	20	20	20	12	20	10	10	15		
0100	20	20	20	12	20	10	10	15		
0200	20	30	20	15	20	10	10	20		
0300	20	30	20	15	20	10	10	20		
0400	20	30	20	15	20	10	10	20		
0500	20	30	20	15	20	10	10	20		
0600	20	30	20	15	20	10	10	20		
0700	20	30	20	15	20	10	10	20		
0800	20	30	20	15	20	10	10	20		
0900	20	20	15	20	40	20	20	20		
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2000	20	30	20	20	30	12	12	20		
2100	20	30	20	20	30	12	12	20		
2200	20	30	20	20	30	12	12	20		
2300	20	30	20	20	30	12	12	20		

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.
 *Look at next higher band for possible openings.

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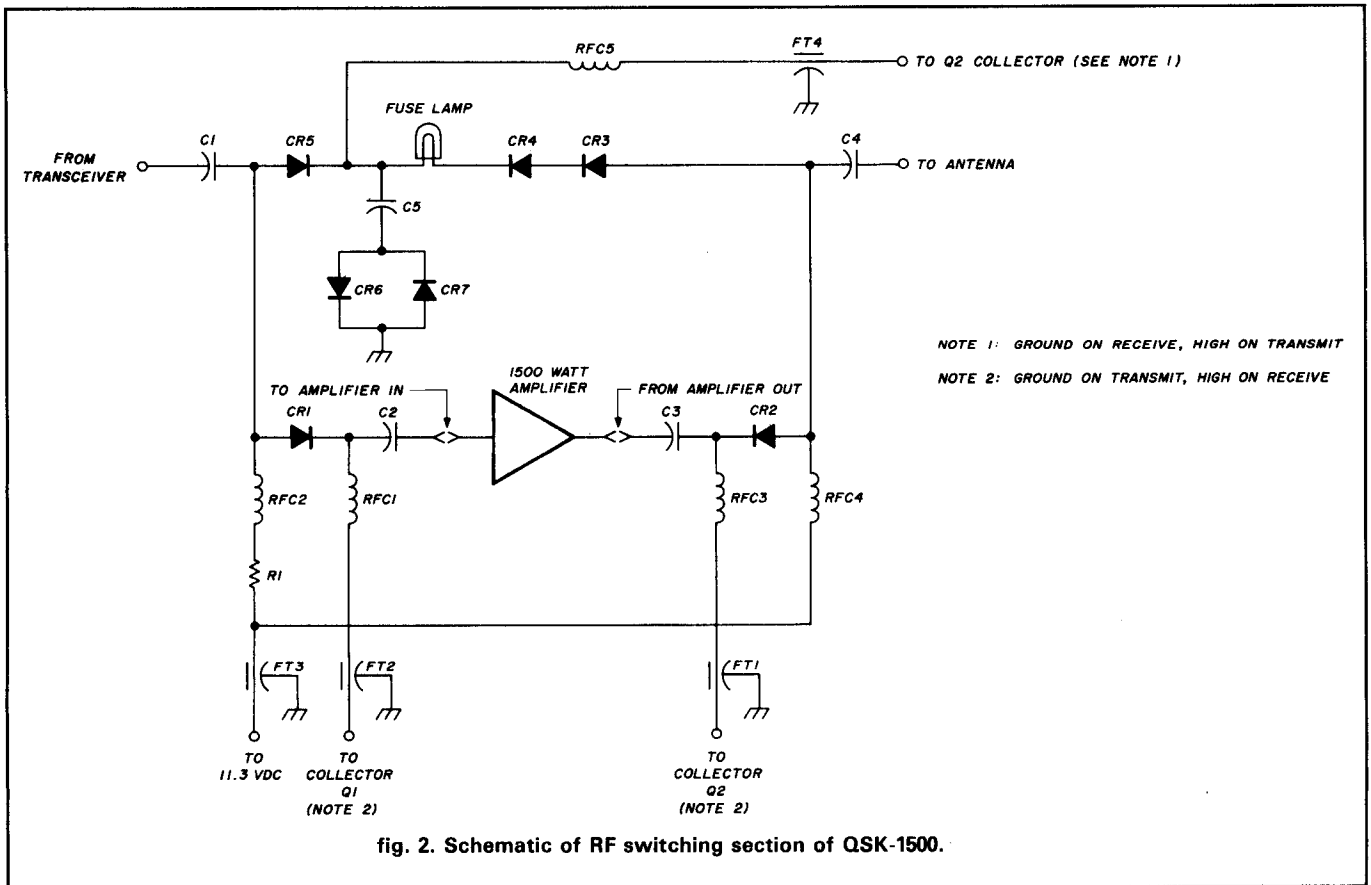
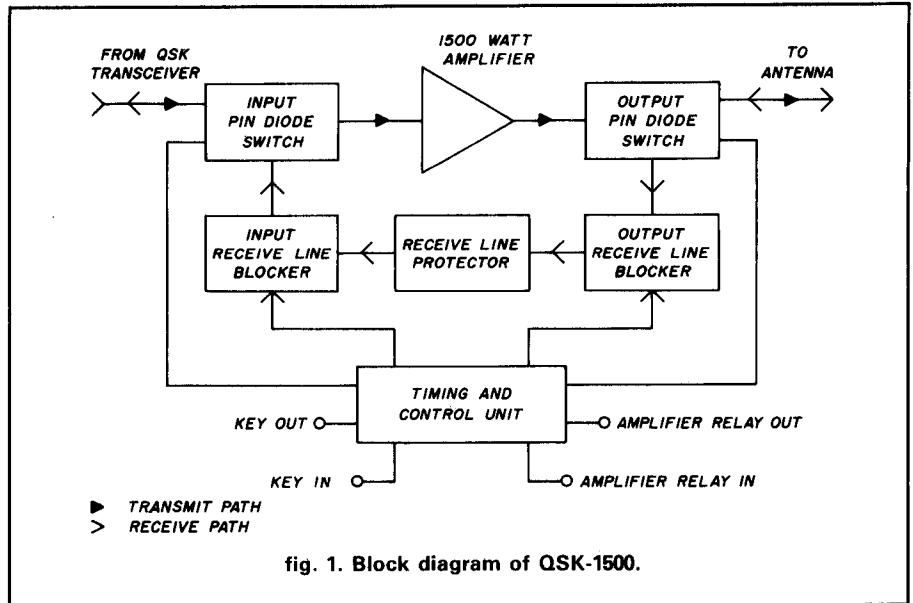
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REVIEW

the DEO QSK-1500

When I was a Novice back in the 1960s, my club station had a Johnson Ranger I, a Hammarlund HQ-120, and a Johnson TR switch. I spent hours on the air and really enjoyed being able to work full break-in CW. I got spoiled in that first year of being a ham, not having to listen to the clunking of relays and switches.

Since then I've dreamed of operating break-in CW again. Until recently, however, I never had a fully QSK-compatible radio. When I finally acquired one of the new do-it-all-but-burp-the-baby radios, QSK was once again possible. But I couldn't operate high power QSK. What to do?



QSK — what is it?

Basically stated, QSK is the ability to hear between the dots and dashes while transmitting a CW signal. A number of methods can be used to effect QSK operation: separate transmit and receive antennas, a vacuum tube TR switch such as the Johnson mentioned above, and expensive vacuum switches.

Each of these designs, however, presents a number of difficulties. Not everyone has a receiver with an AGC circuit capable of operating in the presence of a strong rf field without "folding up" or the space necessary for separate transmit and receive antennas. The TR switch had two significant problems: it created a tremendous amount of TVI and caused attenuation on the receive signal (commonly called "suck-out"). Vacuum switches require complex, precisely timed circuits to prevent "hot-switching" and aren't cheap to manufacture. Another problem was that some of these schemes introduced distortion onto the transmitted signal.

modern technology to the rescue

Several technological advances in the late 1960s brought QSK closer to reality for the average ham. Until the late 1970s, however, TENTE was the only manufacturer to offer a QSK-compatible radio.

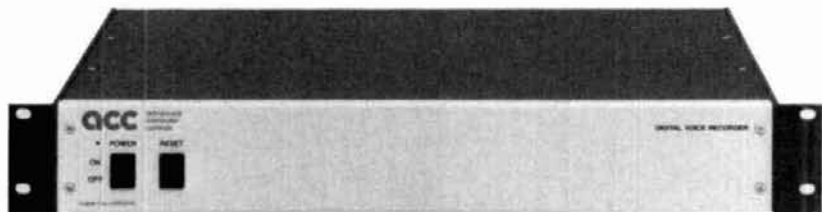
By 1980, transceiver manufacturing took a quantum leap forward and QSK-compatible radios were available from almost all radio manufacturers. Unfortunately, one problem remained; unless you owned either an Ehrhorn Alpha 77 or 78, you couldn't operate high-power QSK. Several after-market, add-on QSK units were introduced, but they suffered many of the same problems as other units.

In mid-1984, John ("Doc") Sheller, KN8Z, designed a solid-state QSK switch that solved many of the problems of earlier QSK units, using state-of-the-art PIN diodes as rf switches.

PIN stand for layer doping (P), intrinsic layer (I), pure and no doping (N) layer. The thickness of the intrinsic layer determines the characteristics of the diode and allows the manufacturer to custom-design the diode for any of many different kinds of applications. A PIN diode is a solid-state device that acts like a variable resistor at rf frequencies. The amount of forward dc bias applied to the PIN diode determines the resistance (impedance) to rf signals. Doc's design is unique in that it has no moving parts — no relays — and can hot-switch high power rf. It needs only low dc voltage biasing to control rf currents, and by design, doesn't introduce any significant waveform distortion on the transmitted signal.

how it works

A block diagram and schematic (figs. 1 and 2) illustrate how the QSK-1500 works. The received signal travels from the antenna through the output line blocker, the receive line protector, and the input receive line blocker into the



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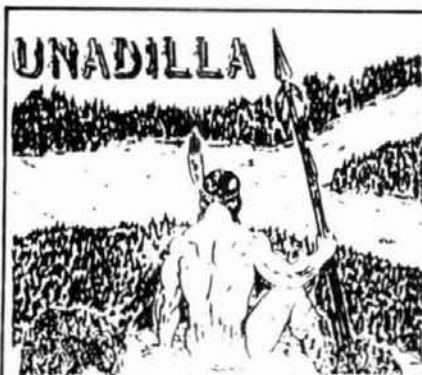
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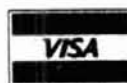
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front end of the transceiver. This circuit completely bypasses the conventional circuit path through the amplifier. The receiver is prevented from seeing the tank circuit of the amplifier by PIN diode CR2, which is reversed biased. PIN diodes CR3, 4, and 5 are forward biased and offer a very low impedance path for the receive signal. Typical insertion loss is less than 0.5 dB.

As you might expect, the transmit path is a little more complicated. Keying the transmitter triggers the timing circuit. The amp relay line is grounded, keying the amplifier. The input and output receive line blockers CR3, 4, and 5 are reverse biased with 525 volts dc; input and output PIN diodes CR1 and 2 are forward biased. The key-out line is closed and keys the transmitter. Rf flows from the transceiver through C1, CR1, and CR2 into the rf amplifier, it then flows through CR2, CR4, and C3 to the antenna. The reverse biasing of CR3, 4, and 5 prevents any rf from passing through the receive line. As soon as the morse character is completed, all the PIN diodes reverse state and the unit is back in a receive mode. PHEW!

So that's how this thing works. When you think about it, it's really quite simple. (For a more complete explanation of PIN diodes and how they work, see Doc's article on PIN diodes that appeared in the January, 1986, issue of *ham radio*.)

hookup and use

Setting up the QSK-1500 requires making four control cables. DEO recommends that you use RG-58 or other shielded cable to eliminate the possibility of rf getting into one of the units. Line 1 connects your key or keyer to the QSK-1500's timing circuit; line 2 connects your transceiver's amplifier control circuit to the timing circuit again. Line 3 connects the timing circuit back to the transceiver's keying line, and line 4 connects the timing circuit to the RF amplifier.

Rf connections are straightforward and can be followed by referring to fig. 1. Now you're ready to run a few tests. Turn your transceiver on. If everything is working correctly, you should hear nothing. This is because the receive line diodes aren't "on," so there's no receive path to your receiver. When you turn on the QSK-1500's power supply, the LEDs should light up; you should also hear signals. After a few more simple tests, you're ready to crank up the power and operate.

Once testing was complete, I was excited to see how QSK could actually be used. Tuning the low end of 160, I decided to look for a European or two; finding a fairly strong G3 calling CQ, I dropped my call in a couple of times, only to hear him start calling CQ again. I stopped calling immediately and waited for him to complete his CQ. Again I called, only to have him call another European. Stopping again, I realized how much QRM could be eliminated if more hams operated break-in CW — no more long-winded responses at 5 wpm while the DX station was calling again. QSO rates would increase dramatically, and many more of the "deserving" would get through.

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LOH 141.7° w	FRQ 145,8093	AZIM 143.3°
HGT 28938 km	DOP -697 Hz	ORBIT 1562
RNG 23571 km	DRFT 5 Hz	78

HOUSTON OSCAR 11 → 1985 JUL 11 03:44:33

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SRT
OBS
EPOCH
ASTRO
MOVE
HELP
QUIT

LAT 30.2° n	ECHO 8 ms	ELEV 29.5°
LOH 93.5° w	FRQ 145,8228	AZIM 9.9°
HGT 691 km	DOP -2628 Hz	ORBIT 7253
RNG 1245 km	DRFT -528 Hz	27

LONDON OSCAR 9 → 1985 JUL 11 04:14:24

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LAT 49.2° n	ECHO 6 ms	ELEV 28.9°
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No. AVT40-10 4 Band 18'9" \$139.95
No. AVT20-10 3 Band 11'4" \$109.95
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Monthly sales averaging \$10,000 for the last three years have provided reasonable salaries and good dividends because of generous profit margins and legitimate tax allowances for home-office rental, pro-rata car operating expense, ham and office equipment depreciation, etc. Easy ownership transfer without interruption of cash flow. "Hand-holding" is available during transition.

Detailed information will be sent only to those judged to have the best chance of success. Accordingly, serious inquirers should write giving biographical data including age, education, technical background, and experience. Box 498 Greenville, NH 03048

The only problem I've had with this unit occurred one evening when I was working in the shack and had the radio on, but the QSK unit off. I dropped a hammer, tripping the radio's VOX. Later when I sat down to operate, I turned the rest of the equipment on, only to find that the receive line through the transmit antenna was dead! Since I could still receive through the Beverages (which bypass the QSK 1500 on receive), I ruled out the possibility of a radio problem. A quick call to DEO solved my problem immediately. Doc suggested I check the receive line fuse lamp; sure enough, it was blown. Inserting a new lamp cured the problem quickly and simply. Doc did advise me, however, to always make sure the QSK 1500 is turned on whenever the transceiver is. Without the fuse lamp, there's the possibility of damaging the PIN diodes, necessitating some fairly extensive repairs.

Doc reports that contesters who have QSK-1500 units have been pleased with the results; I'm anxiously awaiting the next contest to see for myself. I've had a lot of fun operating QSK. I expect the QSK-1500 is one review unit I'll use for quite awhile.

The QSK-1500 consists of a power supply and a control unit. The power supply measures 3 x 6 x 4 inches and the control unit 3 x 7 x 9 inches. One bit of warning: don't place the QSK-1500 on top of a radio, an amplifier, or any other heat-generating unit. The PIN diodes alone create plenty of heat in normal operation, and damage might result from improper placement.

Priced at \$299 (plus \$6 shipping and handling), the QSK-1500 is available from either Design Electronics of Ohio, 4925 S. Hamilton, Groveport, Ohio 43125, or from Universal Electronics, 1280 Aida Drive, Columbus, Ohio 43068.

— de N1ACH



new ICOM catalog

ICOM America, Inc., has released its new 24-page, full product line color catalog. The new catalog includes all new products including the Micro 4AT 440-MHz handheld and the IC-275 all-mode VHF and IC-475 all-mode UHF base station.

For your copy, contact ICOM America, Inc., 2380 116th Avenue N.E., Bellevue, Washington 98004.

Circle #303 on Reader Service Card.

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10 Watts in - 100 Watts out

..... 1 W to 15 Watts

..... 10 Watts in - 100 Watts out

..... 3 Watts in - 50 Watts out

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RF Power In.....

RF Power Out.....

2 Watts in - 100 Watts out

..... 300 mW to 3 Watts

..... 2 Watts in - 100 Watts out

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- TS-940/930/830 Super Selective CW Switch Kit - This new kit allows you to add another CW bandwidth. For example, on the 930/940/830, if you already have 400Hz or 500Hz filters installed, you can now select a set of IR1250Hz filters. Our 250Hz matched set allows for a lower noise floor for super quiet DXing.

- 9 mHz 8-Pole Crystal Filters for Experimenters - All 8-Pole Filters - \$60 All 6-Pole Filters - \$50

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| For SSB | | For FM | |
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| C. 1.8kHz @ 6dB | D. 2.2kHz @ 6dB | For CW | |
| ★ E. 2.2kHz @ 6dB | A. 600Hz @ 6dB | A. 600Hz @ 6dB | ★ B. 600Hz @ 6dB |
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| ★ B. 6.0kHz @ 6dB | \$5.00 shipping and handling on all orders. | | |

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TRAP DIPOLES:

Model	Bands	Traps	Length	Price
D-42	10/15/20/40	2	55"	\$59.95
D-52	10/15/20/40/80	2	105"	84.95
D-56	10/15/20/40/80	6	82"	109.95
D-66	10/15/20/40/80/160	6	163"	129.95

TRAP VERTICALS - "SLOPERS":*

Model	Bands	Traps	Length	Price
V5-41	10/15/20/40	1	28"	44.95
V5-52	10/15/20/40/80	2	49"	59.95
V5-53	10/15/20/40/80	3	42"	69.95
V5-64	10/15/20/40/80/160	4	73"	89.95

*Can be used without radials
*Feed line can be buried if desired
*Permanent or Portable Use

ALL TRAP ANTENNAS are Ready to use - Factory assembled - Commercial Quality - Handle full power - Comes complete with: Deluxe Traps, Deluxe center connector, 14 ga Stranded CopperWeid ant wire and End Insulators. Automatic Band Switching - Tuner usually never required - For all Transmitters, Receivers & Transceivers - For all class amateurs - One feedline works all bands - Instructions included - 10 day money back guarantee!

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Model	Band	Length	Price
D-15	15	22"	18.95
D-20	20	33"	19.95
D-40	40	66"	22.95
D-80	80/75	130"	25.95
D-160	160	260"	34.95

Includes assembly instructions, Deluxe center connector, 14 ga Stranded CopperWeid Antenna wire and End Insulators.

COAX CABLE: (includes PL-259 connector on each end)

Type	Length	With antenna purchase	Separately
RG-58	50'	\$8.00	\$11.95
RG-58	90'	12.00	16.95

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computer rotor control interface

The most often requested new feature for satellite enthusiasts is automatic control of the antenna pointing via a personal computer. While the KENPRO KR0-5400A provides the electro-mechanical interface to the rotor motors, the actual hardware interface to the computer is provided by the KR-010. The KR-010 consists of analog-to-digital converters for both the elevation and azimuth channels and also provides the drive signal for driving the motors in the desired direction. An internal microcomputer provides the RS-232 serial capability at 9600 Baud without handshaking.

The KR-010 attaches to the serial port of the computer and operates with the tracking software (available only from the AMSAT software exchange, not from ENCOMM or KENPRO) written for AMSAT by W3IWI and W0SL. KENPRO and ENCOMM provide the software needed to point the antenna from data entered into a basic program in real time. Subroutines of the automatic tracking program which apply to the KR-010/KR-5400A combination and information on how to install them in the W0SL package is supplied with the KR-010.

The suggested retail price is \$269.95. For details, contact Encomm, Inc., 1506 Capital, Plano, Texas 75074.

Circle #302 on Reader Service Card.

new from RF Concepts

Ken Holladay, K6HCP, and Everett Gracey, WA6CBA, the co-founders of Mirex Communications, have formed RF Concepts, a new company whose products include a state-of-the-art, all-mode 2-meter, 170-watt VHF amplifier with a GaAsFET receive pre-amp; a 450 MHz 100-watt UHF amplifier with a GaAsFET receive pre-amp; a 130-watt 220-225 MHz amplifier with a GaAsFET pre-amp; and a 30-watt HT amplifier, also with a GaAsFET receive pre-amp. Other products include a repeater amplifier for 440 MHz, 220 MHz, and 144 MHz and a full-function repeater controller.

On the drawing boards is a new dual-band amplifier, 144/440 and base station amplifiers, and state-of-the-art switching power supplies from 10 to 35 amps.

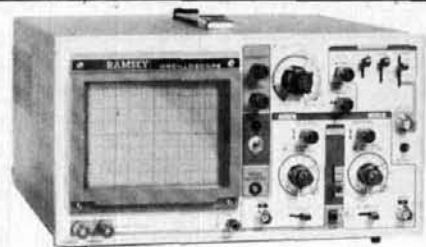
For further information, contact RF Concepts, 2000 Humboldt Street, Reno, Nevada 89509

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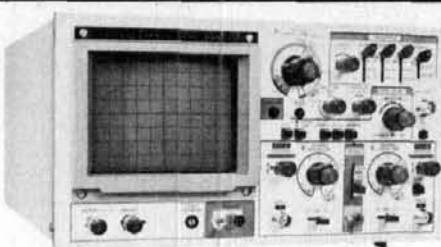
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2500	15 MHz	(2)	3.5 inch	2 mV per div	30 MHz	25 MHz
3500	35 MHz	(2)	8x10CM	1 mV per div	50 MHz	60 MHz

All include high quality 1:1, 10:1 hook on probes, instruction/service manual with schematic and component layout, 1 year warranty.

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CT-50 8 DIGIT 600 MHz



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CT-125 9 DIGIT 1.2 GHz



\$189.95 WIRED INCLUDES AC ADAPTER

MODEL	FREQ RANGE	SENSITIVITY	ACCURACY	DIGITS	RESOLUTION	PRICE
MINI-100	1-500 MHz	Less than 250mv	1 PPM	7	100 Hz, 1 KHz	119.95
CT-70	20 Hz-550 MHz	< 50mv to 150 MHz	1 PPM	7	1Hz, 10Hz, 100Hz	139.95
CT-90	10 Hz-600 MHz	< 10mv to 150 MHz < 150mv to 600 MHz	1 PPM	9	0.1Hz, 1Hz, 10Hz	169.95
CT-50	5 Hz-600 MHz	LESS THAN 25 mv	1 PPM	8	1Hz, 10Hz	189.95
CT-125	10 Hz-1.25 GHz	< 25mv @ 50 MHz < 15mv @ 500 MHz < 100 mv @ 800 MHz	1 PPM	9	0.1Hz, 1Hz, 10Hz	189.95
CT-90 WITH DV-1 OPTION	10 Hz-600 MHz	< 10mv to 150 MHz < 150mv to 600 MHz	0.1 PPM	9	0.1Hz, 1Hz, 10Hz	229.90

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 CONTACTS 5 PIN C
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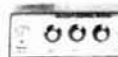
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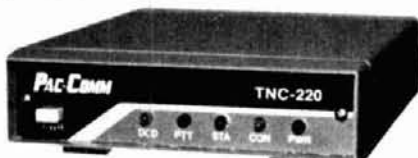
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RECONDITIONED TEST EQUIPMENT \$1.25 for catalog. Walzer, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: Please indicate in your announcements whether or not your Hamfest location, classes, exams, meetings, flea markets, etc., are wheelchair accessible. This information would be greatly appreciated by our brother/sister hams with limited physical ability.

MICHIGAN: April 5. The South Eastern Michigan Amateur Radio Association's 29th annual Hamfest Swap and Shop, 8 AM to 3 PM, Grosse Pointe North High School, 707 Vernier Road, Grosse Pointe Woods. ARRL, DX and Packet forums. Tickets \$1.00 advance; \$3.00 at the door. Tables \$8.00 advance; \$10.00 at the door. Talk in on SEMARA repeater 147.70/10 and 146.52 simplex. For information: SEMARA Hamfest, PO Box 646, St. Clair Shores, MI 48080 or phone Fred Lewis, NK8M (313) 881-0187.

MINNESOTA: April 4. 10th annual Rochester Area Hamfest sponsored by the Rochester ARC, John Adams Junior High School, 1525 NW 31st Street, Rochester. Doors open 8:30 AM. Large flea market, speakers, programs, refreshments and plenty of free parking. Talk in on 146.22/82. W0MXXW. For further information contact: RARC c/o W0BYEE, 2253 Nordic Ct. NW, Rochester, MN 55901.

NEW JERSEY area Chaverim, an organization of Jewish Amateur radio operators completing its 4th successful year, invites all hams and those interested in Amateur Radio to join, particularly those in Monmouth and Ocean counties of NJ. We meet the first Sunday of the month in the Jewish Community Center. Grant Avenue, Deal. Write POB 192, West Long Branch, NJ 07764.

OHIO: April 24, 25, 26. DAYTON HAMVENTION.

The Dayton Amateur Radio Association is now accepting applications for its 1987 Scholarship Program. Any licensed Amateur graduating from high school in 1987 is eligible to enter. For information and application forms write DARA Scholarship Committee, 317 Ernst Avenue, Dayton, Ohio 45405.

OHIO: April 24. The 18th annual B*A*S*H* will be held on FRIDAY NIGHT of the Hamvention, at the Conference Center (Madison Room) of the HARA ARENA AND CONFERENCE CENTER, (the same location as the Hamvention) starting at 7:00 PM. There is no admission charge, and free continuous entertainment. Hot dinner, sandwiches, snacks and beverages are available. Two exciting top awards, and many others. Stay right at HARA when the Hamvention closes on Friday evening and meet your friends and join us for an evening of fun and entertainment. Sponsored by the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401

OHIO: April 24. The Dayton-Cincinnati Chapter 9 of the QCWA announces the 1987 QCWA Banquet, Friday night of the Dayton Hamvention. Neil's Heritage House in Dayton. CODG bar at 6:30; dinner at 7:30. Tickets \$13.00 each. Contact Bob Ding, KA4LAU, 657 Dell Ridge Drive, Dayton, Ohio 45429. QCWA membership is not required to attend.

NEW JERSEY: May 9. The Cherryville Repeater Association's annual Hamfest, Hunterdon Central High School, Flemington, 8 AM to 4 PM. Indoor tables plus tailgating, FCC exams, forums displays and more. Talk in on 146.52, 147.975/.375. For further information or reservations call Bill Inkrote, K2NJ (201) 788-4080 or Don Mazak, NR2H (201) 782-1114.

COLORADO: June 20. The Grand Mesa Repeater Society will hold its 8th annual Western Slope Amateur Radio and Computer Swapfest, 9 AM to 4 PM, National Guard Armory, 482-28 Road, Grand Junction. Free admission. Swap tables \$5.00 each. Indoor swapfest, Amateur Radio exams, auction and refreshments. Talk in on 146.22/.82 and 449.20. For tables or information SASE to Les Scott, W0VF, 2105 Yellowstone Rd, Grand Junction, CO 81503 or call (303) 242-5296.

CALIFORNIA: May 1, 2, and 3: The Fresno Amateur Radio Club will hold its 45th annual Hamfest, Fresno Airport Holiday Inn. Air conditioned dealer spaces and swap tables. FCC exams. Good programs and parking. Demonstrations and forums. Talk in on 146.34-94. For information: Glen T. Caine, Fresno ARC, PO Box 783, Fresno, CA 93712 or (209) 292-4611.

INDIANA: April 4. The 4th annual Columbus Amateur Radio Club Swapfest, Bartholomew County Fairgrounds in Columbus, 8 AM to 5 PM. FCC testing. Advance setup Friday, April 3 with overnight security provided. Free parking. Talk in on 146.790 and 444.950. For information: Dave Mann, KA9UUP, 458N Country Club Rd, Columbus, IN 47201 (812) 342-6302

NEW YORK: May 3. The Suffolk County Radio Club's indoor/outdoor flea market, 8 AM to 2 PM, Republic Lodge No. 1987, 585 Broadhollow Road, Rt. 110, Melville, L.I. General admission \$3.00. Spouse and kids under 12 free. Indoor tables \$10. each. Outdoor spaces \$7.00 each. Includes one free admission. Talk in on 144.61/145.21 and 146.52 simplex. For information: Bill Sullivan, N2ETG (516) 689-9871 evenings.

TENNESSEE: April 5. The Clarksville Amateur Transmitting Society is sponsoring their annual Swapfest, National Guard Armory, Hwy 414, Clarksville. Tables \$5.00 each. Free admission. Talk in on 146.52 or 146.805-.600 or packet 145.03. For tables or information call WD4DBJ, Larry, (615) 232-6141.

NEW JERSEY: May 3. The Tri-County Radio Association's annual Indoor Hamfest/Flea Market, Passaic Township Community Center, Stirling, 9 AM to 3 PM. Donations \$3.00. Tables \$8.00; w/power \$10.00. Reserved tailgating. Refreshments. Talk in on 147.855/.255, 146.52 and 444.975/449.975. For information: Dick Franklin, W2EUF, POB 182, Westfield, NJ 07090. (201) 232-5955.

ARKANSAS: May 2. The Northwest Arkansas ARC will hold its 7th annual Hamfest, Rogers Youth Center, 315 West Olive Street, Rogers, 8 AM to 4 PM. Doors open 6 AM for exhibitors. General admission and parking free. Nearby recreational areas and parking for RV's and campers. Talk in on 16/76 or 03/63. For information: Roy Milliner, AF5W, 2014 S. 16th Street, Rogers, AR 72756 (501) 636-6750.

VIRGINIA: May 3. The Lynchburg ARC will hold its annual Swapfest, Brookville High School grounds, Route 460 West, just outside Lynchburg. Rain or shine. Starts 9 AM. Admission \$1.00. Tailgaters pay general admission plus \$2.00. VE exams beginning 1 PM. Please pre-register. For more information write Lynchburg ARC, PO Box 4242, Lynchburg, VA 24502.

OKLAHOMA: April 5. The Great Plains ARC's 6th annual Northwest Oklahoma Eyeball and Swapmeet, Mooreland, starting 9 AM. Admission \$2.00. Dealer and swap tables available at no charge. Covered dish dinner at noon. Local airport. VE tests Saturday, April 4. Campsites available. Talk in on 147.72/12, 146.13/73 and 146.52 simplex. For information write Gerald Bowman, WG5Z, Box 356 Mooreland, OK 73852 or call (405) 994-5600.

MINNESOTA: April 4. The 10th annual Rochester Area Hamfest, sponsored by the Rochester ARC, John Adams Junior High School, 1525 NW 31st Street, Rochester. Doors open 8:30 AM. Admission \$2.50 advance, \$3.00 at the door. Spouse and harmonics free admission. Tables \$8.00. Selling space \$4.00, bring own tables. License exams. Walk-ins only. Talk in on 146.22/82, W0MXXV. For information and advance tickets: RARC c/o W8OYEE, 2253 Nordic Ct NW, Rochester, MN 55901.

RHODE ISLAND: May 16. The RI Amateur FM Repeater Service will hold their annual Spring Flea Market and Auction, American Legion Fairmount Post 85, 870 River Street, Woonsocket. Flea market opens 9 AM and spaces are \$5.00 each. Auction from noon to 5 PM. Free admission. Food and beverages available. Talk in on 34/94 and 52 simplex. For information: Rick Fairweather, K1KYI, Box 591, Harrisville, RI 02830 or call (401) 56-8-0566, 7-9 PM.

MASSACHUSETTS: April 25. The Montachusett ARA will hold a flea market, Knights of Columbus Hall, Electric Avenue. 9:30 to 3 PM. Doors open 8 AM for sellers. Admission \$1.00. Tables \$8.00. Refreshments available. Talk in on 144.85/145.45 and 52. For reservations send check payable to MARA, c/o James Beauregard, 7 Mountain Avenue, Fitchburg, MA 01420.

SOUTH CAROLINA: May 2 and 3. The Blue Ridge Amateur Radio Society's 48th annual Greenville Hamfest and Electronic Flea Market, American Legion Fairgrounds, Greenville. Saturday 8 AM to 5 PM. Sunday 8 AM to 3 PM. Admission \$3.50 advance and \$5.00 at gate. Indoor/outdoor flea market, license exams, free parking, food. For tickets or information SASE to Blue Ridge ARA, PO Box 6751, Greenville, SC 29606.

NEW HAMPSHIRE: May 9. New England's favorite — The Hoss-traders Spring Tailgate Swapfest, Deerfield Fairgrounds. Admission \$2.00 per person including sellers and commercial dealers. Friday night camping at nominal fee only after 4 PM. Profits benefit Shriners' Burns Hospital. Last year's gift over \$13,500! For map, SASE to WA11VB, RFD Box 57, West Baldwin, ME 04091.

PENNSYLVANIA: May 3. The Delaware County ARA is sponsoring their 8th annual Hamfest, Drexel Hill Middle School, State Road and Penn Avenue, Drexel Hill. Doors open 8 AM. Setup 7 AM. Admission \$3. Reserved indoor tables/elect. available for \$3.00 per space. License exams. Refreshments available. Talk in on 147.96/36, 224.5 and 146.52. For registration and information write

Hamfest, DCARA, PO Bkox 236, Springfield, PA 19064 or call Barbara, N3DLG (215) 535-1616.

ONTARIO: April 11. The 6th annual Durham Region Flea Market sponsored by the South Pickering ARC, Pickering High School, Church Street N., Pickering Village, Ajax. Admission \$3.00. Children under 12 free. Tables \$6.00 each plus admission. For information: VE3DAX (416) 683-7562; VE3NBE (416) 839-9208; VE3JPP (416) 282-3983.

WEST VIRGINIA: April 5. Charleston WV Area Hamfest and Computer Show in Charleston Civic Center, 8 AM to 5 PM. Admission \$4.00. Forums, SSTV, Packet, Flea Market and Swap Tables. Computer demos, ARRL booth. Talk in on 6.22/6.88. For information: Ollis Rinehart, K8TIK, 1258 Ridge Drive, South Charleston, wV 25303. (304) 768-9534.

WISCONSIN: May 2. The Ozaukee Radio Club will sponsor its 8th annual Cedarburg Swapfest, Circle B Recreation Center, Highway 60 and County J, Cedarburg, 8 AM to 1 PM. Admission \$2.00 advance, \$3.00 at the door. 4 tables \$3.00 each. Food and refreshments available. For tickets, reservations or information SASE to 1987 ORC Swapfest, 101 E. Clay Street, Saukville, WI 53080 or phone (414) 284-3271.

TEXAS: April 18. The Austin ARC's Spring Swapfest, Mancha-ca Fire Station on FM 1626. Open 7 AM. Free admission and parking. Tables \$2.00. Talk in on 146.18/78. For info: Dave Harper, WD5N, 109 W. 38, Austin, TX 78705. Tel (512) 454-9206 evenings.

NEW JERSEY: May 17. The Bergen County ARA is sponsoring its Spring Hamfest, Bergen Community College, 400 Paramus Road, Paramus, 8 to 4 PM. Rain or shine. Buyers free. Sellers \$5 per space, tailgate only. License exams. Talk in on 146.19/79 and 520 simplex. For further information: Jim, KK2U (201) 445-2855, 444 Berkshire Rd, Ridgewood, NJ 07450.

WISCONSIN: April 5. The Madison Area Repeater Association, (M.A.R.A.) is please to announce its 15th annual Madison Swapfest, Dane County Exposition Center Forum Building, Madison. Doors open 7:30 AM for flea market sellers and 8 AM for the general public. An all-you-can-eat pancake breakfast available at the Swapfest. Admission \$2.50 advance and \$3.00 at the door. Children twelve and under admitted free. Flea market tables \$5.00 each/advance and \$6.00/door plus admission. The deadline for tickets and table reservations is March 31, 1987. For tickets, tables or information write M.A.R.A., PO Box 3403, Madison, WI 53704 or call (608) 274-5153 day or night.

OPERATING EVENTS

"Things to do . . ."

April 25. The Great River ARC of Dubuque, Iowa will operate N9GBY from 1600Z to 2300Z at the site of the annual Boy Scouts of America Grant Pilgrimage in Galena, Illinois. For a QSL card SASE to N9GBY, PO Box 141, Galena, IL 61036.

April 11. The St. Charles ARC will operate W0H0SI from 1400Z to 2200Z to celebrate its 15th anniversary near its original meeting site in St. Charles, MO. For certificates SASE to St. Charles ARC, PO Box 1429, St. Charles, MO 63302.

April 10, 11, 12. The Lanierland ARC will operate W4IKR 1400Z to 2200Z each day. To honor the 50th anniversary of the completion of the Appalachian trail, the 2000 mile hiking trail from Springer Mt. in Georgia to Mt. Katahdin in Maine. For commemorative certificate send QSL and 9x12 SASE (39 cents) to Lanierland ARC, PO Box 2182, Gainesville, GA 30503.

THE FOUNDATION FOR AMATEUR RADIO, INC. plans to award 26 scholarships for academic year 1987-88 to assist licensed Radio Amateurs who plan to pursue a full-time course of studies beyond high school and are enrolled or have been accepted by an accredited university, college or technical school. For further information write FAR Scholarships, 6903 Rhode Island Avenue, College Park, MD 20740 prior to May 31, 1987.

ARMED FORCES DAY 1987 May 16. The 38th annual Armed Forces Day Communication Test. Traditional military-to-Amateur cross band operation and broadcast of the Secretary of Defense message are the featured highlights and include operations in CW, SSB and RTTY.

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MRF450, A	Q 50W	14.00	31.00	
MRF453, A	Q 80W	15.00	35.00	
MRF454, A	Q 80W	15.00	34.00	
MRF455, A	Q 80W	12.00	28.00	
MRF458	80W	20.00	46.00	
MRF475	12W	3.00	9.00	
MRF476	3W	2.75	8.00	
MRF477	40W	11.00	25.00	
MRF479	15W	10.00	23.00	
MRF485*	15W	8.00	15.00	
MRF492	Q 90W	16.75	37.50	
SRF2072	Q 65W	13.00	30.00	
SRF3662	Q 110W	25.00	54.00	
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MRF238	30W	136-174	13.00	30.00
MRF239	30W	136-174	15.00	35.00
MRF240, A	40W	136-174	18.00	41.00
MRF245	80W	136-174	28.00	65.00
MRF247	75W	136-174	27.00	63.00
MRF607	1.75W	136-174	3.00	—
MRF641	15W	407-512	22.00	49.00
MRF644	25W	407-512	24.00	54.00
MRF646	40W	407-512	26.50	59.00
MRF648	60W	407-512	33.00	69.00
SD1441	150W	136-174	74.50	170.00
SD1447	100W	136-174	32.50	78.00
2N5591	25W	136-174	13.50	34.00
2N6080	4W	136-174	7.75	—
2N6081	15W	136-174	9.00	—
2N6082	25W	136-174	10.50	—
2N6083	30W	136-174	11.50	24.00
2N6084	40W	136-174	13.00	31.00

		MISC. TRANSISTORS & MODULES		
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MRF136	21.00	2N1522	10.50	
MRF136Y	70.00	2N3966	1.25	
MRF137	24.00	2N4048	10.50	
MRF138	35.00	2N4427	1.25	
MRF140	89.50	2N5590	10.00	
MRF148	35.00	2N5642	13.75	
MRF150	89.50	2N5643	15.00	
MRF172	82.00	2N5646	18.00	
MRF174	80.00	2N5945	10.00	
MRF208	11.50	2N5946	13.00	
MRF212	16.00	2SC2097	29.50	
MRF221	10.00	2SC2237	13.50	
MRF260	7.00	2SC1960	3.00	
MRF261	9.00	S10-12	13.50	
MRF262	9.00	SAV6	34.50	
MRF264	13.00	SAV7	34.50	
MRF406	14.50	SC1019	59.90	
MRF428	55.00	SC1027	47.50	
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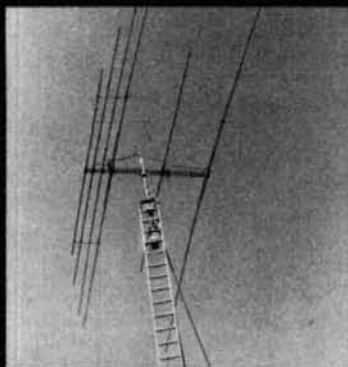
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READER SERVICE #	PAGE #	READER SERVICE #	PAGE #
130 - Advanced Computer Controls.....	99, 101, 103	194 - MFJ Enterprises	13
158 - Advanced Receiver Research	69	157 - Micro Control Specialties	62
153 - AEA	76	143 - Micro Systems Institute	92
111 - All Electronics Corp	110	172 - Midland Technologies	48
180 - Alpha Delta Communications	37	174 - Mirage Communications	51
* - Amateur Electronic Supply	96	184 - Mirage Communications	28
178 - Amateur Packet Alaska	43	188 - Mirage Communications	22
124 - Amateur Wholesale Electronics.....	104	173 - Mission Communications	48
108 - American Reliance, Inc.	112	162 - Missouri Radio Center	67
156 - AMSAT	62	136 - Motron Electronics.....	101
132 - Antennas Etc.....	103	110 - NCG	113
114 - Antique Electronic Supply	111	185 - Nel-Tech Labs, Inc.	26
200 - Antique Radio Classified	48	133 - Nema Electronics	101
169 - Arnold Co	48	140 - Nuts & Volts	95
150 - ARRL	83	* - Opportunity	106
154 - Astron Corp	73	187 - P.C. Electronics	23
* - Barker & Williamson	37	105 - Pac-Comm Packet Radio Systems, Inc.	112
* - Barry Electronics	111	127 - Pacific Rim Communications	104
166 - Bilal Company	54	170 - Palomar Engineers	48
144 - Buckmaster Publishing	92	113 - Processor Concepts	111
171 - Buckmaster Publishing	48	129 - The PX Shack	105
* - Butternut Electronics	100	151 - Radio Amateur Callbook	82
* - C.O.M.B.	43	182 - Radiosporting	38
* - Caddell Coil Corp	84	118 - Ramsey Electronics, Inc.	109
149 - California Packet Concepts	84	* - RF Parts	117
128 - Coaxial Dynamics, Inc.	105	164 - Robot Research, Inc.	59
198 - Communication Concepts, Inc.	87	160 - Roensch Microwave	69
103 - Communications Specialists	120	109 - S-Com	113
141 - CTM	94	167 - Satman	54
* - Dayton Hamvention	119	181 - Sias Engineering Co	38
119 - Digitrex	107	122 - Silicon Solutions	106
137 - Down East Microwave	99	104 - Sommer	118
121 - Eaglestone	107	* - Spec-Com	100
175 - EGE, Inc.	46	192 - Spectrum International	17
* - Engineering Consulting	92	116 - Spi-Ro Manufacturing Co	108
* - Falcon Communications	107	117 - Stone Mountain Engineering Co	108
161 - Fluke Manufacturing Co	69	125 - Stridsburg Engineering Co	104
186 - GLB Electronics	27	155 - STV/OnSat	72
165 - Grove Enterprises.....	54	135 - Synthetic Textiles, Inc.	101
138 - H.L. Heaster, Inc.	99	177 - TE Systems	43
193 - HAL Communications Corp.	18	189 - Tel-Com	20
183 - Hal-Tronix	32	145 - Telewave, Inc.	87
179 - Hall Electronics	43	106 - Testware	112
* - Ham-Com 1987	26	112 - Texas Radio Products	111
* - Ham Radio Magazine	99	191 - Torontel Technology Systems Ltd.	20
163 - Ham Radio Outlet	60, 61	201 - Total Electronic Concepts	115
* - Ham Radio's Bookstore.....	17, 42, 62, 72, 110, 111	199 - Transverters Unlimited.....	100
* - The Ham Station.....	42	148 - Unity Electronics	87
* - Hamtronics, NY	88	* - University Microfilm Int.	20
* - Hamtronics, PA	113	142 - Vanguard Labs	92
176 - ICOM America, Inc.	44, 45	195 - Varian	16
197 - ICOM America, Inc.	CII	159 - W9INN Antennas	69
139 - IIX Equipment Ltd.	99	123 - Western Electronics	106
107 - Interceptor Electronics, Inc.	112	102 - Yaesu Electronics Corp	CIII
120 - International Radio	107	168 - Yaesu Electronics Corp	52, 53
115 - Jun's Electronics	108	134 - E.H. Yost Co	101
146 - Kagil	87		
196 - Kantronics.....	1		
* - Trio-Kenwood Communications	2, 5, 7, 15, CIV		
152 - Madison Electronics Supply	80		
147 - Elaine Martin, Inc.	87		
190 - Glen Martin Engineering, Inc.	20		
131 - The Meadowlake Corp	103		
126 - Metal & Cable Corp.....	104		

PRODUCT REVIEW/NEW PRODUCTS

* - Design Electronics of Ohio	102
302 - Encomm, Inc.	108
303 - ICOM America, Inc.	106
301 - RF Concepts	108
* - Universal Electronics	102

*Please contact this advertiser directly.

Limit 15 inquiries per request.

Please use before May 31, 1987.

DAYTON **HAMVENTION**

April 24, 25, 26, 1987

Early Reservation Information

• General Chairman, Jim Simpson, WB8QZZ

• Asst. General Chairman, Bill McNabb, WD8SAY

Grand banquet tickets are limited, please place your reservations early.

- **Giant 3 day flea market • Exhibits**
- **Door prizes • License exams**
- **CW proficiency test**

Flea Market Tickets

We increased Flea Market area by nearly 400 spaces this year and all were sold out by January 10.

Special Awards

Nominations are requested for "Radio Amateur of the Year", "Special Achievement" and "Technical Achievement" awards. Contact: Awards chairman, Box 44, Dayton, OH 45401.

License Exam

Novice thru extra exams scheduled Saturday & Sunday by appointment only. Send current FCC form 610, copy of present license and check for \$4.35 (payable to ARRL/VEC) to: Exam Registration, 8830 Windbluff Point, Dayton Oh 45459

Slide Show

35 mm slide/tape presentation about the HAMVENTION is available for loan. Contact Dick Miller 2853 La Cresta, Beavercreek, OH 45324

Parking

Free parking is available at Hara Arena. In addition, there will be free shuttle bus service from all major motels and designated parking lots. Parking and road information will be available on DARA's 146.34/.94 repeater.

Free Bus Service

Free Bus Service will be provided between many Motels and Hara Arena. See the schedules at the motel registration desks. Avoid parking problems at the Arena by taking the HAMVENTION buses.

Campers & Trailers

Campers and Trailers may be parked at Montgomery County Joint Vocational School. A HAMVENTION bus will provide transportation between the camper parking area and the Arena. No campers or travel trailers will be permitted to park in the Arena lot or Flea Market area.

Wheelchairs

Wheelchairs will be available. Send S.A.S.E. for details to "Wheelchair" P.O. Box 44, Dayton, OH 45401.

Alternate Activities

HAMVENTION is for everyone. We have planned activities for the YL or your non-ham family members.

Deadlines

Award Nominations: April 4
Lodging: April 4
License Exams: March 28
Advance Registration and banquet: USA - April 11
Canada - April 4

Information

General Information: (513) 433-7720
or DARA Box 44 Dayton, OH 45401
Flea Market Information: (513) 223-0923
Lodging Information: (513) 223-2612
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This is the year for you to attend the internationally famous Dayton HAMVENTION. Come with your friends to hear enlightening forums, see the latest equipment, and visit a flea market that has everything! No matter what you are looking for, you can find it in Dayton!

HAMVENTION is sponsored by the Dayton Amateur Radio Association Inc.



Food for thought.

Our new Universal Tone Encoder lends its versatility to all tastes. The menu includes all CTCSS, as well as Burst Tones, Touch Tones, and Test Tones. No counter or test equipment required to set frequency—just dial it in. While traveling, use it on your Amateur transceiver to access tone operated systems, or in your service van to check out your customers' repeaters; also, as a piece of test equipment to modulate your Service Monitor or signal generator. It can even operate off an internal nine volt battery, and is available for one day delivery, backed by our one year warranty.

- All tones in Group A and Group B are included.
- Output level flat to within 1.5db over entire range selected.
- Separate level adjust pots and output connections for each tone Group.
- Immune to RF
- Powered by 6-30vdc, unregulated at 8 ma.
- Low impedance, low distortion, adjustable sinewave output, 5v peak-to-peak
- Instant start-up.
- Off position for no tone output.
- Reverse polarity protection built-in.

Group A

67.0 XZ	91.5 ZZ	118.8 2B	156.7 5A
71.9 XA	94.8 ZA	123.0 3Z	162.2 5B
74.4 WA	97.4 ZB	127.3 3A	167.9 6Z
77.0 XB	100.0 1Z	131.8 3B	173.8 6A
79.7 SP	103.5 1A	136.5 4Z	179.9 6B
82.5 YZ	107.2 1B	141.3 4A	186.2 7Z
85.4 YA	110.9 2Z	146.2 4B	192.8 7A
88.5 YB	114.8 2A	151.4 5Z	203.5 M1

- Frequency accuracy, $\pm .1$ Hz maximum - 40°C to + 85°C
- Frequencies to 250 Hz available on special order
- Continuous tone

Group B

TEST-TONES:	TOUCH-TONES:	BURST TONES:			
600	697 1209	1600	1850	2150	2400
1000	770 1336	1650	1900	2200	2450
1500	852 1477	1700	1950	2250	2500
2175	941 1633	1750	2000	2300	2550
2805		1800	2100	2350	

- Frequency accuracy, ± 1 Hz maximum - 40°C to + 85°C
- Tone length approximately 300 ms. May be lengthened, shortened or eliminated by changing value of resistor

Model TE-64 \$79.95

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One of the most complex operating controls of our high-performance mobiles.



You don't have to sacrifice performance to gain simplicity in your mobile operation.

Yaesu's 2-meter FT-211RH and 440-MHz FT-711RH give you all the performance you look for in a sophisticated, microprocessor-controlled mobile.

With controls that couldn't be more straightforward and easy to learn. Which means no

radios are based on the very same technology.

To begin with, you get an autodialer mic with 10 lithium backed memories, each capable of storing any key sequence up to 22 digits long.

Plus you get: 45 watts output (35 watts on 440 MHz). LCD readout. 10 memories that store frequency, offset and PL tone.

(7 memories can store odd splits.) Scan all memories or selected memories at 2 frequencies per second. Band scan at 10 frequencies per second. Tx offset storage. Priority channel scan.

Tuning via tuning knob, or up/down buttons. PL tone board (optional). PL display.

Independent PL memory per channel. PL encode *and* decode. LCD power output and "S"-meter display. Eight-key control pad. Keypad lock. High/low power switch (low power: 5 watts VHF, 3 watts UHF).

What's more, each radio is perfect for overhead mounting. Just remove a few screws and flip the control panel 180°

Discover the 2-meter FT-211RH and 440-MHz FT-711RH at your nearest Yaesu dealer today. If you can turn a knob and push a button, you'll have high-performance mobile operation mastered.

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operating complexities to interfere with your driving.

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TS-440S Compact high performance HF transceiver with general coverage receiver

Kenwood's advanced digital know-how brings Amateurs world-wide "big-rig" performance in a compact package. We call it "Digital DX-citement"—that special feeling you get every time you turn the power on!

Covers All Amateur bands

General coverage receiver tunes from 100 kHz—30 MHz. Easily modified for HF MARS operation.

Direct keyboard entry of frequency

All modes built-in USB, LSB, CW, AM, FM, and AFSK. Mode selection is verified in Morse Code.

Built-in automatic antenna tuner (optional)

Covers 80-10 meters.

VS-1 voice synthesizer (optional)

Superior receiver dynamic range

Kenwood DynaMix™ high sensitivity direct mixing system ensures true 102 dB receiver dynamic range. (500 Hz bandwidth on 20 m)

100% duty cycle transmitter

Super efficient cooling permits continuous key-down for periods exceeding one hour. RF input power is rated at 200 W PEP on SSB, 200 W DC on CW, AFSK, FM, and 110 W DC AM. (The PS-50 power supply is needed for continuous duty.)

Adjustable dial torque

100 memory channels
Frequency and mode may be stored in 10 groups of 10 channels each. Split frequencies may be stored in 10 channels for repeater operation.

TU-8 CTCSS unit (optional)

Subtone is memorized when TU-8 is installed.

Superb interference reduction

IF shift, tuneable notch filter, noise blanker, all-mode squelch, RF attenuator, RIT/XIT, and optional filters fight QRM.

MC-43S UP/DOWN mic. included

Computer interface port

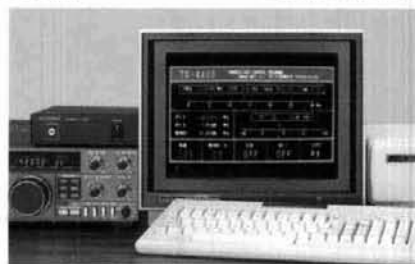
5 IF filter functions

Dual SSB IF filtering

A built-in SSB filter is standard. When an optional SSB filter (YK-88S or YK-88SN) is installed, dual filtering is provided.

VOX, full or semi break-in CW

AMTOR compatible



Optional accessories:

- AT-440 internal auto. antenna tuner (80 m—10 m)
- AT-250 external auto. tuner (160 m—10 m)
- AT-130 compact mobile antenna tuner (160 m—10 m)
- IF-232C/IC-10 level translator and modem IC kit
- PS-50 heavy duty power supply
- PS-430/PS-30 DC power supply
- SP-430 external speaker
- MB-430 mobile mounting bracket
- YK-88C/88CN 500 Hz/270 Hz CW filters
- YK-88S/88SN 2.4 kHz/1.8 kHz SSB filters
- MC-60A/80/85 desk microphones
- MC-55 (8P) mobile microphone
- HS-5/6/7 headphones
- SP-40/50B mobile speakers
- MA-5/VP-1 HF 5 band mobile helical antenna and bumper mount
- TL-922A 2 kw PEP linear amplifier
- SM-220 station monitor
- VS-1 voice synthesizer
- SW-100A/200A/2000 SWR/power meters
- TU-8 CTCSS tone unit
- PG-2S extra DC cable.

Kenwood takes you from HF to OSCAR!



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

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