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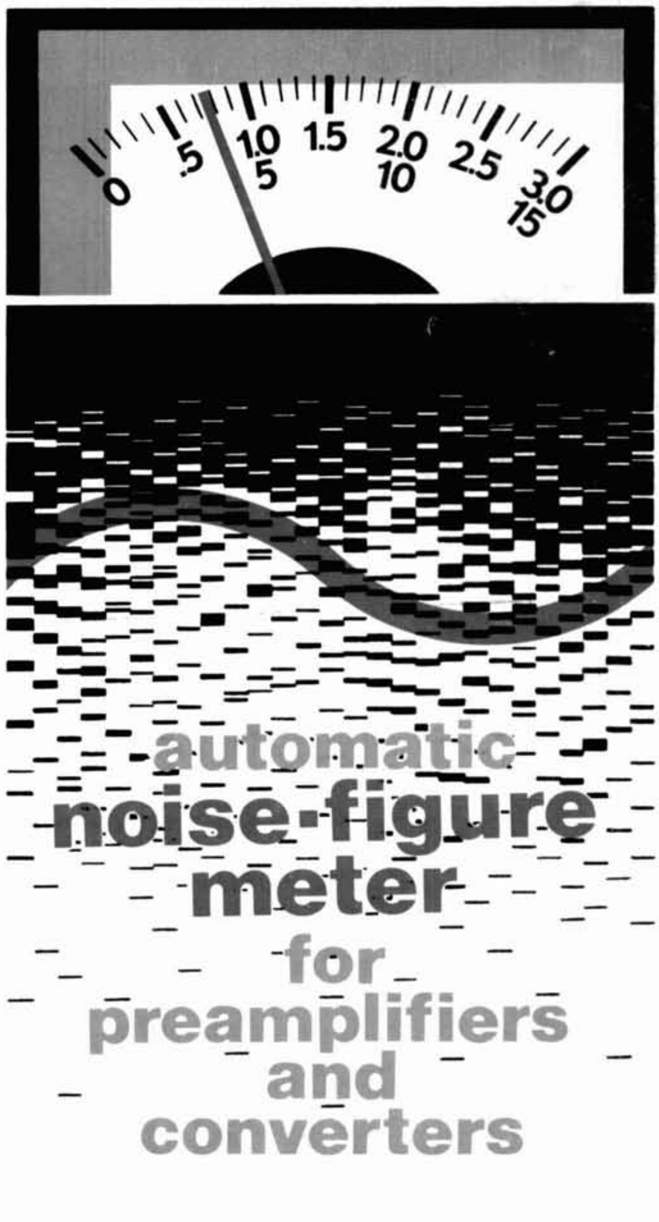
ham radio

magazine

FEBRUARY 1981



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Tempo was the first with a synthesized hand held for amateur use, first with a 220 MHz synthesized hand held, first with a 5 watt output synthesized hand held...and once again first in the 440 MHz range with the S-4, a fully synthesized hand held radio. Not only does Tempo offer the broadest line of synthesized hand helds, but its standards of reliability are unsurpassed...reliability proven through millions of hours of operation. No other hand held has been so thoroughly field tested, is so simple to operate or offers so much value. The Tempo S-4 offers the opportunity to get on 440 MHz from where ever you may be. With the addition of a touch tone pad and matching power amplifier its versatility is also unsurpassed.

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With 12 button touch tone pad...$399.00
With 16 button touch tone pad...$419.00
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13.8 VDC power amplifier...$149.00

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More Details? CHECK—OFF Page 98

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SANTEC HT-1200	YAESU FT-207R	KENWOOD TR-2400
Texas Instruments TMS-1000 microprocessor	NEC-650	NEC-650
Rx on 143 to 149.995 MHz Tx on 143 to 148.995 MHz (1200 channels with MARS coverage)	Rx & Tx on 144 to 147.995 MHz, Ham band only (800 channels)	Rx & Tx on 143.9 to 148.495 (900 channels with some MARS coverage)
Direct keyboard entry of all frequencies. Keyboard entry of 5kHz digit which stays in memory'	Keyboard enrty of 10kHz steps with a switch for 5kHz steps	Direct keyboard entry of Ham band only. MARS frequencies must be entered into a memory by stepping and recalling.
10 programable memories with frequencies preloaded on cold boot.	5 programable memories. All memories loaded with 144.00 on cold boot.	10 programable memories. All memories loaded with 145.00 on cold boot.
Up/Down variable scan steps in any multiples of 5kHz over whole band or auto-scan of 10 memories. Scan (restart) or search (lock) modes for both band and memory modes.	Up/Down scan with 10kHz steps only. Misses every other 15kHz by 5kHz. Locks without restart.	Scans 10 memories only. Restart only: lock mode not available. Continuous band scan/search not available.
Full 16 button TTP with LED display of number as it is dialed.	12 button TTP only.	Full 16 button TTP. Readout of the number dialed is not available.
9.6v 500mah battery (included)	10.8v 450mah battery (included)	9.6v 500mah battery (included)
Tx High: 3.5W (4W nominal) Tx Low: 1W	Tx High: 2.5W Tx Low Tx Low: 200mW	Tx at 1.5W only.
Readout: LED	Readout: LED	Readout: LCD
Volume: 543cc 170mm(H) x 68mm(W) x 47mm(D)	Volume: 664cc 181mm(H) x 68mm(W) x 54mm(D)	Volume: 640cc 192mm(H) x 71mm(W) x 47mm(D)

The Santec HT-1200 is approved under FCC Part 15 and exceeds FCC regulations limiting spurious emissions.

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ham radio

magazine

incorporating

HAM RADIO HORIZONS

FEBRUARY 1981

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T. H. Tenney, Jr., W1NLB
publisher and editor in chief

Alfred Wilson, W6NIF
editor

editorial staff

Martin Hanft, WB1CHQ
production editor

Joseph J. Schroeder, W9JUV
Leonard H. Anderson
associate editors

W.E. Scarborough, Jr., KA1DXQ
graphic production manager

Irene Hollingsworth
editorial assistant

Wayne Pierce, K3SUK
cover

publishing staff

J. Craig Clark, Jr., N1ACH
assistant publisher and advertising manager

Susan Shorrock
circulation manager

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contents

Observation & Opinion

The urge to compete seems to be a part of human nature. Amateur Radio has its share of competition: DXCC, T-Hunts, Field Day. Competition stimulates people to improve, and that's healthy.

Another contest, which was popular years ago but which hasn't received much publicity recently, is the world high-speed CW championship. This is the challenge of challenges for operating skill: to break the record of Ted McElroy, ex-W1JYN, who made the *Guinness Book of World Records* by copying Morse code at a speed of 75.2 words per minute in a contest at Ashville, North Carolina, on July 2, 1939. Ted's record still stands. It's time for someone to try to break it.

Many readers will sniff in disdain at such a contest: "Who needs it?" "What will it prove?" I'm here to tell these people that Morse code is here to stay, like it or not.

Since I became editor of *ham radio*, I've received many letters from readers who scoff at the Morse code requirement in the Amateur license examination. For the most part, their reasons are that Morse isn't necessary for today's communications. I won't argue this point except to say that these people are somewhat misguided and know not whereof they speak. Listen to the parts of the Amateur bands devoted to traffic handling, the extra-class sub-bands, the Novice bands. It's CW, in whatever form.

A contest to break the world's record in Morse code reception is challenging, exciting, and in the best tradition of Amateur Radio for those who like to compete. Ted McElroy's record has been unbroken for 42 years. Who will be the next champ?

We at *ham radio* are proposing a contest for those who wish to try to break Ted's record. The contest will be conducted under official rules similar to those in effect during the contest in Ashville, North Carolina, in 1939. Appropriate prizes will be awarded to the winners.

Right now the contest is in the planning stage. The first contest will probably be held at one of the larger ham conventions in the spring of this year. We haven't yet decided which convention it will be. At any rate, the contest will be held under strictly controlled conditions, in a room devoid of distractions and noise. A contest of this sort must be done by the rules to guarantee fairness to all.

More information on the contest will be upcoming in future issues of both *ham radio* and *HR Report*. Look for it and plan to enter. CW is certainly not dead!

Our ex-*Horizons* readers will find some familiar topics in this issue: an equipment owner's questionnaire on three popular transceivers, Bill Orr's column on "Ham Radio Techniques," and of course Garth Stonehocker's "DX Forecaster." Next month we'll have the results of the Collins KWM2 and KWM2A equipment survey. Also look for the popular Q and A column and an interesting article by W7JWJ on the world's champion high-speed Morse code operator.

Alf Wilson, W6NIF
editor

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ICOM

comments

hyperbolic navigation

Dear HR:

Amateur Radio experts often reinvent the wheel in novel ways. A case in point is the article in *ham radio*, September, 1980, by Henry S. Keen, W5TRS, on "Navigational Aid for Small Boat Operators." The whole idea here is basically a hyperbolic navigation method and is markedly similar to the DECCA system used in Europe for many decades. DECCA operates on harmonic-related frequencies in the 75 kHz to 150 kHz range and uses exact integral divisors or multipliers as the intermediate beat frequency, very similar to Keen's idea of the second harmonic or subharmonic of the beat note. A problem that Keen has not addressed is that of "lane jumping," or resolution. At transmitter frequencies in the 10-meter range, the lane width between adjacent hyperbolas will be 5 meters, which taxes the resolution of the phase detector (and the steering navigator) for keeping track of which hyperbola the navigator is seeking.

At short ranges (a few miles), the use of the 1-watt-limit, 1750-meter experimenter band at 160 to 190 kHz might possibly make Keen's idea practical with lane widths of 1/2 wavelength or 875 meters for one cycle of phase change of the beat note. Thus 875 meters becomes 360 degrees of the phase detector output. Typical phase detectors resolve to 1 per cent or so at audio beat frequencies, so in practice 8.75-meter resolution of the path over a particular hyperbola might be resolved.

Another system is available from Hastings-Raydist Division of Teledyne Corporation. This has used transmitters in the 500 kHz to 2.5 MHz range, with hyperbolas 100 meters apart and resolution to within 1 meter. Still other worldwide hyperbolic navigation systems like OMEGA (10.2 to 13.6 kHz) and LORAN-C (100 kHz) use time-sequenced bursts or pulses to enable a large number of stations to all transmit on the same frequency and very effectively measure phase difference by the time of arrival or sequential memory-aided phase-locked loop methods. The very high power (1 megawatt) Navy vlf communications transmitters operating in the 14 kHz to 20 kHz range have also been used for worldwide navigation. Here, they are all atomic clock controlled, so all one has to do in principle is to convert the receiver measurement to a common i-f in the audio range at something like 100 Hz.

There are a great many pitfalls in devising new navigation systems that have been thoroughly worked over in the past 50 years. A general reference on the subject is Kayton and Fried *Avionics Navigation Systems*, John Wiley & Sons, published in 1969. Another reference, particularly on the early history of hyperbolic navigation, is NBS Monograph #129, "The Development of LORAN-C Navigation and Timing," U.S. Government Printing Office, $4.50, published October, 1972. A major problem with any CW-type hyperbolic system is that of proper lane identification; that is, how does the navigator know where to start the phase measurement, which lane or which line of position is he on? The SHORAN system that Keen mentions is not a CW hyperbolic method but rather a direct ranging time of arrival of pulse technique operating in the 200-300 MHz range. CW hyperbolic systems have largely been replaced by direct ranging ideas using coherent phase-locked transponders, but this always makes the system hardware more complex. When engineers think about these methods, what invariably happens is that the actual working hardware becomes much more complex and expensive than the inventor originally intended. Also, it is very hard these days to come up with something new. There are so many people thinking about the same idea at any given time that ten people will come up with the same idea at once, and there is just too much prior art to study.

Ralph W. Burhans
Athens, Ohio

RST

Dear HR:

I would like to make a comment to the "Observations & Comments," *ham radio*, September, 1980. I wholeheartedly agree with the idea of changing the RST system to something like Q1-Q2-Q3 as reporting signal conditions.

I am not a DXer or contest man, but I have put in some time logging on Field Day. It sure looks ridiculous to see the whole log with just RST 599. The RST system doesn't give a true picture of conditions. I like the Q1-Q2-Q3 because it seems to cover the total spectrum. But as it is done now, the whole contest report could be just all Q3s. Just for the record I think I will start using this system of Q1,2,3 and wait and see how many ask about it. It is worth a try.

George F. Schmidt, WØUCK
St. Louis, Missouri

More Details? CHECK—OFF Page 98

presstop

PLAIN LANGUAGE AMATEUR RULES docket released in the form of a Notice of Proposed Rule Making, PR Docket 80-279, includes both the texts of the present Part 97 rules and the proposed rewrite, plus explanations. It was printed in the Federal Register of December 19.

Significant Changes, such as the deletion of all logging requirements and addition of a requirement that licensees must keep a copy of the Amateur rules on hand, also include a warning that the FCC can inspect a station "at any time during the business day or any time your station is transmitting or has just finished transmitting." The new rules no longer tell an Amateur how he should make various measurements, but only how the FCC inspector would make them during a station visit. The proposed new rules also clarify rules on interconnects (phone patches), antenna heights, and emergency and prohibited communications. They also include much more information on exams, both written and code, and propose that applicants with FCC Commercial CW licenses get credit for the equivalent Amateur CW exam. The rewrite also endeavors to consolidate and reorganize the rules into a more concise and logical arrangement.

In Addition To Being Printed in December 19's Federal Register, PR Docket 80-729 is also available in limited quantities from the FCC Office of Consumer Assistance, Room 258, 1919 M Street NW, Washington, D.C. 20554. Comments won't be due until June 19, 1981, and Reply Comments on August 19, 1981.

ITALIAN EARTHQUAKE AREA COMMUNICATIONS depended very heavily on Amateur Radio until telephone and other services were partially restored in mid December, with Amateur Radio still providing a major channel for relief traffic. Initial traffic from the disaster area included casualty lists, relayed from VHF links in the area by I8ULL near Naples. Many stations throughout the U.S. and Canada participated, with KA1BQ and WB2FID leading the action on 28802 mornings and 14240 later in the day when conditions permitted.

The Need For Transatlantic Relief communications links should continue for at least several more months, because of the extent of the damage (over 26,000 square kilometers devastated) and the magnitude of the U.S.-originated relief effort. KA1BQ (I8CZW) left for Italy last month to work on the scene and attempt to convince the Italian government to continue its third party OK.

I8ULL, Principal Terminal for U.S./Italy traffic, has been shut down several times by aftershock damage. KLM donated a replacement beam, which Alitalia shipped to Naples, to that station.

Participants In Relief communications represent all ethnic backgrounds, KA1BQ noted, a graphic demonstration of the fraternal character of the Amateur Radio service.

30-METER BAND OPERATION was begun in early December by VE3OB, using his Canadian commercial callsign, VE9LFZ, for preliminary one-way transmissions with partner VE9LIN (VE3BPD). They plan to operate daily at 1500, 1700, and 1900Z, plus random times evenings starting at 2200Z. CW and (later) ASCII will be the modes, using 10.101 and 10.149 MHz.

VE9LFZ Was Solid Copy on 10.101 MHz in the Midwest Thanksgiving afternoon, with a more than hour-long transmission that peaked as high as 5-6 even though Larry was running only 5 watts to an inverted V. These are test transmissions only, authorized by the Canadian government, and Amateur operations on this band will be illegal until January, 1982.

CALIFORNIA BRUSH FIRES, which destroyed about 350 homes and burned hundreds of thousands of acres in December, brought out plenty of volunteer Amateur Radio support. More than 30,000 acres burned east of El Toro, where more than 50 operators used W6TIO/R and simplex channels to provide Red Cross, paramedics, and fire headquarters with needed radio services. A link with the National Traffic System was also set up providing health-and-welfare communications for the fire fighters, who came in from a seven-state area.

Another 150 Or So Amateurs served in San Bernardino/Riverside areas, where, along with many CBers, they provided communications for the Lake Arrowhead Evacuation Center.

Another Fire In The Big Basin Redwood State Park destroyed 350 acres of young redwoods there. Sixty-five-year-old K6TEH set up his portable 34-94 repeater there at the request of the Division of Forestry. A dozen others also took part.

PHASE III OF FCC'S CALLSIGN assignment system went into effect December 15, and with it all license classes except Novice became eligible to request callsign changes. Any Amateur who now holds a callsign not appropriate for his license class (1x3 for General/Technician, 2x2 for Advanced, 1x2/2x1 for Extra) may request a change to a callsign of the proper format, though without any choice as to the specific callsign he will receive.

FCC's New Form 610 (August, 1980 edition) is required by Phase III, so all previous editions are now obsolete and may not be used. As before, licensees still have the option of retaining their old callsigns when upgrading or changing call areas. A new callsign will not be assigned unless it's specifically requested by the licensee's having checked item 2F on the new Form 610.

Though Both The 4th and 6th call areas are running low on 2x1 callsigns for Extras, the Commission does not expect to be ready to go back to 1x2s until 1983 at the earliest.

More Details? CHECK—OFF Page 98

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D-20	20	33	24.95	20.95
D-15	15	22	23.95	19.95
D-10	10	16	22.95	18.95
Shortened dipoles				
SD-80	80/75	90	31.95	27.95
SD-40	40	45	28.95	24.95
Parallel dipoles				
PD-8010	80,40,20,10/15	130	39.95	35.95
PD-4010	40,20,10/15	66	33.95	29.95
PD-8040	80,40/15	130	35.95	31.95
PD-4020	40,20/15	66	29.95	25.95
Dipole shorteners - only, same as included in SD models				
S-80	80/75		$11.95 pr	
S-40	40		$10.95 pr	

All antennas are complete with a Hi-Q Balun or Hi-Q Antenna Center Insulator, No. 14 antenna wire, ceramic insulators, 100' nylon antenna support rope (SD models only 50'), rated for full legal power. Antennas may be used as an inverted V and may also be used by MARS or SWLs.

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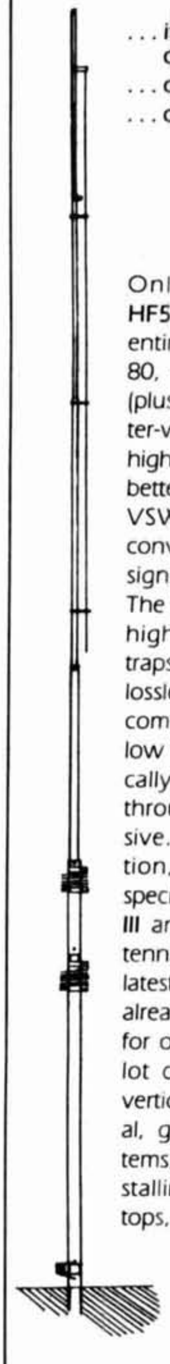

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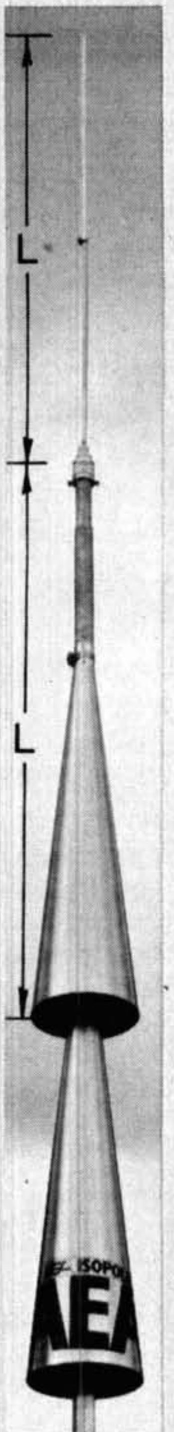

The IsoPole is building a strong reputation for quality in design and superior performance. The IsoPole's acceptance has already compelled another large antenna producer to make a major design modification to his most popular VHF Base Station antenna. Innovative IsoPole conical sleeve decouplers (pat. pend.) offer many **new** design advantages.

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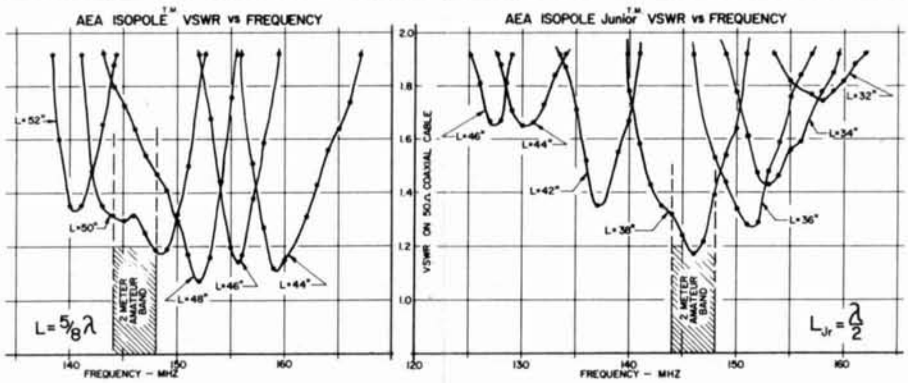

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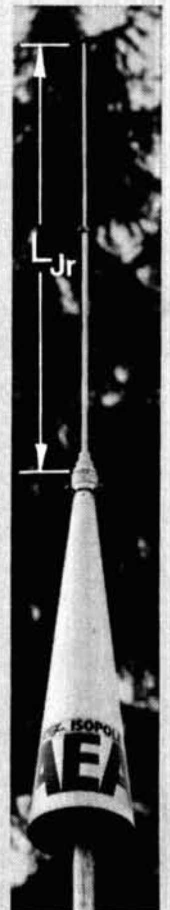

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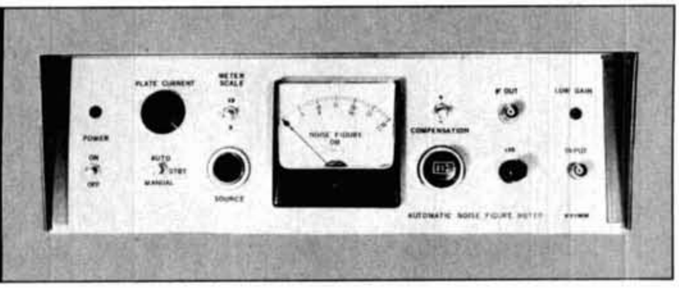


automatic noise-figure meter

for preamplifiers and converters

A recent article by W6NBI[1] describes an easily duplicated temperature-limited diode noise generator and power supply. This generator provides both a reproducible source of noise and, by measuring the diode plate current, an easy method to measure the noise power output of the source. As will be discussed, verification of a receiver's noise figure by the use of this generator is straightforward, using either the y-factor or twice-power methods.

Optimization of the noise figure of a receiver is quite tedious because each time a parameter is varied, two measurements (insertion of a pad and some calculations) are required. Also, since the two measurements are taken at different times, any gain drift in the system will introduce error into the measurement. If you are willing to restrict the noise source to a temperature-limited diode, it's possible to build a very accurate automatic noise-filter meter that avoids some of the compromises multisource meters must make to accommodate different sources.

derivation of the noise-figure concept

Noise factor (F) of a receiver is defined as

$$F = \frac{S/N_I(Ideal\ Receiver)}{S/N_A(Actual\ Receiver)} = \frac{N_A}{N_I}$$

$$= \frac{Noise\ Power\ Actual\ Receiver}{Noise\ Power\ Ideal\ Receiver} \quad (1)$$

and in logarithmic terms

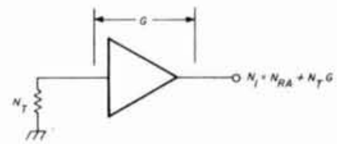
$$Noise\ Figure\ (NF) = 10 \log F \quad (2)$$

It's important to keep the *log NF* terms and the *numerical ratio F* terms separate in your mind; if you mix them up you'll make errors when calculating correction factors.

If we call N_I the noise output power from the receiver when its input is terminated in a 50-ohm resistor at a temperature of 290 K (62 F), then

$$N_I = N_{RA} + N_T G \quad (3)$$

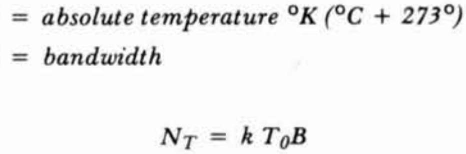
where N_{RA} = *receiver added noise power*
N_T = *termination resistor noise power*
G = *receiver gain*



If we then apply an excess noise to the receiver input we can define:

$$N_2 = N_{RA} + N_T G + (EN)G \quad (4)$$

where EN = *excess noise power.*



Noise power $(N) = kTB$

and k = *Boltzman's constant* $1.374 \times 10^{-23} \frac{Joules}{°K}$

T = *absolute temperature* °K (°C + 273°)

B = *bandwidth*

then

$$N_T = k T_0 B \quad (5)$$

where T_0 = *290 K (62 F) (by IEEE Convention)*

$$EN = kTB \quad (6)$$

where T = *effective temperature of the excess noise source in °K.*

The definition of the noise factor of a receiver (**eq. 1**) is

$$F = \frac{N_A}{N_I} = \frac{N_I}{kT_0BG}$$

By Edward T. Gisske, K9IMM, 7256 Mineral Point Road, Verona, Wisconsin 53593

The amount of noise added by the receiver is

$$\frac{N_2}{N_1} = \frac{kT_0BG + (F-1)\,(kT_0BG) + kTBG}{kT_0BG + (F-1)\,(kT_0BG)}^{*}$$

$$= \frac{FT_0 + T - T_0}{FT_0}$$

Rearranging

$$F = \left(\frac{T-T_0}{T_0}\right)\left(\frac{1}{\frac{N_2}{N_1} - 1}\right) \quad \textbf{(7)}$$

or, in dB

$$NF = 10\log\left(\frac{T-T_0}{T_0}\right) - 10\log\left(\frac{N_2}{N_1} - 1\right) \quad \textbf{(8)}$$

The first term is equal to the excess noise ratio (ENR) of the source expressed in dB

$$NF = ENR(dB) - 10\log\left(\frac{N_2}{N_1} - 1\right) \quad \textbf{(9)}$$

The term N_2/N_1 is referred to as the y-factor and is used by automatic noise-figure meters such as the HP-340 series to measure noise figure in the following manner:[2]

With reference to **fig. 1**, note that the source is gated on and N_2 is measured; then the source is gated off and N_1 is measured. This sequence is repeated at about a 500-Hz repetition rate. The ENR of the source is known, and N_2 is used to set the gain of the i-f amplifier so that N_2 is a fixed output regardless of the receiver gain. Then, by measuring N_1, the ratio N_2/N_1 is known. The meter is calibrated to solve the equation:

$$NF = ENR - 10\log\left(\frac{N_2}{N_1} - 1\right) \quad \textbf{(10)}$$

This ratiometric measurement factors receiver gain out of the measurement, so it's possible to tune for minimum noise figure without recalibration. Note that the i-f amplifier agc loop must be very tight, as the measurement depends on maintaining a constant reference (N_2) level. The i-f amplifier and detection circuitry must also be linear over a range of agc levels to get a valid N_1 measurement.

Another noise-figure measuring method, usually used for manual measurements, is called the twice-power method. This method forces $N_2 = 2N_1$ by adjustment of the source ENR.

$$NF = ENR(dB) - 10\log\left(\frac{N_2}{N_1} - 1\right)$$

$$N_2 = 2N_1$$

$$= ENR - 10\log\left(\frac{2}{1} - 1\right)$$

$$NF = ENR \quad \textbf{(11)}$$

Fig. 2 shows a block diagram of the measurement. In practice, the measurement is performed as follows: with the noise source off and the pad **out** of the circuit; the power meter reading is noted. The source is then activated and the 3-dB pad inserted into the circuit. The source ENR is adjusted to produce exactly the same reading as before. When this condition is satisfied, the ENR of the source is equivalent to the receiver noise figure. The ENR of the source is usually adjusted by padding the source with a precision variable attenuator.

The attenuator setting is subtracted from the known ENR of the source to compute the receiver noise figure. This works for temperature-limited diode, argon discharge, or semiconductor diode souces. With the temperature-limited diode, an additional method of adjusting source ENR is to adjust the diode plate current (by varying the filament current) and thus vary the effective source temperature. Coincidentally, in a 50-ohm system, the noise factor $F \cong \frac{\text{diode plate current}}{1000}$ and noise figure, $NF \cong 10 \log I(mA)$. (See reference 1 for a derivation of this relationship.) The advantage of the twice-power method over the y-factor method is that the power-measuring circuit is always operated at the same signal level; therefore linearity of the amplifiers and de-

*The transition from **eq. 6** to **eq. 7** may not be obvious to some readers, in view of the appearance of the term $(F-1)$. The equation for N_2/N_1 is correct, however, as the following shows:

By definition, the noise temperature, T, of the input terminating resistor is related to the equivalent noise factor, F, contributed by the terminating resistor by the relationship

$$T = T_0(F-1)$$

rearranging

$$\frac{T}{T_0} = (F-1)$$

therefore

$$(F-1)\,(KT_0BG) = \frac{T}{T_0}\,(KT_0BG) = KTBG$$

Note that if $T = T_0$, $(F-1) = 1 \cdot (F-1)$ accounts for the noise power at the amplifier output due to the input terminating resistor operating at some temperature $T°K$. When $T \neq T_0$, then a correction for overall noise figure is made. **Editor.**

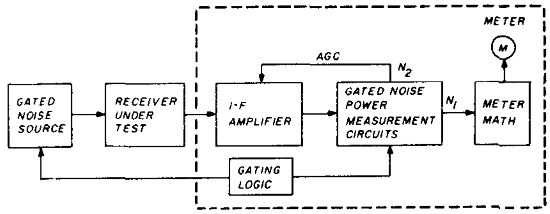


fig. 1. Simplified block diagram of the HP-340B Automatic Noise Figure Meter (copyright Hewlett-Packard Company, 1959. Used by permission). Commercial instruments such as this use the y-factor method of noise-figure measurement, where the y-factor is the ratio of N_2/N_1 (see eq. 10).

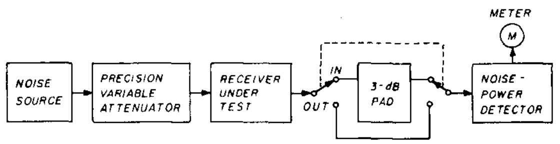


fig. 2. Block diagram of the twice-power method of manual noise-figure measurement. This method forces $N_2 = 2N_1$ by adjustment of the source excess noise ratio (ENR).

tector in the circuit has no effect on the accuracy of the measurement.

an automated twice-power measurement system

Commercial automatic-noise-figure (ANF) meters use the y-factor method because they're designed to work with a combination of sources, argon discharge, semiconductor or temperature-limited diode, necessary to cover a wide range of frequencies up to the GHz region. Frequencies of major interest to Amateurs are in the 10-600 MHz region. This allows the use of the temperature-limited diode as the only necessary source and greatly simplifies the measurement problem.

With reference to the block diagram of **fig. 3** and the timing chart in **fig. 4**, note that the noise source is gated both in the filament circuit and in the plate circuit. The plate gate turns the source *on* or *off*, and the filament gate adjusts the ENR by pulse-width modulating the filament current. Because of its long thermal time constant, the filament is not used for noise source on-off gating.

The receiver under test can be any combination of amplifiers and/or mixers with an output frequency of 28 MHz and an output impedance of 50 ohms.

The ANF-meter input is through an electronically switched 50-ohm 3 dB attenuator. This pad is switched into the circuit when the source is on, and out of the circuit when it is off. A train of noise pulses from the attenuator is fed to the i-f amplifier, which amplifies the noise pulses to a usable level. The input level of this meter should be 7 μV (-90 dBm) or greater for proper operation.

The i-f strip output is detected with a half-wave hot carrier diode detector and fed through an amplifier to the two sample-and-hold (S&H) amplifiers. The S&H amplifiers are circuits that will follow the input voltage when in sample mode and hold the final voltage of the sample period in hold mode until updated with a new sample pulse. S&H 1 samples the last half of the *attenuator out-source off* period. Sampling is done in the last half of the period to allow attenuator and source switching transients to die out. S&H 2 samples the last half of the *source on-attenuator in* period. The outputs of these two S&H amplifiers are presented to the comparator. If the comparator decides S&H 1 output is higher than S&H 2, it turns on the source filament. If S&H 2 output is higher than S&H 1, it turns off the filament. This sample-and-compare cycle is repeated at a 170-Hz rate.

At this update rate, the filament thermal inertia will average the on and off pulses to give a filament temperature corresponding to exactly 3 dB ENR. The S&H 1 output is fed back to the agc input of the i-f strip to reduce the i-f gain for converters or amplifiers with very high gain or poor noise figure. This is a rather loose agc loop but, for this applicaton, where the sample-and-hold amplifiers are compared only to see which one is higher, it's only necessary to keep the i-f output within the common-mode range of the comparator and the linear range of the sample-and-hold amplifiers. The agc loop time constants and loop gain are somewhat critical because this is a sampled-data servo loop. Such loops, unless properly compensated, will tend to be unstable. Linearity of the system is unimportant, as the i-f amplifier, the detector and the S&H preamplifier are all working at the same level of noise when the system is in balance.

The source plate current is detected as a voltage across the 100-ohm resistor in the source plate supply return. As this current is being pulsed by the gating circuitry, S&H 3 is needed to sample during the *source on* period and hold during the *source off* period. S&H 3 output, which is proportional to the source plate current, is supplied to the log amplifier, which converts noise factor to noise figure and allows the meter to have a linear scale. The log amplifier output feeds a "perfect" rectifier, which ensures that only positive voltage reaches the meter. The summing junction of this op amp makes a handy place to supply a calibrated offset voltage to add or subtract a correction factor for the source ENR *vs.* frequency error and for loss pads.

manual noise figure measurement circuit

This meter can also be used for manual twice-power-method noise figure measurements when it is difficult to get a 28-MHz output from the receiver under test. In this mode the source is turned on all the time, and a comparator monitors a voltage corresponding to desired plate current (set by a front-panel pot) and the voltage across the meter. When the meter voltage is higher than the set-point voltage, the filament is turned off and *vice-versa*. When the source is adjusted to give exactly 3-dB ENR, the

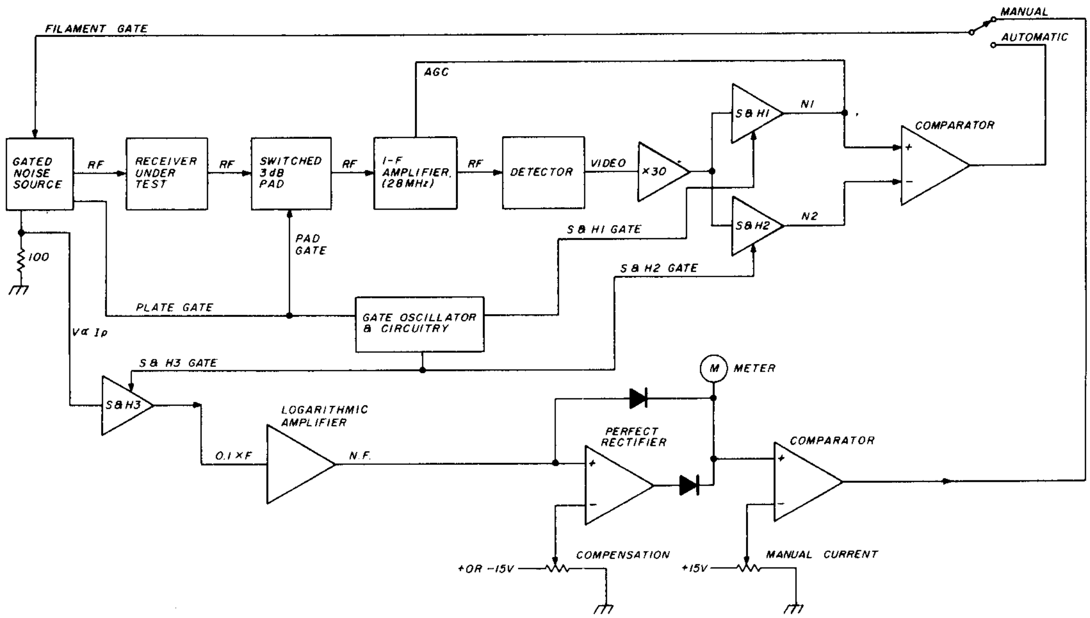


fig. 3. Automatic noise-figure measuring system using the twice-power method. System is designed for use in the 10-600 MHz region, which allows the use of a temperature-limited diode as the only necessary noise source, thus greatly simplifying the measurement problem.

noise figure can be read directly from the meter. The i-f loop is disabled during operation in manual mode.

circuit details

Fig. 5 shows the noise-figure meter schematic. The noise source can be either an HP-343A or the homebrew 5722 source described by W6NBI[1]. I prefer the 5722 source because the price for the tube is $9.00 compared to the $90.00 that HP charges for their noise diode. The homebrew source, if carefully constructed, should be as good as the HP source. The meter power supply is a fairly conventional ± 15V regulated supply. The 10-μF bypass capacitors for U13 and U14 should be mounted immediately at the device terminals or oscillation is a likely result.

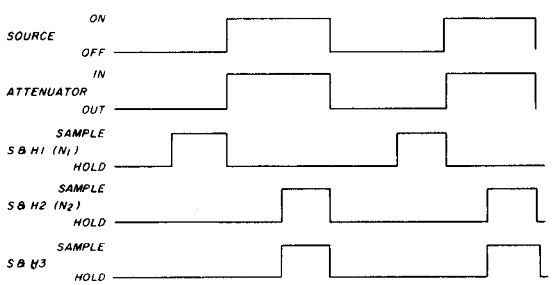
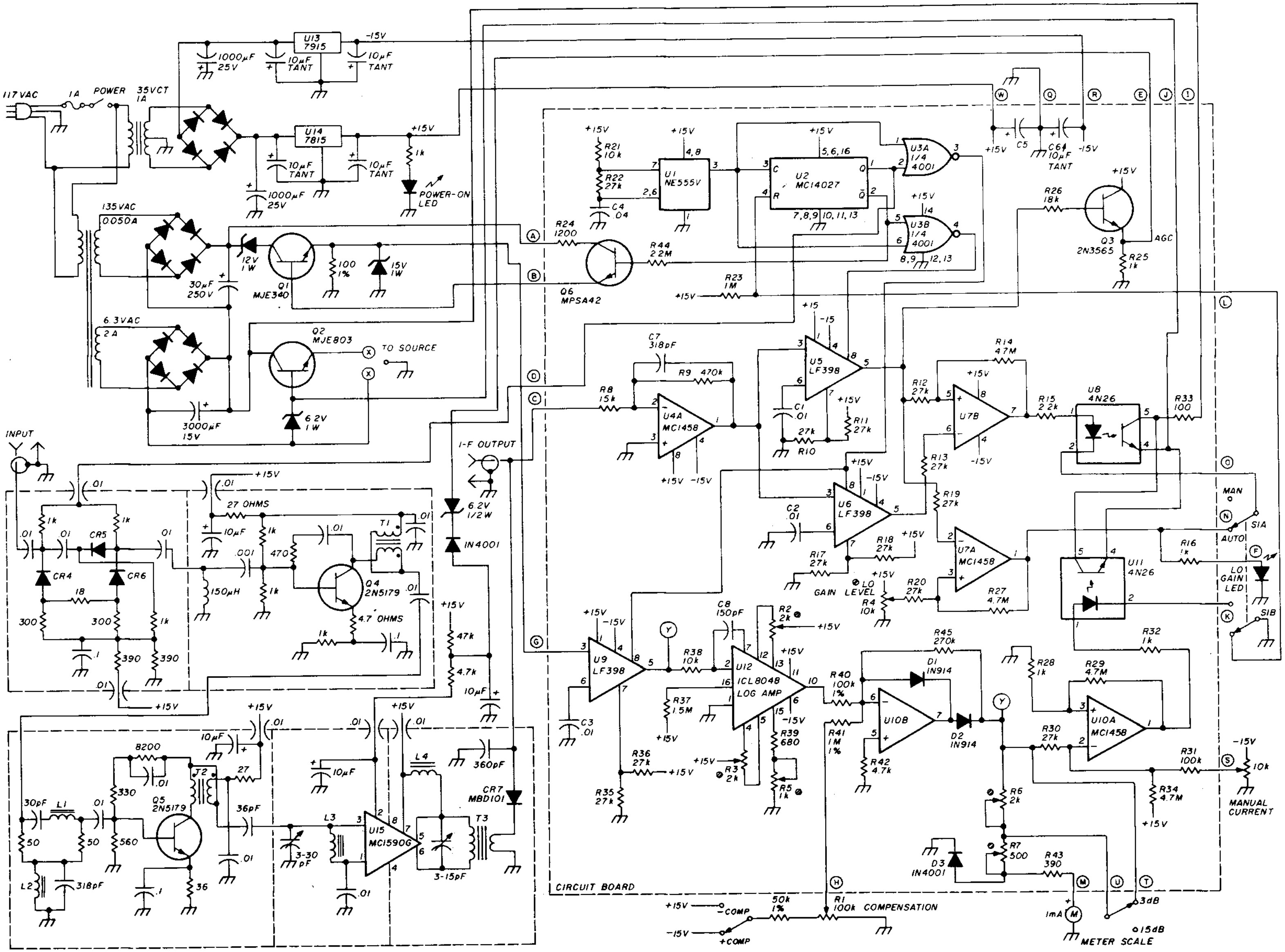


fig. 4. Timing chart for the twice-power noise-figure meter.

The source has a grounded plate, so the 200-volt plate supply has its positive end grounded through the gate transistor, Q1, and the current measuring 100-ohm resistor. Q1 and Q4 form a quasi-Darlington switch designed to minimize base current injection into the current-measuring circuit. The negative end of the plate supply is fed up the filament line to the source. Q2 gates the filament current, and the 6.2-volt zener diode ensures that the maximum filament voltage is 4.9 volts. Q2 is driven by opto-isolators, as it is floating 200 volts below ground.

The receiver noise output is applied to the electronically switched 3-dB pad at the input to the i-f strip. When the voltage at point D is + 15 volts, CR5 is turned on and CR4 and CR6 are back biased. This removes the pad from the circuit. When point D is grounded, the pad is switched into the circuit by CR4 and CR6, while CR5 is back biased. This pad should be symmetrically constructed using minimum lead lengths. The ultimate accuracy provided by the ANF meter depends on the accuracy of this pad. One-quarter-watt carbon film resistors and small disc ceramic capacitors should be used for best results. The output of the pad is connected to the next stage through a highpass filter to reduce the effect of switching transients on the next stage. Q4 is a broadband amplifier designed for 50-ohm input and output



C1,2,3	.01 µF polystyrene or polypropylene.
D4,5,6	Motorola MPN3402 PIN diode.
D7	Motorola MBD101 hot-carrier diode.
S1	Double pole double throw, center off, toggle switch.
U5,6,9	National LF 398 sample and hold amplifiers.
U12	Intersil ICL8048CCPE log amplifier.

L1	14t No. 22 T37-2 core
L2	3t No. 22 T37-2 core
L3	11t No. 22 T50-6 core
L4	12t No. 22 T37-61 core
T1,T2	6t No. 22 bifilar wound T37-61 core

fig. 5. Schematic of the automatic twice-power NF meter.

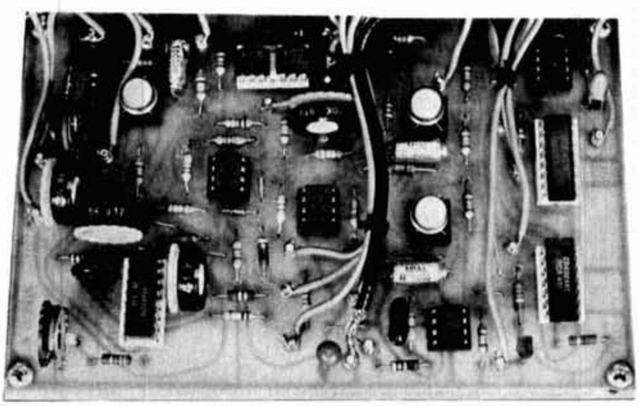
impedance and has 10 dB gain. A diplexer interstage network tuned to 28 MHz feeds another 11-dB broadband amplifier. The broadband negative feedback amplifiers are used to ensure unconditional stability and proper wideband termination for the pad.

U15, an MC 1590G, provides the bulk of the gain (50 dB) as well as a 60-dB agc range. Both input and output of this amplifier are tuned to 28 MHz; since it packs a lot of gain into a small package, shielding is necessary for stability. A hot-carrier diode coupled to the MC 1590 tank is the detector, and the 360-pF capacitor provides light filtering of the detected noise level. Averaging of the noise level is done by the memory capacitors in the sample-and-hold amplifiers. The detector output is coupled to U4A, which has a dc gain of 31 and provides some additional filtering.

N_1 and N_2 pulses are separated by the sample-and-hold amplifiers U5 and U6. Q3 buffers the output of U5. The 6.2 V zener diode and forward-biased silicon diode level-shift the agc for the MC 1590. The agc level starts at 6.8 volts and goes up with increasing signal level.

U7B compares N_1 and N_2 levels and decides whether or not to turn on the filament in the source. U7A is another comparator that decides if there is sufficient noise signal for an accurate measurement. If the N_1 pulse level is too low, the front panel "low gain" LED is lit, and the cathode of the LED in U8 is elevated to +15 volts. This action disables the filament circuit, keeping the terminating resistor cool and saving the noise tube.

U1, U2, and U3 compose the timing circuit for generating the gate signals. U1, an NE555V, functions as an astable multivibrator at 340 Hz. U2, a flip flop, divides U1's output by 2 and by combining U2 and U1 outputs in U3, the nonoverlapping quadrature sampling pulses are synthesized. In the manual



PC board.

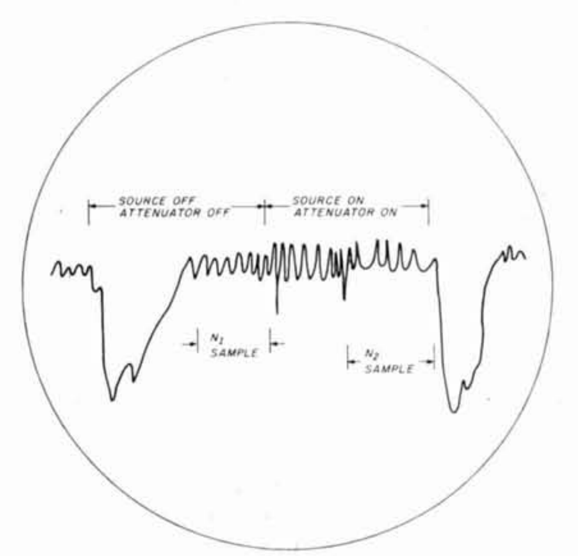


fig. 6. Sketch of oscilloscope pattern taken at the i-f output jack.

mode, U2's reset input is held high. This action enables the source continuously and U9, the meter circuit sample-and-hold amplifier, operates at a 340 Hz rate. In the automatic mode U9 samples during the last half of the source-on period and supplies a smoothed voltage proportional to plate current to the log amplifier, U12. The signal at the output of U12 corresponds to noise factor × 0.1, which the log amplifier converts to noise figure. This chip is quite expensive (about $35) and can be eliminated if the meter is recalibrated. The meter will now read noise factor and should have 3 mA and 30 mA full-scale positions. If you decide not to use the log amplifier, pull U12 and disable U10B by removing its external components. Connect points Y together and do not install the compensation pot, R1, or its switch.

U10B, a perfect rectifier, blocks negative voltages from pinning the meter down-scale. U12 output is negative for source plate currents of less than 1 mA and −15 V for zero plate current. U10A is the comparator for the manual mode. It decides if the source current, as represented by a voltage across the meter circuit, is higher or lower than the desired current, set by the front panel manual-current control. The 4.7 megohm resistor between the output of U10A and its noninverting input supplies a minor amount of hysteresis, and the 4.7 megohm resistor between the inverting input of U10A and +15 V swamps out any offset in U10A and ensures that the plate current of the source can be zeroed with the manual current control.

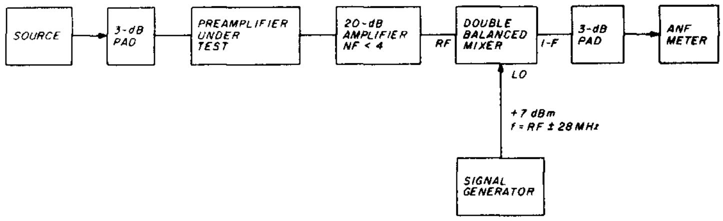


fig. 7. Block diagram showing technique for measuring noise figure of preamplifiers or converters.

construction

The i-f strip is built into an enclosure made of double-sided PC board. All voltages are brought into the strip through feedthrough capacitors. Button ceramic capacitors are used for bypass purposes, and ceramic standoffs are used for unbypassed tie points. Even though the i-f strip operates at 28 MHz, the F_T of Q4 and Q5 is 1.4 GHz, and the MC 1590 operates up to 300 MHz; therefore vhf construction techniques must be used to ensure stability. A shield should be mounted across pins 8 and 4 on the MC 1590 to isolate the input and the output of this stage.

The logic and analog circuits are all on the PC board. Sockets should be used for the integrated circuits as some must be pulled during calibration. All resistors are 1/4 watt 10 per cent tolerance, and all capacitors are disc ceramics unless otherwise noted. The power supply and switching transistors are mounted onto the chassis. Q2 should be bolted to the chassis for a heat sink. This transistor and the rest of the filament power circuit should be treated with caution, as it is at −200 volts.

Don't use a 2-pin connector for the source and rely on the connector shell for a ground; otherwise you may get a 200-volt surprise when you plug in the source with the power on. Use a 3-pin plug with the ground routed through a pin. U13 and U14 should be heat-sinked, and attention should be paid to the wiring of these ICs. Even though they are complementary ICs, the pinouts are different. The meter has a 1-mA 1000-ohm movement and is calibrated for 0-3 and 0-15 dB noise figure by erasing the original scale with an electric eraser and recalibrating with transfer letters. The compensation pot is a 10-turn linear digital readout pot with 10 volts applied across it. Either +10 or −10 volts can be used to give −10 dB to +10 dB compensation. This adjustment is used to null out the excess ENR of the source with increasing frequency and to compensate for loss pads between the source and the receiver.

checkout and alignment

First unplug all the ICs and confirm that the power supply voltages are correct. If they are, plug in U1, U2 and U3 and see if you get quadrature pulses out of U3A and U3B. Connect a 28-MHz signal generator to the input jack and an oscilloscope to point C. Increase the signal generator output until you see a train of pulses at point C. Peak the U15 input and output circuits. Reduce the signal level to some convenient value and verify that the levels of the pulse train are 3 dB apart. This must be done with a calibrated attenuator of known accuracy. Connect an rf pad (10 dB for example) between the signal generator and the input of the ANF meter. This standardizes the signal-generator output impedance to 50 ohms. Unplug U2 and connect point D to +15 V. Connect a DC voltmeter (preferably a digital voltmeter) to the i-f output jack and note the reading on the voltmeter. Then connect point D to ground and plug a 3-dB pad between the 10-dB pad and the ANF meter. The volt-

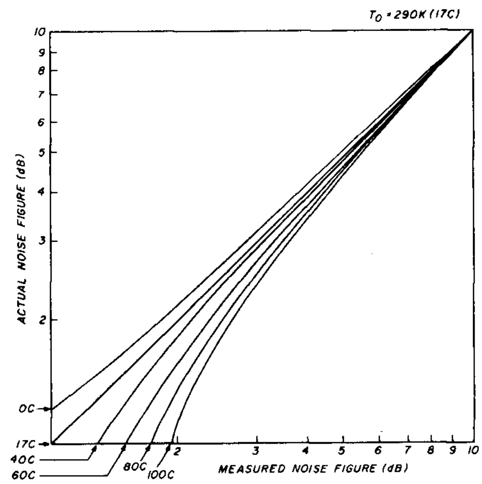


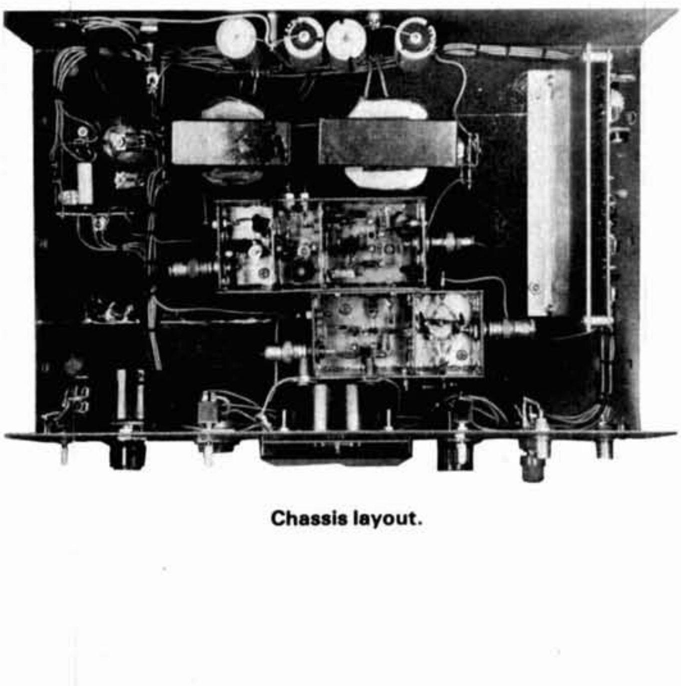
fig. 8. Temperature correction for a temperature-limited diode noise source (courtesy Hewlett-Packard Company).

meter level should remain the same. Disconnect the ground on point D and plug in U2. Plug in U4 and U5, and check to see if the pulse train is amplified by a factor of about 30 in U4A. Increase the signal level and see if the pulse train level at the i-f output jack stabilizes. This verifies that the agc loop is operational. Plug in U6 and U7. With the signal generator set for 7 μV and the pads removed, adjust R4 until the low gain LED just goes out. Turn the signal down and the LED should come on. Plug in U12 and temporarily solder a 10 k resistor between pins 2 and 7 of this IC. Ground pin 5 of U9 (make sure U9 is out of the socket), and adjust R3 for zero volts at U12 pin 7. Unsolder the 10 k resistor and unground U9 pin 5.

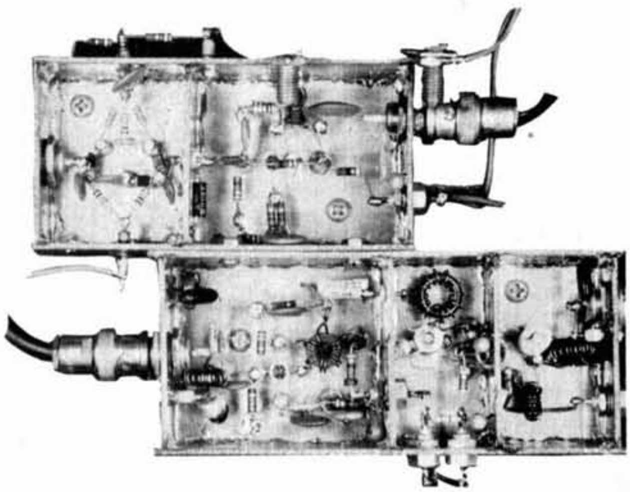
Connect an accurate source of 0.1 volt to U9 pin 5 (make sure U9 is still out of its socket) and adjust R2 for zero volts at pin 10 U12. Now turn up the source to +10 volts and adjust R5 for exactly −2.0 volts at pin 10 U12. Plug in U10, turn the compensation pot to zero, and adjust the voltage source to +0.2 volts at U9 pin 5. With the meter switch in the 3-dB position, adjust R7 for full scale. Change the meter switch to 15 dB, increase the source voltage to +3.16 volts, and adjust R6 for full scale.

Disconnect the voltage source and plug in the remainder of the ICs. Turn the AUTO STBY MAN switch to manual and plug in the noise source. The manual current control should vary the plate current (as expressed in NF) from zero up to 12 or 14 dB before saturating. Flip the switch to automatic and verify that the filament remains dark.

Connect the output of an amplifier to the input jack and apply power to the amplifier. The filament should light and the low-gain LED should extinguish. The meter should swing to saturation (12-14 dB). Now connect the source to the amplifier. Assuming the noise figure of the amplifier is in the range of the instrument, the meter should now indicate its noise figure.



Chassis layout.



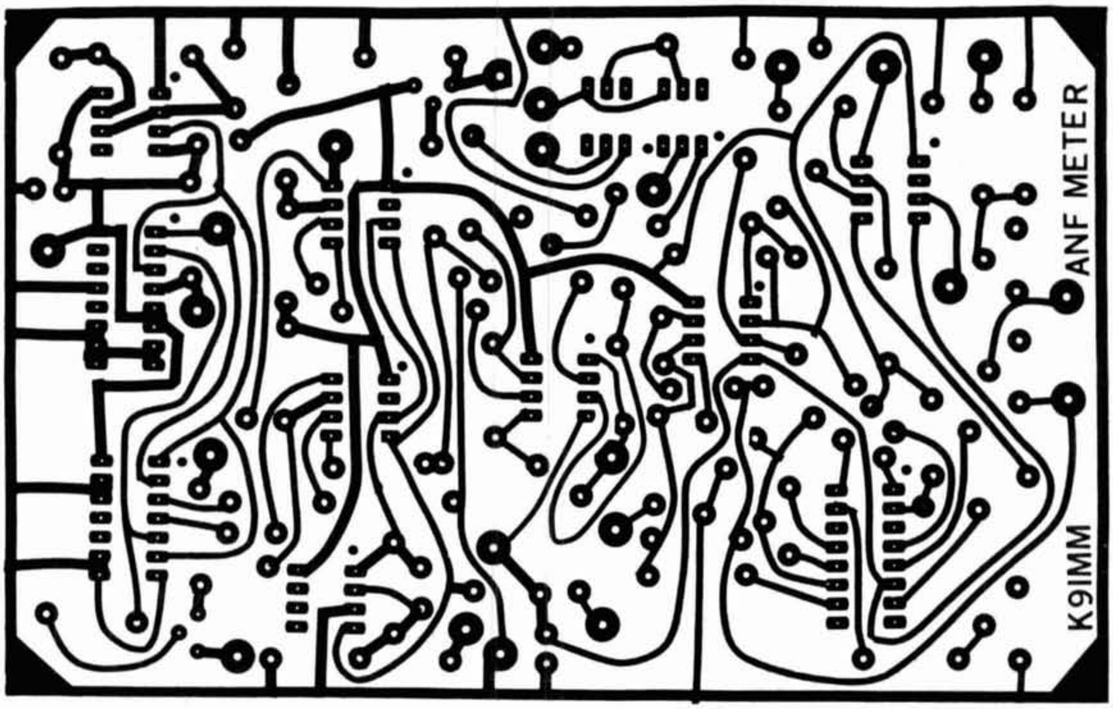
I-f strip.

operation

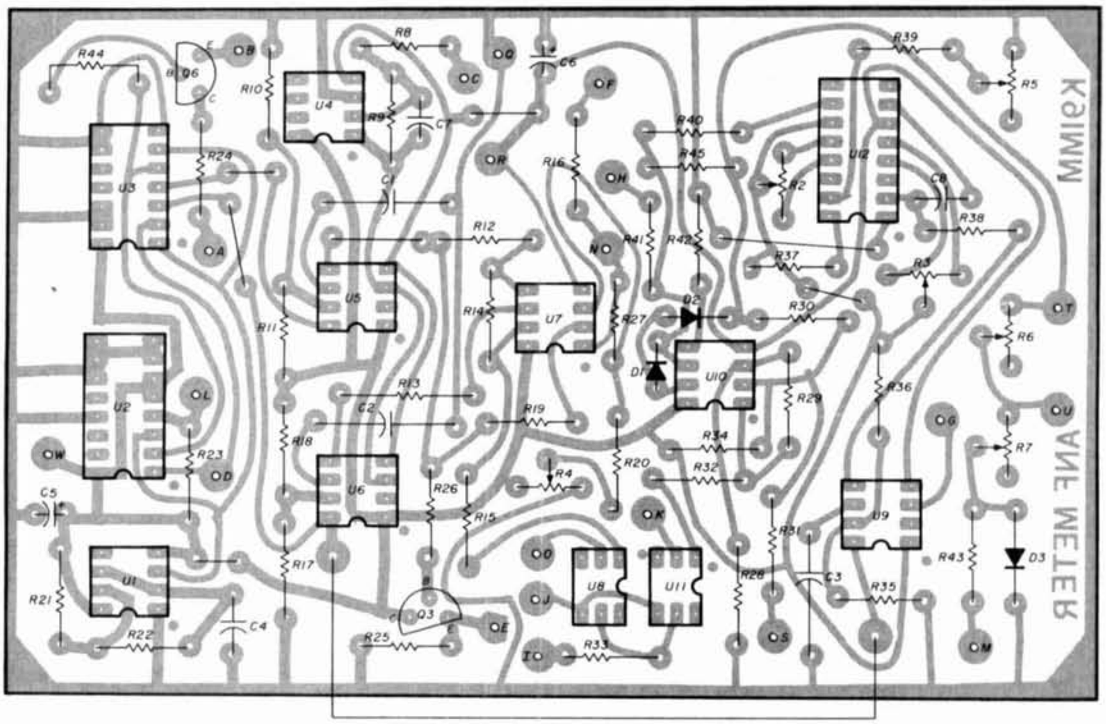
The reader is referred to the article by W6NBI on automatic noise-figure measurements for background on proper use of ANF meters.[3] A few comments pertaining to this particular meter are in order. Fifty-ohm output impedance of the receiver under test is required for accurate operation of the input pad in the meter. Amplifiers and converters seldom have 50-ohm output impedance, so a 3-dB or greater pad should be inserted between the receiver under test and the meter input. This pad won't cure the mismatch, but it will swamp out its effects on the measurement.

The receiver under test must be stable, since the meter is unable to tell the difference between noise and oscillation. In the same vein, if you are in a high rf-level area, the receiver should be well shielded. I live within three miles of four TV stations, four commercial fm stations, and a multitude of commercial and Amateur repeaters and paging services; consequently I find it difficult to get proper measurements on unshielded converters or preamps.

An i-f output jack is mounted on the front panel for connection to a scope to verify proper operation of the comparison circuitry. **Fig. 6** shows what the scope picture should look like. The important thing is that the N_1 and N_2 sample periods should be at the same level. The large negative pulse at the beginning of the source *off* period is caused by the source not turning off immediately. This is because of the finite discharge time of the bypass capacitors in the filament circuit of the source. A 5-way binding post mounted on the front panel is used as a handy source



PC-board layout, foil side.



PC-board layout, component side.

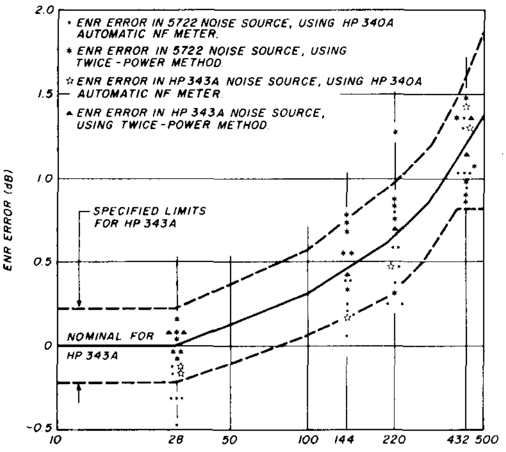


fig. 9. ENR error as a function of frequency for the HP-343A or 5722 source (from reference 1).

of +15 volts to supply converters and amplifiers. When working with preamps with a sub-3-dB noise-figure, be aware that the input impedance for best noise figure is seldom 50 ohms. This presents a mismatch to the source and indicates an erroneous noise figure, which is usually too high. A 3-dB pad between the source and the receiver input will swamp out most of the error. The 3-dB-pad effect can be nulled out by the compensation pot, so the meter will read noise figure directly. I use the technique in **fig. 7** to measure converters or preamps. The first 3-dB pad terminates the source, and the 20-dB-gain amplifier after the preamp under test reduces the effect of mixer loss on noise figure to a negligible level. The mixer is a broadband, double-balanced mixer (ANZAC MD-108) packaged in a small box with BNC connectors for the ports. The 3-dB pad after the mixer terminates the switched pad in the meter. The signal generator supplies local-oscillator power at $f \pm 28$ MHz. If the amplifier has a reasonable noise figure (≤ 4 dB), and the preamp has a reasonable gain (≤ 15 dB), then the compensation for a 2-meter preamplifier would be figured as follows: At 144 MHz, ENR error is +0.45 dB. A 3 dB pad is used so that compensation is $0.45\,dB - 3\,dB = -2.55\,dB$.

Set the compensation pot to −255 and read the system noise figure directly from the meter.

The method for determining preamplifier noise figure from system noise is:

$$NF_1 = 10 \log \left[F_S - \left(\frac{F_2 - 1}{G_1}\right) - \left(\frac{F_3 - 1}{G_1 G_2}\right) - \ldots \right]$$

NF_1 = *preamp noise figure (dB)*

F_S = *system noise factor*

G_1 = *preamp numerical gain*

G_2 = *second stage numerical gain*

F_2 = *second stage noise factor*

F_3 = *third stage noise factor*

Note that this formula uses noise factors and numerical gains, *not* noise figures and decibel gains.

If the source terminating resistor temperature is other than 290 K (62 F), a correction must be made for this error. 290 K (62 F) is below room temperature, and the resistor is very close to a hot tube, so this correction is usually necessary. Unfortunately this is a non-constant error, so the compensation pot can't be used. The best technique is to factor the other errors into the compensation pot correction and then use that figure to work up the temperature correction from **fig. 8**. The ENR correction for frequency for the HP 343A source and the W6NBI 5722-type source is given in **fig. 9**. This error should be nulled out as a + compensation with the compensation pot. Be aware that the source accuracy is the limiting factor for the measurement accuracy. I suggested that the hot-cold resistor noise figure measurement system be used to verify the excess-noise ratio output of the source at various frequencies.[4] If this is done, this meter should be as accurate as the commercial models.

Noise figures, like antenna gains, are numbers that are often claimed but seldom measured. My involvement in this project has been well worth the effort, especially when I see the long faces (mine included) on the owners of "hot" preamps when the meter reads otherwise.

I'd like to acknowledge the assistance of W9XM and WA9ACI in reviewing this paper and for many spirited discussions on the philosophy of noise-figure measurement.

references

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2. "Noise Figure Primer," Hewlett Packard Application Note 57, January, 1965.
3. Robert S. Stein, W6NBI, "Automatic Noise Figure Measurements — Fact and Fancy," *ham radio*, August, 1978, page 40.
4. Benjamin L. Lowe, K4VOW/WA5UUM, "Hot and Cold Resistors as UHF Noise Sources," *QST*, September, 1976, page 32.

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1. Fisk, James R., WIDTY, "Receiver Noise Figure, Sensitivity and Dynamic Range — What the numbers mean," *ham radio*, October, 1975, page 8.
2. Krauss, Geoffrey H., WA2GEP, "VHF Preamplifiers," *ham radio*, December 1979, pg. 50.
3. Intersil 8048/8049 Application Note, Intersil, Inc., Cupertino, California.
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ham radio

Amateur Radio equipment survey

Detailed reports from owners of several models of popular Amateur Radio gear were featured in various issues of *Ham Radio Horizons*. These reports were received with much favor by readers who were contemplating the purchase of equipment, either new or used.

This month we continue our owners' equipment survey by selecting for review three radios that have enjoyed much popularity in the Amateur community. The radios have been in use long enough so that a fairly broad sample of opinion and experience can now be collected. They are the Icom 701, Drake TR-7, and Kenwood TS-520 series — all high-frequency transceivers.

The items in the owners' report form have been chosen to extract the most information of use to the prospective buyer in making his choice. The results of the survey will show what owners really think about their equipment — what was best liked, what was disliked, what types of problems were encountered and how they were resolved, and in general what owners felt about performance, maintainability, and reliability.

Reading the results of the survey will surely help you decide how to spend your money for that new rig. You can profit from the experience of Amateurs who have learned by doing — putting the equipment to work under actual operating conditions. Such information can be much more meaningful than a laboratory report made from tests under controlled conditions.

The report sheets can be even more useful if comments are added, in addition to answers being checked off where called for. Feel free to let us know your opinions. The more information we can gather, the better we'll be able to serve prospective buyers.

Next month we'll be publishing the first part of our two-part rundown on Collins gear, in which we'll present the data we've collected from users' comments on the KWM-2 and KWM-2A transceivers. In the April issue, we'll appraise the 32S-series transmitters and the 75S-series receivers. As in the past, the readers of *ham radio* have made some perceptive and revealing comments on the equipment they own and use. If Collins equipment interests you, be on the lookout for these two articles.

Owner's Report on Amateur Radio Equipment

(Please report only from your own experience. Type or print clearly.)

1. Make and Model (please circle the exact unit you are reporting on).

ICOM 701 | Drake TR7 | Kenwood 520
520 S
520 SE

2. What year did you buy it? ________ New? ________ Used? ________
3. Where did you buy it? Dealer________ Mail Order________ Individual________ Flea Market________ 800 Number________ Other________
4. Would you buy from the same source again?________
5. Amount of use: Daily________ Often________ Occasional________ Seldom________
6. Is this your primary________ or backup________rig?
7. What modes have you used? CW______ SSB______ RTTY______ SSTV______ AM______ Other______
8. What is the rig's best feature?____
9. Worst feature? ____
10. Have you had any problems?________ Explain ____
11. Have you had the rig serviced?________ Where? Manufacturer________ Dealer________ Other________
12. Was the service satisfactory? Yes________ No________
13. What accessories have you purchased for this rig? ____
14. Have you been able to obtain all the accessories and parts you need? ________________
15. Have you been satisfied with these accessories? Yes________ No________
16. If not, why? ____

17. Additional features you would like to see built into a rig of this type ______________________

____

____

____

18. Give the equipment a score from 1 to 10 (with 1 being poorest, 4 to 6 average, and 10 perfect).

Ease of operation ______________

Reliability ______________

Durability
(in continuous use) ______________

Instruction Book ______________

Factory/Dealer Service ______________

Quality of Workmanship ______________

Performance ______________

Maintenance ______________

Parts Availability ______________

Accessories
(ease of connection) ______________

Price ______________

Flexibility ______________

19. How long have you been licensed? ________ Your Age ________ License Class ______________

Principal activities: Contest ________ DX ________ Rag Chewing ________

Traffic Handling ________ Experimenter ________

20. What antenna do you use most? Beam ________ Wire ________ Vertical ________ Other ________

21. What rig would you like to see reported on in the future? ______________________

22. Would you buy this same rig again? ______________________

23. (Optional: fill in the following only if you wish.)

Submitted by: Name ______________________ Call ________

Address ______________________

City ______________ State ______________ Zip ________

____

(Signature)

(Your signature authorizes *ham radio* to quote portions of your comments in our report.) May we use your name and/or call?

Yes ________ No ________

Note: If you own more than one of the rigs indicated, please use a separate form for a report on each rig.

Completed survey forms must be returned no later than March 31, 1981, to be included in our report.

Mail To: Ham Radio Owner's Report No. 4, Greenville, NH 03048

80-meter receiver
for the experimenter

Basic building blocks
for those
who like to build
their own gear

This simple receiver covers the 80-meter band. Other bands can be received by adding converters. This article has been submitted with the idea of giving other experimenters some ideas about receiver construction. A simple low-frequency i-f filter is also included. It provides a bandwidth of about 1200 Hz — great for CW if the band isn't too crowded. The nice thing about the filter is that it's inexpensive: two crystals at $2.00 each plus a few other parts (also inexpensive).

description

My receiver was built on a 5×9 inch (12.7×23 cm) chassis with a panel measuring 9×5 inches (23×12.7 cm). I found room for a power supply and small speaker. The dial was Japanese. With a switch on the back of the chassis I can change to battery operation instead of the ac supply, thus making a nice portable receiver that will fit into a travel bag.

antenna input circuit

No rf stage was used in this receiver design because I anticipated that converters would be used for higher-frequency bands. An rf stage isn't needed on 80 meters — this saves some space. The antenna is fed directly into an attenuator, which I found necessary to prevent receiver overload. The attenuator was made from a dual 10k-section pot (see **fig. 1**).

The antenna coils are double tuned with separate 100-pF capacitors for simplicity rather than a split stator capacitor. This eliminates the tracking problem and hard-to-find parts.

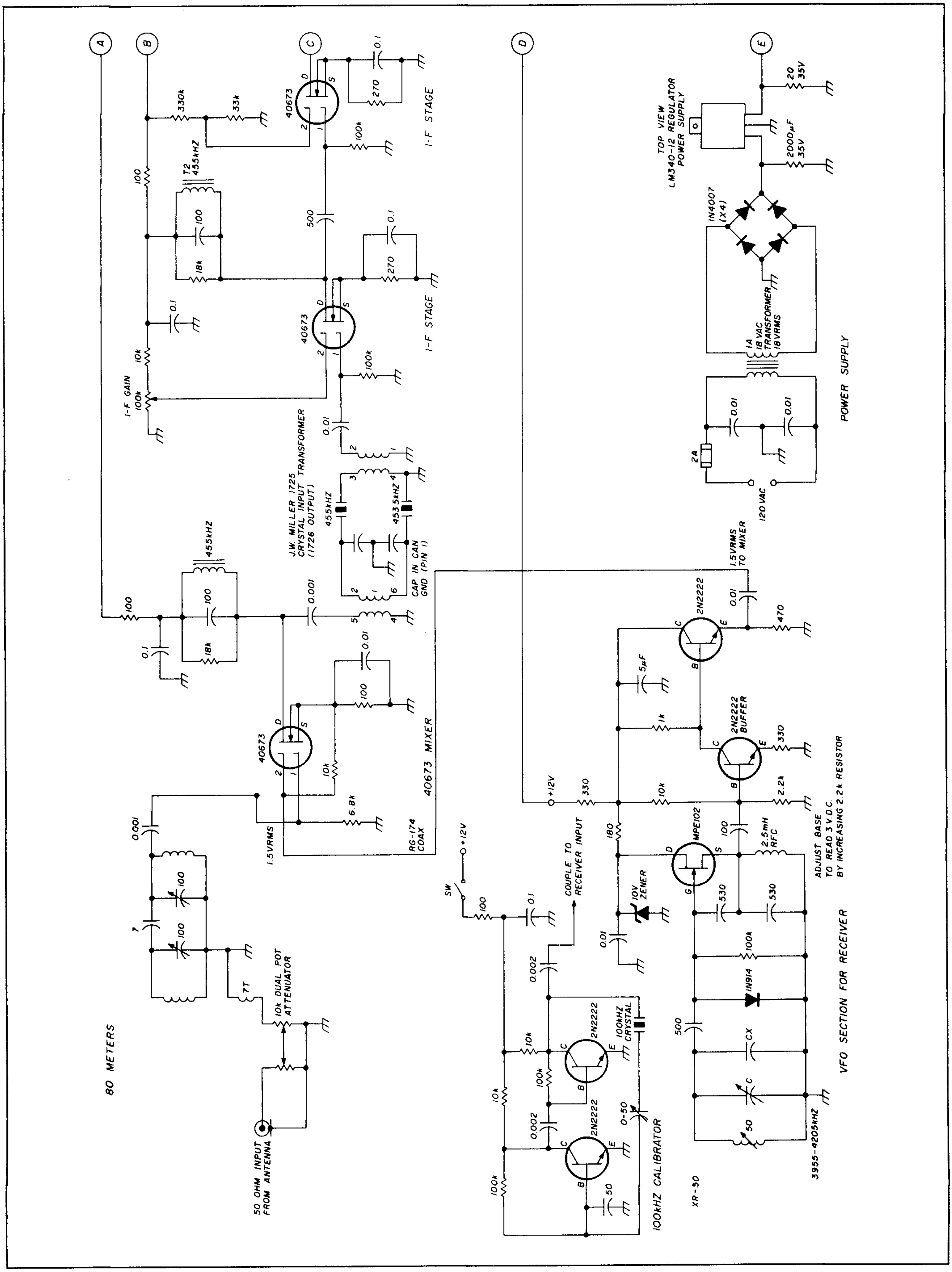
The antenna coils are wound on red Amidon cores. I used T-80-2, 1 inch (25.4 mm) in diameter, because they were available. The 68-2 will also be correct — anything to resonate at 80 meters. It takes about forty-five turns of no. 26 (0.51-mm) wire using a 100-pF variable for tuning. I used a 50-pF and padded it for resonance at mid-capacitor position. Resonance can be checked using a grid-dip oscillator with a loop around its coil and one around the toroid.

mixer

A 40673 dual-gate MOSFET was used for the mixer. The resistor in gate 1 isn't critical and can be anything from 6.8k to 100k; however, the lower values are better, as overloading is possible with higher values.

I lucked out for a coil in the drain circuit: a small, potted, ceramic toroid about 1 inch (25.4 mm) square. I found it in a bag of coils sold by Radio Shack. It's not listed in their catalog, so it is worth looking for in the store. Otherwise, a J.W. Miller variable slug coil 4515, a 350 μH to 475 μH inductance

By Ed Marriner, W6XM, 528 Colima Street, La Jolla, California 92037



80 METERS
50 OHM INPUT FROM ANTENNA
10k DUAL POT ATTENUATOR
7T
40673 MIXER
1.5VRMS
RG-174 COAX
I-F STAGE
I-F STAGE
I-F GAIN
T2 455kHZ
J.W. MILLER 1725 CRYSTAL INPUT TRANSFORMER (1726 OUTPUT)
CAP IN CAN GND (PIN 1)
455kHZ
453.5kHZ
40673
1.5VRMS TO MIXER
2N2222
2N2222 BUFFER
MPE102
10V ZENER
2.5mH RFC
ADJUST BASE TO READ 3 V.D.C. BY INCREASING 2.2k RESISTOR
VFO SECTION FOR RECEIVER
3955-4205kHZ
XR-50
COUPLE TO RECEIVER INPUT
100kHZ CRYSTAL
100kHZ CALIBRATOR
TOP VIEW LM340-12 REGULATOR POWER SUPPLY
1N4007 (X4)
1A 18 VAC TRANSFORMER 18VRMS
120 VAC
2A
POWER SUPPLY

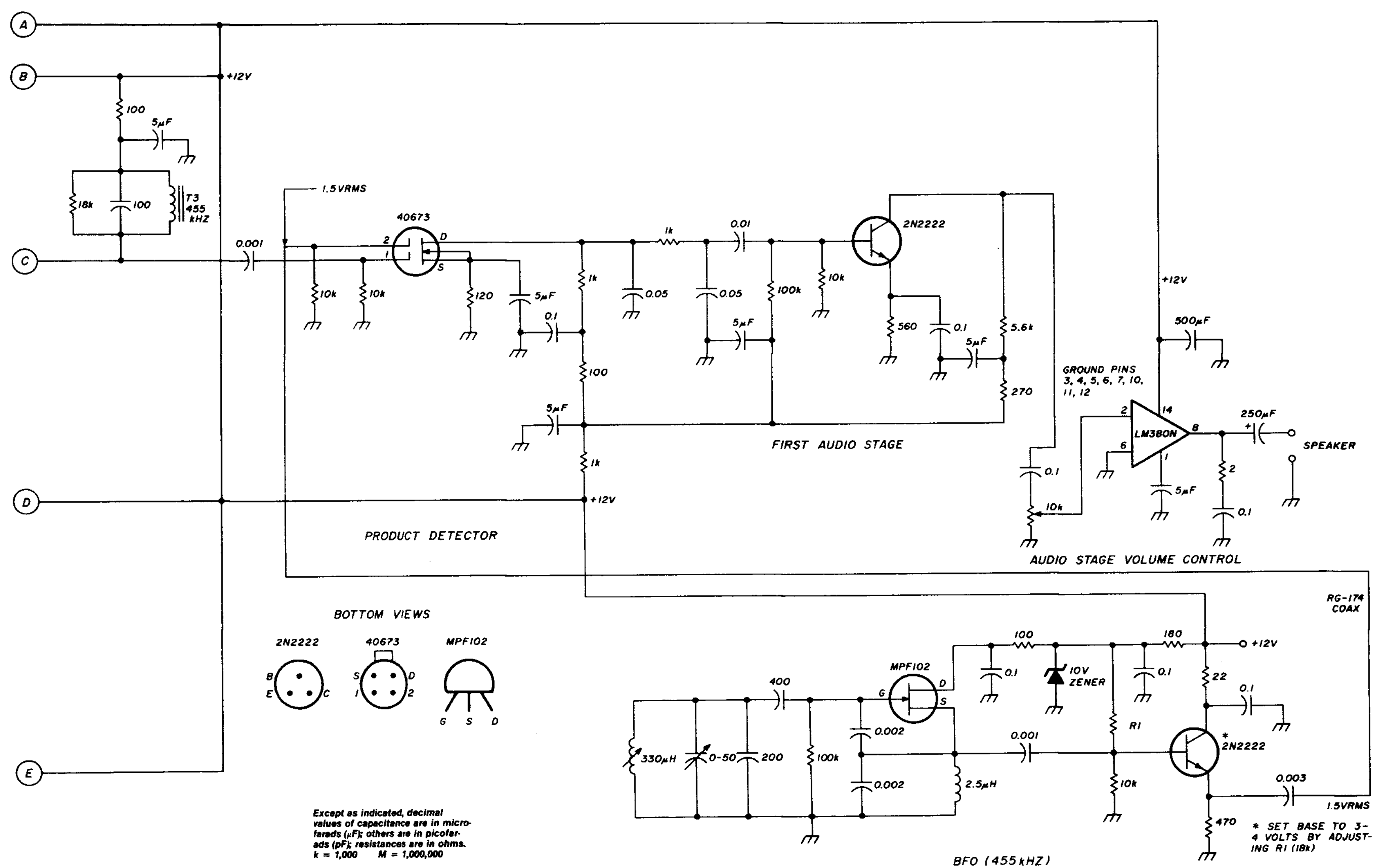


fig. 1. Schematic of the 80-meter receiver. Easy to build and easy to get working, it's a good project for those who like to experiment.

with a 220-pF capacitor should work. Another solution is to use half of an old 455 kHz i-f transformer.

crystal filter

The filter was made with two crystals, one of 455 kHz and one of 453.5 kHz. This is great for CW with a bandwidth of about 1200 Hz. However, a little wider spacing could be in order for SSB. The reason for these particular crystals is they are inexpensive: $2.00 each from the source indicated. The J.W. Miller Company offers an input and output transformer (1725 and 1726) to match crystals for the filter. Note that the capacitors in the output of the 1725 transformer are inside the can. You don't need to add them; just ground pin 1.

i-f stage

Two i-f stages were used, although it might be possible to get by with only one as the gain of this receiver is pretty high. I used two stages and reduced the gain with the i-f gain control.

product detector and audio

Rather than using a passive detector with diodes I used a 40673, which has some gain, to drive the first audio stage: a 2N2222. The product detector needs about 1.5 volts RMS injection voltage. I've had pretty good luck with this audio circuit, which delivers about 2 watts into a big speaker. Other lower-powered chips are available, but I used this circuit, which is in most of the commercial sets in use at the moment.

BFO

A crystal-controlled BFO could be used, but I've had difficulty making crystals oscillate at this frequency so I used the variable BFO. It's also very useful for zeroing in SSB signals and changing CW pitch. It's very stable, with no pulling.

VFO

The VFO tunes 3955-4205 kHz to cover the 80-meter band. It will take a little playing around to get it right. It's nice if you can borrow a counter or have a receiver that will cover this range. I used a National XR-50 5/8-inch (15.9 mm) diameter slug-tuned form for the coil. I had some silver wire that was nylon covered. It was no. 24 (0.511-mm) diameter. Using this wire, the VFO was very stable, but locating such wire is difficult. (This wire was found at a flea market; enameled wire is the next best choice.)

The variable capacitors available will determine your bandspread: capacitors of 0-50 or 0-100 pF will cover the range, or you can use a switch and two silver mica padding capacitors, which is what I did to cover the 80-meter band in two steps.

The capacitors I used from gate to ground and the coil coupling capacitor also have an effect on the tuning. The coupling capacitor has an effect on the oscillation and its value sometimes must be reduced. However, with the values shown, the VFO worked well and produced 1.5 volts rms — enough for mixer injection. While many circuits don't show buffers, I've found that signals don't pull the oscillator if a buffer is used. Thus, the set is more stable.

construction

Because of the difficulty in obtaining parts these days, it's impossible to specify an exact component (see table 1). I search the surplus ads, flea markets, and surplus stores. One of the best sources is the flea markets that radio clubs sponsor. There I've found all my parts.

This set was made from copper board and black drafting tape and dots, then etching. The board was mounted onto the chassis using spacers. I cut my panels with a hacksaw, with the aluminum held between two pieces of angle iron secured in a vice. After drilling, I dipped the panel in lye water. Or I let the panel set in *Lime Away*, a grocery store product, overnight.

Sometimes I spray the panel with black crackle paint and bake it in the oven at 175F for 15 minutes. The finished products look good. The boards can be dipped into a tinning solution to make them commercial-looking.

Holes for parts are drilled with a number 60 drill. There are a lot of little things you can do to make your homebrew projects look nice if you so desire. Many hams just want them to work, but I like to have them look nice, too.

table 1. Suggested sources for parts.

J.W. Miller Co. 19070 Reyes Avenue Compton, California 90221
Low Frequency Crystals ($2.00 each): John L. Winton 8062 San Mateo Cr. Buena Park, California 90621
Radio Shack Stores
Integrated Circuits Unlimited 7889 Clairemont Mesa Boulevard San Diego, California 92111
Semiconductor Supermart P. O. Box 3047 Scottsdale, Arizona 85257

ham radio

More Details? CHECK—OFF Page 98

solid-state power for 1296 MHz

Here's a Class C amplifier that delivers 2 watts of output power

In the past few years quite a few construction articles describing preamplifiers, converters, filters, and test equipment for the 1296-MHz band have been published. However, there has been a conspicuous lack of articles about transmitters for this band. This is a serious omission, because, for reasonable communications range, a transmitter power of at least a watt or two is needed — far more than the milliwatts produced by most converters.

background

Most of the 1296-MHz transmitters[1,2,3] described in the last decade have used surplus microwave tubes as the active components in the final stages. These planar triode tubes, commonly known as "lighthouse" tubes because of their distinctive shape, have been available on the surplus market for many years. And, although amplifiers built around this type of tube are capable of delivering tens or even hundreds of watts output, they suffer from at least two drawbacks that have limited their popularity.

First, the tubes are normally operated in a cavity or coaxial type of structure. Fabrication of these types of matching networks is not particularly difficult, but some machine shop work is generally required. Second, these tubes, usually of the type 2C39 family, are becoming somewhat scarce on the surplus market, as the original users of the tubes have largely switched to diode and transistor devices for medium-power amplification at the lower microwave frequencies.

At the present time, bipolar transistors are commercially available that are capable of at least 40 watts of continuous output at one GHz and 20 watts at 2 GHz, approximately the same power level to be expected from a single 2C39 tube.

Few Amateurs have built transistor power amplifiers for the 25-cm ham band, however. This is mainly because microwave transistors have a reputation for being expensive, tricky, and easily destroyed[4]. But with recent improvements in technology, rf transistors have become not only better but also more rugged, while the cost per unit has dropped because of the tremendous increase in the volume of production. As a result, transistor amplifiers capable of producing several watts output power are now easily within the technical and financial reach of Amateurs.

This article describes a two-stage amplifier chain that will deliver over 2 watts output. The total gain of the two units together is at least 30 dB, which means that a few milliwatts of drive power from a low-level converter or exciter will produce full output. The amplifiers are sufficiently broadband to cover the entire 1250-1300 MHz range, so that once tuned up they need not be retuned for operating frequency changes.

Construction of these amplifiers is straightforward and only hand-tool work is required. They are built using a mixture of microstrip line and lumped component techniques on copper-clad boards. Each stage is built as a module with 50-ohm input and output to simplify testing and to increase flexibility.

The first stage produces at least 100 milliwatts of linear output power. The gain at this power level is approximately 17 dB. The next stage provides a minimum of 2 watts output. It is operated Class C and has a gain of 13-14 dB. When the two are cascaded together as shown in **fig. 1**, a 2-milliwatt drive signal will produce over 2 watts output. Alternatively, the first stage may be used alone as lower-power transmitter to produce up to 150 milliwatts of saturated power.

By Jerry Hinshaw, N6JH, 4558 Margery Drive, Fremont, California 94538

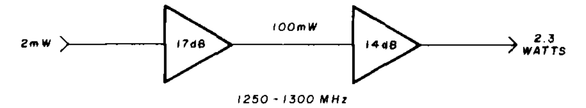


fig. 1. Block diagram of the microwave amplifier chain showing typical power levels. Total stage gain is at least 30 dB. Broadband design ensures coverage between 1250-1300 MHz, so that no retuning is needed for frequency changes.

100-mW stage

The first stage of the chain is a linear, Class A amplifier. As such, it's designed using the small-signal, grounded-emitter, s-parameter information provided by the manufacturer.

The transistor used in this stage is a Hewlett-Packard 2N6679, also known as the HXTR 2101. There are several reasons for selecting this transistor in preference over another, even though at approximately $22 it isn't the most inexpensive device available. This microwave transistor is designed to produce moderately high linear output power and is therefore fully characterized for this type of operation. Second, it has a higher matched gain than most less-expensive transistors. This means that one less gain stage is needed to amplify a signal up to the 100 mW level required to drive the output stage. This reduction in system complexity more than offsets the higher initial cost of this one stage.

The design of this amplifier follows the approach described in detail by Shuch.[5] Each input and output matching networks consists of a series quarter-wave transmission line transformer that matches the real part of the transistor input or output impedance. Shunt capacitance is used to compensate for the reactive portion of the output impedance. The input impedance of this transistor is by coincidence nearly a pure resistance at this frequency, so that satisfactory matching is achieved without a shunt capacitor at the transistor base. A schematic diagram of this amplifier is shown in **fig. 2**.

The design is accomplished using s-parameter data for 1275 MHz, data which were obtained by interpolating the 1.0- and 1.5-GHZ data provided by the manufacturer. Although such a design is, strictly speaking, valid for only this one frequency, in practice the s-parameters do not vary rapidly for small changes in frequency, so the usable range is a band several per cent wide centered at 1275 MHz.

The input impedance of this transistor at 1275 MHz is found to be approximately 5.3 ohms. As it is very nearly a pure resistance, it can be matched with only a quarter wavelength transformer whose characteristic impedance is 17 ohms. This 17-ohm transformer is realized on 1/32 inch (0.8 mm) glass-epoxy board by a microstrip line approximately 0.225 inch (5.7 mm) wide and 1.14 inches (29.0 mm) long.

At the transistor collector the impedance is approximately 50.6 ohms in series with 110.9 ohms of capacitive reactance. The real part of this impedance presents a good match to the 50-ohm load as it is, without further matching. The reactive part of the impedance is tuned with a capacitor located a quarter wavelength from the transistor. This capacitor must have a reactance of 22.5 ohms, corresponding to 5.7 picofarads at 1275 MHz. The quarter wavelength line itself is 0.056 inch (1.4 mm) wide and 1.15 inches (29.2 mm) long on this 1/32-inch (0.8-mm) printed circuit board material.

Dc power is provided by the constant-current bias network shown in the schematic. In an earlier version of this amplifier a simpler, passive bias scheme was used, but it proved to be a false economy because bias instabilities destroyed the HXTR 2101 transistor. The present circuit automatically compensates for the current gain variations from one transistor to another and for variations that result from temperature changes. Thus it provides for stable, Class A operation. Collector current is held at 25 milliamperes while the collector-to-emitter voltage (V_{ce}) of the rf transistor is stabilized at approximately 15 volts.

construction

The amplifier is built on a piece of double-clad Fiberglass epoxy PC board material 1/32 inch (0.8 mm) thick. One side of the board is left fully clad with copper to serve as a ground plane for the microstrip

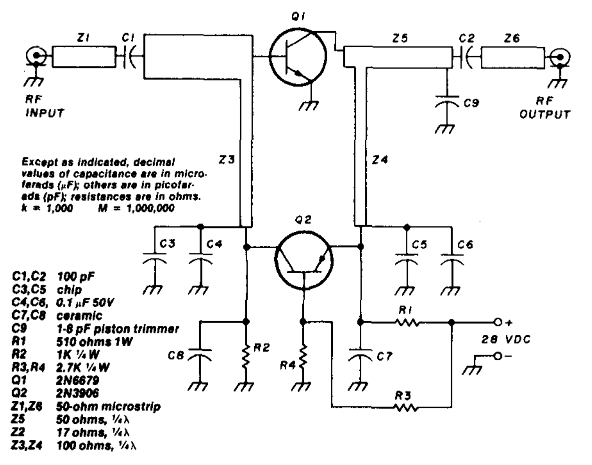


fig. 2. Schematic of the 100-mW driver. Stage operates as a Class A amplifier, producing linear output using the H-P 2N6679 device (also known as the HXTR-2101).

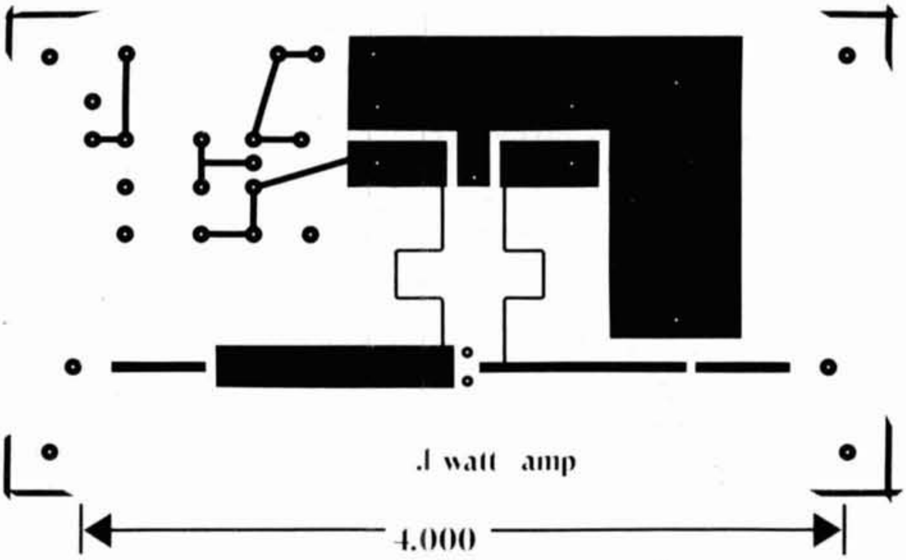


fig. 3. Full-size illustration of the PC board for the 100-mW driver stage.

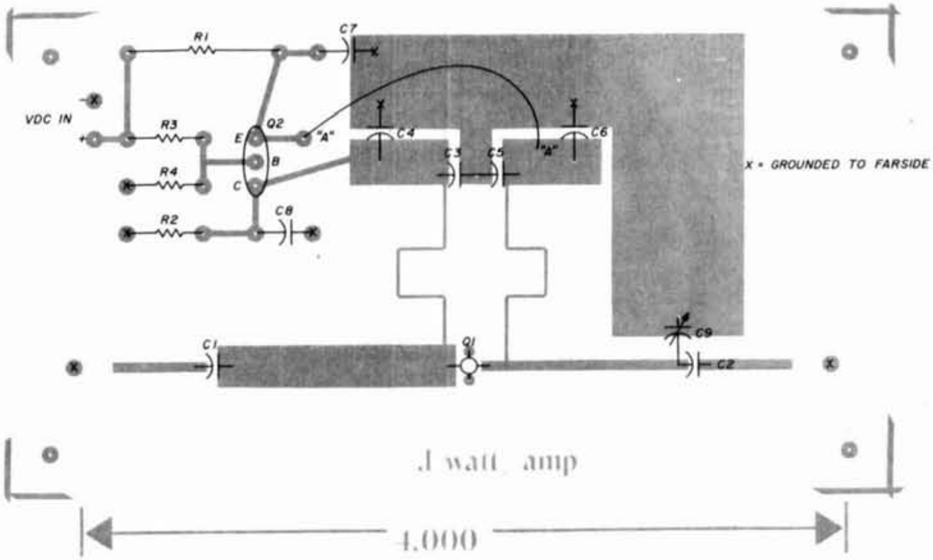


fig. 4. Component placement for the 100-mW stage.

lines. The other side is etched to form the matching and bias circuit mounting areas. A full-scale illustration of this side of the board is shown in **fig. 3**.

Feedthrough eyelets are mounted in the board at the positions marked on **fig. 4** with an **X**. These eyelets are used to help ensure that the two emitter leads of the transistor are connected to ground by low impedance paths and at the ground ends of the tuning and bypass capacitors. Solder the eyelets to the copper ground plane on the bottom of the board.

Where component leads will protrude through the board, the ground plane must be cleared away to prevent shorts. This can be done with a small drill. Clear away enough copper from around the hole to ensure that the component doesn't touch anything but Fiberglass.

Once the fabrication of the board is done, the components are installed as shown in **fig. 4**. The 2N6679 and the chip capacitors are mounted on the top of the board, while the bias circuit components are installed on the ground-plane side. Be careful when soldering the transistor that solder does not flow underneath the package and short out any of the leads. The transistor package is very small, and the leads come quite close together on the bottom side, so that it's easy for even a tiny amount of solder to wick along one lead and short to another. Also, note that the variable capacitor in the output circuit must be mounted right at the output end of the quarter-wave line for maximum performance.

The two emitter leads on the 2N6679 are bent at right angles to the transistor package so that they will fit through the two eyelets to the ground plane side of the board to be soldered. It's important that the emitter leads be connected to the ground plane by low impedance paths if the amplifier is to perform properly. To get a low impedance path, the transistor package should be mounted flush with the PC board so that the emitter leads are as close as possible to the feedthrough eyelets.

The input and output connections are made with small-diameter 50-ohm coaxial cable such as RG-179 or similar. The coax-cable shield is twisted into a pigtail and run through the hole in the PC board near the end of the microstrip matching lines. Then the shield is soldered to the ground plane of the circuit board. On the etched side of the board, the center conductor is soldered to the end of the microstrip transformer. The distance between the end of the coaxial shield and the center conductor soldered connection should be as short as reasonably possible. A length of 0.2 inch (5.1 mm) works well. A jumper wire is run between the two points marked with an **A** in **fig. 4**.

The completed board is installed in a small sheet-metal chassis box by means of four standoffs. The two coaxial cables are run to connectors mounted in the box wall and a feedthrough capacitor is used to bring in dc power for the amplifier.

tuning

Connect the amplifier to a 28-volt power supply and check to see that the supply current is approximately 30 milliamperes. As a further check of the bias, measure the voltage at the 2N6679 collector. It should be about 15 volts above ground.

If you're hesitant about trying out an untested bias circuit with your somewhat expensive rf transistor, you can first test the circuit by soldering a small-signal NPN transistor in place of the 2N6679. The amplifier won't work at 1250 MHz, of course, but the small-signal transistor will duplicate the dc operation of the rf transistor. If the bias circuit properly regulates the collector voltage and current of this "fuse," then it is safe to go ahead and install the 2N6679.

Tuning the amplifier requires a signal source of about 1 milliwatt and a power meter or signal detector of some sort. Apply dc power and the rf drive signal, then tune the output matching capacitor until maximum output is seen. If the amplifier is tuned up at 1275 MHz, it will cover the range 1250-1300 MHz with less than a dB of gain "droop" at the band edges. Once the output capacitor is peaked, no other adjustments are necessary.

2-watt stage

The final stage of this chain is a Class C common-base amplifier capable of 2.5 watts output power. The transistor used in this stage is the Motorola MRF2001, the lowest-power member of a family of 2-GHz, common-base power transistors. The 2000 series transistors have been available from several manufacturers for a number of years and have come to be an unofficial standard type of microwave transistor. They have been used in many military and industrial designs, but their high cost has kept them out of the range of most Amateur experimentation. However, in late 1978, Motorola announced a low-cost version of this 2000 series designed to penetrate the commercial marketplace.

These transistors are rated for 2-GHz operation, and this means that their performance in the 1250-1300 MHz range can be fairly impressive. It's generally difficult to achieve high gain and high power output simultaneously, but when the device is operated well below its design frequency these two goals can be more easily met. Thus, the MRF2001, rated at 1 watt minimum output with 9 dB gain at 2 GHz, when operated at 1.3 GHz, easily produces 2 watts with an associated gain of 13-14 dB. The cost of these transistors is about $19 each.

design

Since this stage is to be operated as a large-signal, Class C amplifier, the s-parameter design approach used to design linear Class A amplifiers is not appropriate. Instead, matching networks must be designed to present to the transistor the impedances that produce the best input match while simultaneously giving the highest output power.

This approach is different in several ways from that used for small-signal linear amplifiers. In the case of a low-level amplifier, the matching networks are chosen to maximize gain, and perhaps to minimize the input and output VSWR. By contrast, in a large-signal amplifier, only the input circuit is designed to provide a low VSWR to maximize the driver power transferred into the amplifier. This part of the design is not wholly unlike that of the small-signal amplifier.

At the output of the Class-C stage, however, the intent is to provide that matching network which maximizes the output power. This requirement usually means that output VSWR and overall power gain are compromised somewhat in the interest of increasing output power and efficiency. A detailed discussion of this design approach and of the tradeoffs involved is provided by Pitzalis.[6]

To permit such a design the manufacturer publishes typical impedance data for their rf power transistors. The input impedance they specify gives a good match and hence a good power transfer into the transistor. The output impedance they list is that into which the transistor delivers maximum power at a given drive level.

As with the 100-milliwatt stage, the fractional bandwidth required for the amplifier to cover this ham band is small enough so that the design can be at a single frequency. This approach is much simpler than would be a broadband optimization.

The data sheet for the MRF2001 lists the eqivalent series input impedance at 1250 MHz as

$$Z_{in} = 7.6 + j10.3 \text{ ohms}$$

The best input match will be obtained with a network that presents the complex conjugate of this impedance to the transistor. Thus, we want to design a network that will transform the real 50-ohm input feedline impedance to

$$Z^*_{in} = 7.6 - j10.3 \text{ ohms}$$

where the asterisk indicates the complex conjugate.

This series input impedance is equivalent to the parallel impedance[7]

$$Z^*_{in} = 21.6 \| -j15.9 \text{ ohms}$$

a resistance of 21.6 ohms in parallel with a capacitive reactance of 15.9 ohms.

At 1250 MHz, a capacitor with this reactance has the value of

$$C = \frac{1}{2\pi f X_c} = 8.0 \, pF$$

If this capacitor is shunted to ground very close to the transistor emitter, the result will be a driving impedance at the emitter of 21.6 ohms, a pure resistance without a reactive component.

To transform this pure resistance to the 50-ohm input impedance, we can use a quarter-wave microstrip transformer whose characteristic impedance is the mean of the two end point resistances, or

$$Z_{input\ line} = \sqrt{(50)(21.6)} = 33 \text{ ohms}$$

This input circuit is shown in the schematic, **fig. 5**.

The required output series impedance is given by the data sheet as

$$Z^*_{out} = 9.6 + j23.1 \text{ ohms}$$

The real part of this series equivalent impedance can be matched with a quarter-wave transformer whose characteristic impedance, as before, is given by

$$Z_{output\ line} = \sqrt{(9.6)(50)} = 22 \text{ ohms}$$

The imaginary term can be dealt with by either a series inductive reactance at the transistor collector or by a shunt capacitive reactance at the transformer output. Since it is easier to obtain good, high-Q, variable capacitors than inductors at high frequency, it's better to choose the latter. The reactance required is equal to

$$X_c = \frac{Z_0^2}{X_s} = \frac{(22)^2}{23.1} = 20.8 \text{ ohms}$$

This corresponds to a capacitor of

$$C = 6.1 \, pF$$

A 1-8 pF trimmer capacitor will provide some adjustment range. The completed output circuit is shown in **fig. 5**.

The matching circuits are built on 1/32-inch (0.8-mm) Fiberglass epoxy board material of the type known as G-10. The 33-ohm input line is 0.110 inch (2.8 mm) wide and 1.14 inches (29.0 mm) long. Since the transistor emitter tab is wider than this matching line, and since a shunt capacitor from the emitter to ground is needed, a portion of the required capacitance is distributed in a short open-ended stub. This stub provides about 2.7 pF of the required 8 pF and it is wide enough so that the emitter lead can be soldered over its entire width.

The 22-ohm output transformer is realized with a line 0.190 inch (4.8 mm) wide and 1.11 inches (28.2 mm) long. The output must have a dc blocking capacitor; this should be a microwave chip-type capacitor. A chip capacitor with a 50-mil (1.3-mm) package width will fit the 0.056-inch (1.4-mm) 50-ohm output line with minimum discontinuities. The variable tuning capacitor must be mounted right at the end of the 22-ohm transformer.

Bias voltages are applied to the transistor through two high-impedance quarter-wave lines bypassed to ground at the end away from the matching networks. Since the input is grounded, no dc blocking capacitor is required. At the output, the supply end of the bias line is bypassed to ground with a chip capacitor in parallel with a higher capacitance ceramic capacitor. The chip capacitor provides effective decoupling at microwave frequencies, while the larger capacitor maintains a low-impedance path to ground down to low frequency.

construction

The construction of this 2-watt stage is complicated somewhat by the fact that both thermal and rf factors must be considered. Good rf circuit techniques are needed so that the potential performance of the amplifier can be achieved, while at the same time the transistor must have a good path for heat transfer, or the heat generated might destroy the transistor or shorten its life. For this reason, the transistor package is designed so it can be mounted to a heat sink. If this is properly done, the transistor junction temperature can be kept low enough so that no damage or degradation will occur.

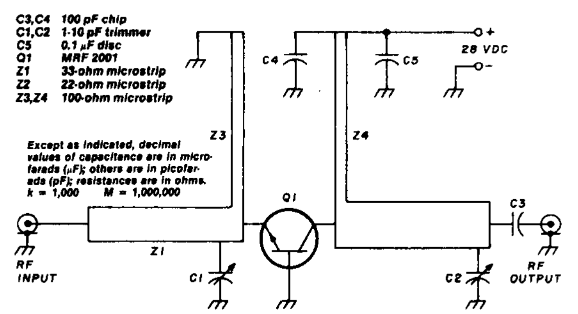


fig. 5. Schematic of the 2-watt amplifier. Simple stripline construction makes for easy assembly. Separate input and output stages are shown, which require two boards.

thermal considerations

Heat is a major enemy of power transistors. The temperature of the transistor die or chip inside the package is the critical factor. For a silicon transistor, the maximum allowable junction temperature is usually specified as 392 F (200 C). However, it is very desirable to operate it at a lower temperature than this, since, as a rule of thumb, the transistor lifetime doubles for every 18 F temperature reduction.[8] If the junction temperature is kept below 302 F (150 C), the average time to failure for a gold-metalized transistor, such as the MRF2001, is measured in decades. Since in Amateur applications a transmitter is used only intermittently, a transistor transmitter operated below this temperature should easily outlive its creator.

The data sheet for the MRF2001 states that the thermal resistance, measured from the transistor junction to the mounting flange of the package, is 45 F per watt of dissipation. If the amplifier has an efficiency* of 50 per cent, then, when it is producing 2.5 watts of rf output, it will also be generating 2.5 watts of heat. This waste heat must be carried away from the die. In this instance the transistor junction will be approximately $2.5 \times 45\ F = 112.5\ F$ hotter than the case of the package. If the case itself is held at a temperature of 122 F (50 C), for example, the junction temperature will be a fully acceptable 243.5 F (117.5 C).

To keep the transistor's case below this target temperature of 122 F (50 C), it must be coupled to a cool heat sink with a connection of low thermal resistance. At this power level, if the transistor flange is bolted to a small finned aluminum heat sink the transistor case temperature will remain well below 122 F (50 C) when the heat sink is located in a normal room temperature environment.

assembly

Because of the flange mounting transistor, the matching circuit is mounted on two separate circuit boards, one for the input and one for the output. The board material is 1/32 inch (0.8 mm) thick type G-10 Fiberglass epoxy double clad with copper. Full-scale illustrations of the boards are shown together in **fig. 6**. The other side of the board is left unetched to serve as a groundplane for the microstrip lines. After etching, the boards are separated along the marked edges. Components are mounted on the boards as indicated in **fig. 7**.

The transistor is bolted directly to a small aluminum heatsink, which is approximately 1.2 by 4 inches (3 by 10 cm) in size. The heat sink is located on the outside of the chassis box with the transistor mounted on it and projecting through the wall of the box in a hole cut with a 1-inch (2.5-cm) chassis punch.

The boards for the input and output circuits are inside the chassis box and are attached to the heat sink with machine screws that run through the box wall. The distance from the bottom of the transistor case up to the bottom of its stripline leads is 0.12 inch (3.0 mm). This distance is taken up by the box wall, which is about 0.05 inch (1.2 mm) thick, by the circuit board thickness of 0.031 inch (0.8 mm), and by a spacer, which is a small piece of the same type of circuit board. Thus the total height is about 0.11 inch (2.8 mm), so that the circuit board traces fit snugly

*Efficiency of rf power semiconductors has been defined in many ways. Here it is taken to be the ratio of the rf output power to the dc collector input power × 100 per cent.

beneath the transistor leads, ready for soldering. A cross-sectional view of this construction method is shown in **fig. 8**. Device outlines are shown in **fig. 9**.

tuning

One of the more challenging problems facing anyone who tries to tune Amateur microwave equipment is to devise test and tune-up procedures that are effective but don't require the use of elaborate test gear, not available to most hams. For instance, if you have access to a microwave rf sweeping signal generator, a spectrum analyzer and a calibrated power meter, you'll not have any difficulty in tuning this amplifier chain for peak performance. Unfortunately, few of the ham shacks I've visited have been quite so well equipped.

When I first tuned up these two units, I made use of a spectrum analyzer and power meter to tune and

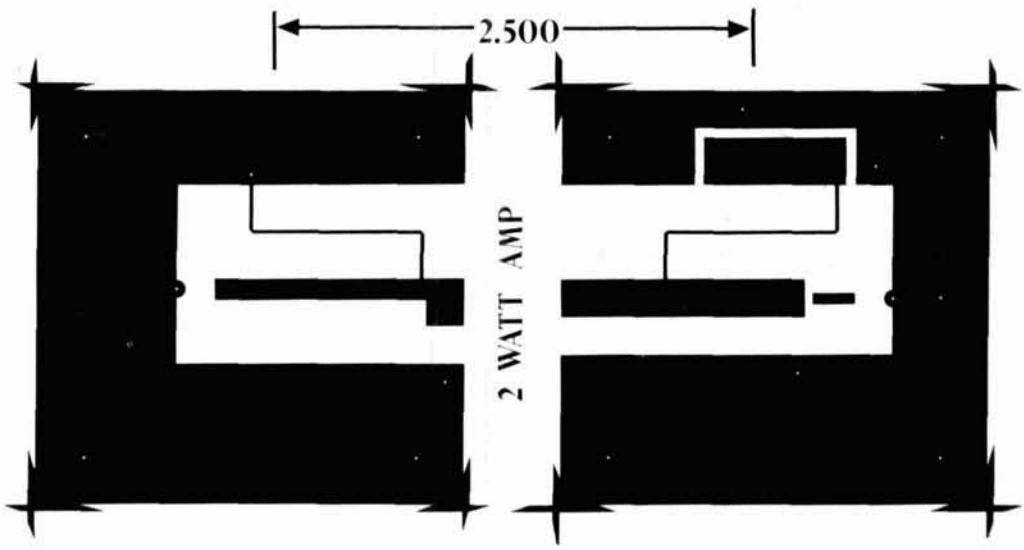


fig. 6. Full-size PC board illustration for the 2-watt amplifier. After etching, the two boards are separated along the marked edges as shown.

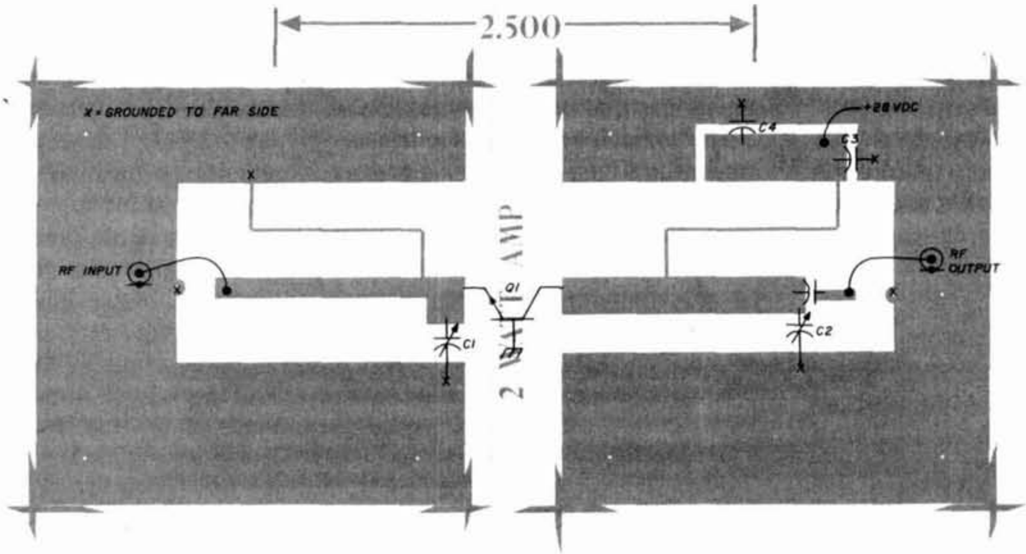


fig. 7. Component placement for the 2-watt amplifier. See text for correct assembly.

to verify that each stage was operating correctly and had clean, spurious-free outputs. Then I detuned them and retuned each at a single frequency using only the power meter, to show maximum signal.

This simplified method worked surprisingly well. Each amplifier had a single peak in output power, a fact that makes tuning with no spectral display easy and not misleading. When the amplifiers are peaked in this way, their bandwidths are wide enough to cover the band and nearly as flat in amplitude response as when they had been tuned with a swept signal source.

To tune up the 2-watt stage using the "no-equipment" approach, you need a signal source that will give a 100-milliwatt output at 1275 MHz, or else at

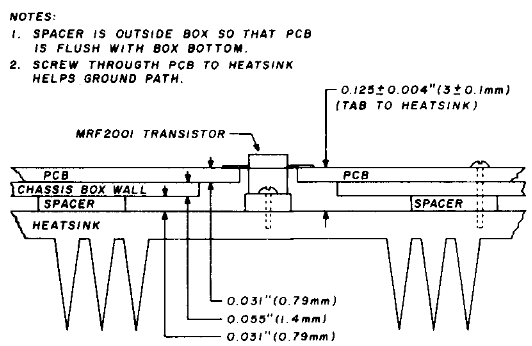


fig. 8. Cutaway view showing transistor mounting detail.

the operating frequency if you prefer to optimize performance there. The first stage driven with a 2-milliwatt signal will serve as a driver, of course. To indicate power, a receiver with a signal-level meter, an rf power meter, an rf millivoltmeter, or a crystal detector may be used.

The output from the final stage should not be fed directly into most types of signal indicators, though, because the power level expected would probably damage them. The signal must be reduced to a level that the power meter can safely handle.

One way to do this is to use a directional coupler as shown in **fig. 10**. The amplifier signal is sent through the directional coupler, and a low-level sample of the signal appears at the coupled port.

Alternatively, the output signal can be attenuated with a resistive network as in **fig. 11**. A suitable attenuator pad can be built to do this. It is a pi-section attenuator with a loss of 20 dB, so that the output of the Class C final stage is reduced in passing through it to a level of about 20 milliwatts. If this is still too much power for your detector, two attenuators can be placed in series.

The attenuator pad can be built using 1/2-watt, 5 per cent carbon composition resistors by following the schematic shown in **fig. 12**. The 1/2-watt resistors are used because, although 2-watt-rated resistors would be better able to stand the heat generated by the amplifier output, their physical size leads to excessive reactance at high frequency.

I realize that any microwave engineer would cringe

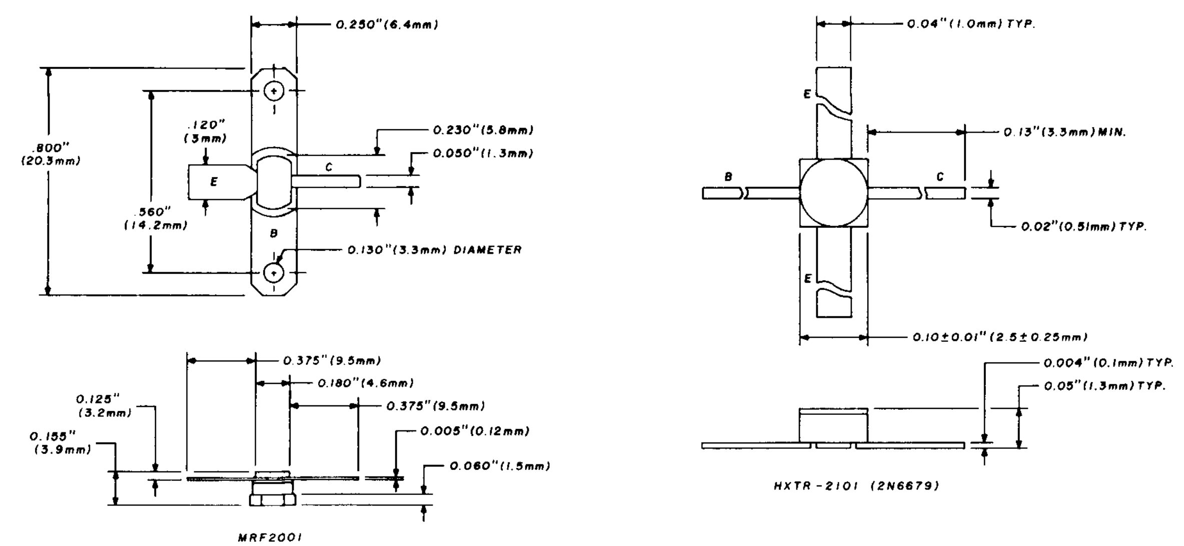


fig. 9. Transistor packages.

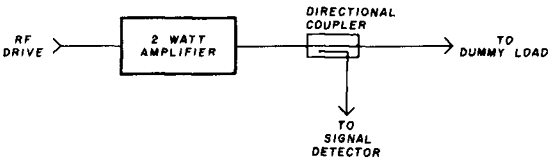


fig. 10. Tuning setup with directional coupler.

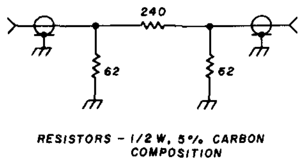


fig. 12. Schematic of the 20-dB pi network attenuator.

at the thought of using such a pi-attenuator at 1300 MHz. However, even though this attenuator is a bit crude, if it's built on a small piece of copper-clad board with the shortest possible lead lengths, it can give an acceptably low VSWR.

This attenuator won't stand 2 watts of rf input for long, so tune rapidly and give the resistors a chance to cool down often. Remember that once heated above a certain point, the resistors will not return to the same resistance they had when new. This will change network attenuation and could lead to errors in tuning.

Once you have the setup together, the hardest part of the task of tuning is done. All that remains is

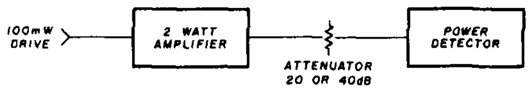


fig. 11. Tuning setup using an attenuator.

to apply the drive signal and peak the two variable capacitors for maximum power output. It's best to go back and forth between the two capacitors a few times; their adjustments do interact a bit.

Fig. 13 shows the final result of this method of tuning. It shows the output of the cascaded system when the input signal is held at a constant 2 milliwatts and swept in frequency from 1250 to 1300 MHz. The output is greater than 2 watts across the band and is nearly 3 watts at the center point.

The final amplifier stage draws about 165 milliamps at 28 volts when producing 2.3 watts output. This indicates that it's operating at approximately 50 per cent efficiency. No attempt was made to tune for better efficiency, although this can often be done.

conclusions

The amplifiers described here are stable, fairly inexpensive, and simple to reproduce. Second stage output power is sufficient to provide solid horizon coverage for voice communication when used with only a moderate gain antenna. The flexibility of the modular construction means the amplifiers can be adapted to a range of uses.

The final stage was biased as a Class C amplifier for use in an fm system. In SSB or ATV service, however, a more linear amplifier is needed. It is likely, as the manufacturer suggests, that the transistor could be operated in Class B, although I have not tried this.

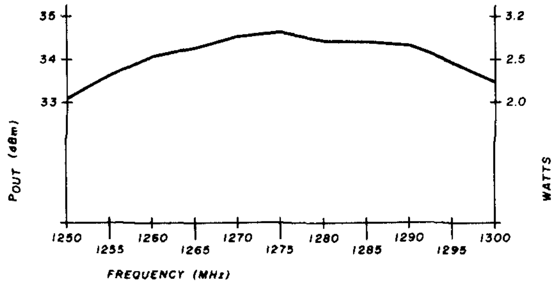


fig. 13. Power output as a function of frequency for the amplifier chain shown in fig. 1. Input is 2 milliwatts.

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More Details? CHECK—OFF Page 98

ham radio TECHNIQUES

Bill Orr W6SAI

Diana to EME in 20 years

Sharp-eared newcomers to the Amateur bands may occasionally hear operators talking about EME schedules. A recent issue of *QST* magazine listed the scores earned by stations engaging in an EME contest. How did it all get started?

For years man has dreamed of reaching the moon. Jules Verne wrote about it. In 1930, a well-known inventor and science writer told how it would be possible to use the moon as a reflector to bounce very short radio waves around the earth (**fig. 1**). But nobody had attempted the task; it was an idle dream. No one was even sure that radio waves would penetrate the ionosphere and reach the moon. Perhaps one day this fanciful idea would take root.

While others dreamed, John DeWitt, Jr., W4ERI, decided to act. A Radio Amateur and avid astronomer, John had combined his hobbies while still in high school, and while a young student in Nashville, Tennessee, he had constructed vhf receiving equipment sensitive enough so that he could hear radio noise from the Milky Way — noise which Karl Jansky had discovered a few years earlier.

fig. 1. A 1930 issue of *Short Wave Craft* magazine featured transmission by moon reflection. A later issue discussed the subject in depth, and noted it was impractical because of the high power needed to overcome path loss. But, the intent was feasible — the article was just ahead of its time.

Finally, in 1940, young DeWitt assembled an 80-watt transmitter, a high-gain antenna, and a sensitive receiver and attempted to send a signal from his transmitter to the moon and back. The experiment was a failure and John set about to determine the cause of his problems. He finally understood that more power, bigger and better antennas and a more sensitive receiver were the keys to success.

The onset of World War II brought all of W4ERI's plans to a stop. Amateur Radio was closed down for the duration. But, as John assumed military service, the idea remained planted in the back of his memory. By then a recognized authority in broadcasting, John rose rapidly in the expanding field of military communications, and by 1944 had risen to the rank of lieutenant colonel and was in charge of the Army's Evans Signal Laboratory at Belmar, New Jersey.

The highly classified work of the laboratory drew to a close in August, 1945, with the abrupt end of the war. DeWitt's important jobs quickly evaporated, but he did not have sufficient discharge points to return to civilian life. He fretted under the boredom and inaction.

project Diana

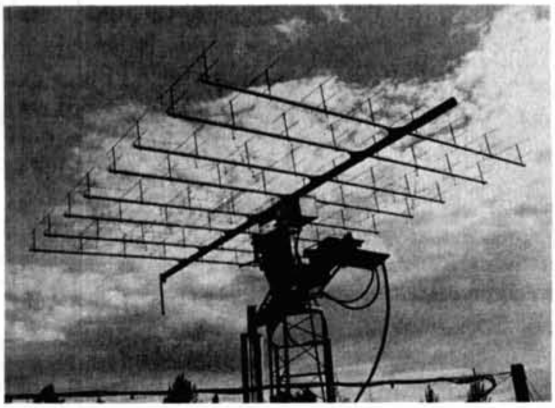
While John bided his time, the Pentagon had a vital question at hand, brought about by the V-2 rocket attacks on London during the war. Was there a way an enemy could direct a rocket attack on the United States using radio-controlled rockets? Could radio and radar signals actually penetrate the ionosphere, or not? Perhaps DeWitt and his remaining laboratory staff could provide the answer.

Once given the go-ahead, John assembled his staff and went to work. Among the group were radio amateurs E.K. Stodola, W3IYF, Frank Elacker, ex-W2DMD, and Harry Kauffman, W2OQU (who would be the first to hear a returned echo from the moon).

The transmitter was an old SCR-271 radar set, much modified, which provided 4-kW output on 111.5 MHz. By combining two radar antennas, DeWitt created an array of sixty-four elements and directors which provided a power gain of nearly 22 dB. His crew built a special superheterodyne receiver of very narrow bandwidth and high sensitivity. All of this, everyday equipment now, was state-of-the-art in 1945.

DeWitt and his crew had to design very special test equipment to make sure their exotic moon-bounce station was performing properly. Calculations showed the project would work and that reflected signals could be detected from the moon — but nobody was *sure* it could be done.

No matter. DeWitt forged ahead, and by January, 1946, he was ready for on-the-air tests. Preliminary tests in December had been inconclusive but he felt that his team was on the verge of success.



The elaborate 432-MHz moonbounce antenna array of ZE5JJ in Rhodesia boasts 128 elements and can track the moon automatically. Peter has since constructed a parabolic reflector using aluminum angle stock and wire screening. It has been very successful.

On the morning of January 10, 1946, DeWitt and his staff fired up the "moonbounce" transmitter. The cumbersome antenna was aimed at the moon, and one-second pulses were sent out every four seconds. Finally, after an agonizing wait, they heard the first echo return on a loudspeaker and saw the returned signal on an oscilloscope connected to the receiver. For the first time, man had touched the moon with an electronic signal and the moon had answered back.

Project Diana was a success and fired the imagination of the public in a manner which may seem surprising in today's more technically sophisticated atmosphere. Soon the U.S. Navy had a microwave moonbounce link from Annapolis to Pearl Harbor. and, very shortly, Amateur moonbounce (EME) circuits would come into being.

amateur moonbounce experiments

It was not until July, 1960, that the first two-way Amateur moon-bounce contact was recorded.[1] It happened on the 1296-MHz Amateur band between W6HB (The Eimac Radio Club) and W1BU (The Rhodedendron Swamp VHF Society). Hank Brown, W6HB, and Sam Harris, W1BU, and their crews worked for weeks getting the equipment ready to go, and finally made contact. Moonbounce communications, using Amateur power levels, was proven possible and it

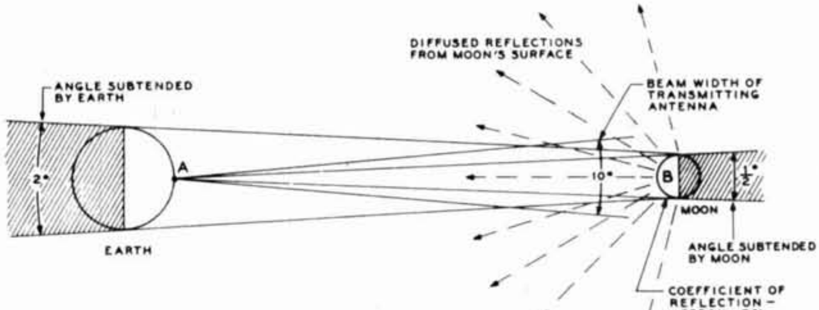


fig. 2. The Earth-Moon-Earth radio circuit. This illustration shows the great obstacles which would seem to make detection of moon-reflected signals highly improbable. The path to the moon and back is about half a million miles and the moon reflects only about 7 per cent of the radio energy striking it. The reflected energy is diffused all over the heavens and only a small portion of the energy which left the radio transmitter is reflected back to earth. Finally, the largest vhf receiving antenna is only a fraction of the earth's pickup area facing the moon. In spite of these staggering difficulties, the earth-moon-earth radio circuit can be made to work with equipment well within the capabilities of many vhf-minded radio Amateurs (drawing courtesy of *Radio Publications, Inc.*).

now remained only to see if other enterprising Amateurs would follow the lead established by these two radio clubs.

Interest grew slowly (possibly because the experiments had been done in a little-used and relatively unknown ham band), and it was not until 1964 that Amateur interest in EME communications was awakened with a jolt when Bill Conkel, W6DNG, ran 2-meter schedules via the moon with Lenna Suomienen, OH1NL, of Finland. Here was real moonbounce with everyday Amateur vhf equipment.

Since those early years, EME interest has grown until today hundreds of Radio Amateurs maintain schedules and experimental contacts via the moon on 144, 220, and 432-MHz.[2] At last, EME has come of age.

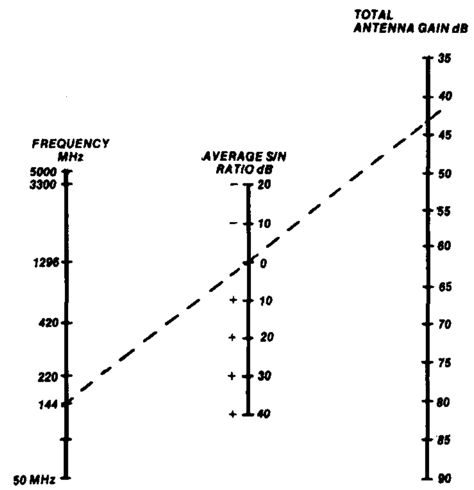


fig. 4. Moonbounce Nomograph provides guideline to successful moonbounce QSO. The graph is based upon 590 watts transmitter power output, zero decibel receiver noise figure, and 100-Hz bandwidth. Lay a straight edge across any two columns and read the desired unknown in the third column. The antenna-gain figures represent a compromise between calculated gain required based upon free space losses and the experience of successful moonbounce experimenters. At 144 MHz, for example, for an average signal-to-noise ratio of zero decibels, a total antenna gain (for both ends of the path) is about 42 decibels. Two 22 decibel gain antennas should do the job.

the EME path

What kind of equipment and antennas does it take to establish a moonbounce station? Who can be worked? Does all the work have to be done at night?

The moon is about 2160 miles in diameter and orbits the earth at a distance that varies from 221,463 to 252,710 miles. An orbit takes about twenty-eight days and is somewhat eccentric, so that the moon travels across a different segment of sky each night of the lunar month. And, although the moon looks quite large when it is full, it subtends an arc of only about 0.5 degree when seen from the earth (**fig. 2**). Even the highest gain Amateur vhf antenna has a beam width much greater than this, consequently only a small portion of the signal aimed at the moon actually strikes it; the rest passes out into limitless space. Furthermore, an estimated 93 per cent of the signal that does strike the moon is absorbed. Also, as our astronauts verified, the moon surface is exceedingly rough; thus the 7 per cent of the reflected energy is diffused all over space.

Viewed from the moon, the earth subtends an arc of about 2 degrees, and the vhf signal that returns to earth is spread over half the earth's surface, or an area of about 98,470,000 square miles. Clearly, only a small fraction of the transmitted energy ever reaches the eager ears of the moonbounce listener.

EME path loss	distance miles (km)	50 MHz (dB)	144 MHz (dB)	432 MHz (dB)	1296 MHz (dB)	2400 MHz (dB)
Perigee	221,463 (356,334)	177.89	187.08	196.62	206.17	211.43
Apogee	252,710 (406,610)	179.03	188.21	197.76	207.21	212.56

fig. 3. Free-space path loss for earth-moon-earth circuit at perigee (221,463 miles) and apogee (252,710 miles). The nominal 1.14 decibel difference in signal loss between moon perigee and apogee becomes 2.28 decibels for the round trip to the moon and back.

the EME path loss

Because radio signals travelling through space are attenuated at the square of the ratio of the frequency, the path loss to the moon and back is 8.3 times (9 dB) greater on 144 MHz than on 50 MHz. A similar increase in loss occurs between 144 MHz and the 420-MHz band, and between the 420 and the 1250-MHz band (**fig. 3**). In addition, transmitter efficiency and receiver noise-figure both tend to become worse with increasing frequency. Thus, there are compelling reasons to use as low an operating frequency as possible for EME work.

On the other hand, the power gain of a beam antenna of a given *size* increases by the same ratio that the path loss increases and, because the antenna gain is realized in both transmission and reception, there is a net circuit signal gain with increase in frequency, even after deducting the increased circuit losses.

The various factors point to the use of the 144 MHz and 430 MHz Amateur bands for EME operation: good equipment is available, antenna size is not too great, and there's enough international activity on these bands to make the investment in time and effort worthwhile. On 2 meters, the portion of the band between 144.00 and 144.10 MHz is an international segment, and it is in this region that a great deal of serious moonbounce activity takes place.

how much power? how big an antenna?

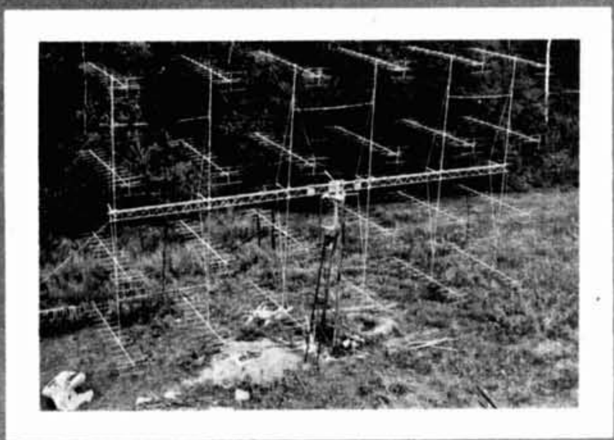
Fig. 4 provides a guideline to successful moonbounce communications. The scale at the right-hand side of the nomograph labeled Total Antenna Gain shows, for example, that if two stations having a combined antenna gain of about 43 dB and are satisfied with an average signal-to-noise ratio of zero dB, they can achieve two-way communications.

This graph is based upon a transmitter power output of 590 watts, a zero-dB noise figure, and a receiver bandwidth of 100 Hz. I'm sure that no moonbounce station fits these esoteric requirements, but hundreds of them come close, and some may surpass these figures. This is how it is done in the real world.

The Total Antenna Gain scale tells us that if one EME station is equipped with a 26-dB-gain antenna, the station at the other end of the circuit will need only 17 dB antenna gain, all other things being equal. *The greater the antenna gain at one end, the smaller the array needs to be at the other end.* On 2 meters, there are many stations equipped with high-gain antennas for meteor scatter and other forms of long-distance communication. Long Yagi beam antennas are available from several manufacturers, and many will provide a power gain of 17 dB. Two of them properly arranged can provide about 20 dB power gain. Again, all else being equal, a station equipped with a 20-dB antenna array can theoretically contact another station which has an antenna array with a power gain of 23 dB. And, if contacts are made on a rising or setting moon, the stations can take advantage of ground reflection of the signals to pick up an additional few dB of signal strength.

Moonbounce for everyone. That is just about possible using the new antenna recently assembled by Cushcraft's Chief Engineer, David Olean, K1WHS. Dave's antenna (below) makes it possible for him to contact stations with single Boomers and only modest power. Recently, he contacted Dave Redman, G4IDR, who was using a single Cushcraft 32-19 Boomer and approximately 200 watts at his antenna. Many other hams with single Yagis from Europe and the U.S. have had their first moonbounce contacts this year with K1WHS.

Dave Olean's new antenna consists of 24 Cushcraft Jr. Boomers model 214B. These Boomers are only 15 feet (4.6 m) long. The total antenna gain is about 26 dBd including feedline losses. This is more than sufficient to cover the 450,000 miles (725,000 km) round trip to and from the moon.



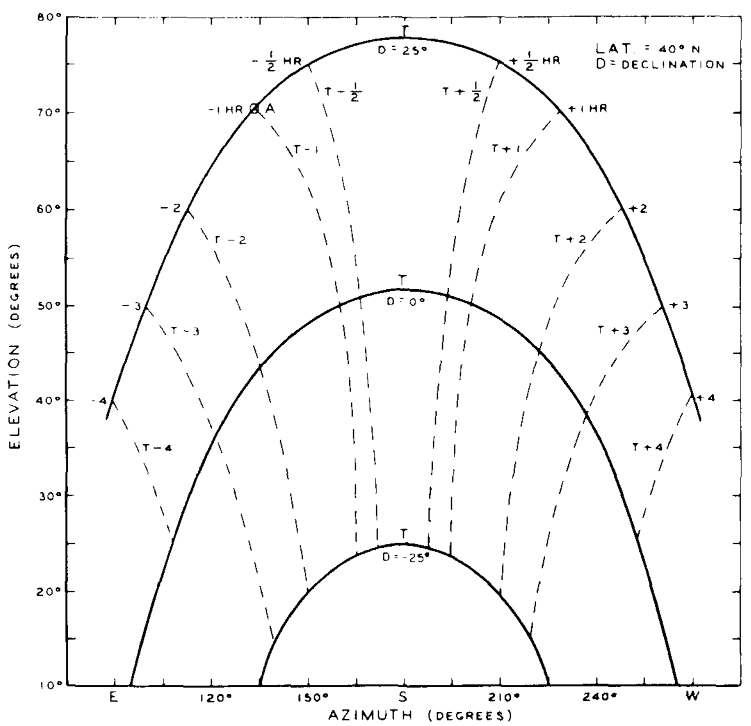


fig. 5. Moon position from center of U.S may be determined with this chart. The altitude of the moon above the horizon and its azimuth change minute by minute every day, but they repeat each lunar month. The "Nautical Almanac," or similar manuals, predict the moon position far in advance. This graph is plotted for a position of 40° North, with the observer facing south. At time T the moon is due south of the observer. At T-1 (one hour before T), the moon is at point A. Given the date, the declination, and the local time that the moon appears due south, the azimuth and elevation may be found. Curves for other values of declination can be interpolated and drawn in between these three curves. (Reference: Lund, "How High the Moon", *QST* July, 1965).

Today, stations using relatively modest antennas and everyday equipment are enjoying EME contacts on 2 meters. Many vhfers who have good antennas and high power are inadvertently bouncing their signals off the moon without even knowing it.

I recently spoke on the telephone with Dave Olean, K1WHS, who is a prominent 2-meter operator and well-known moon-bounce enthusiast. Dave has an elaborate antenna system, and tells me that he can aim his array at the moon and tune around 144.1 MHz during periods of high activity and hear Amateurs working each other. Their antennas are positioned so that the moon sweeps through the beam and their signals are reflected back to earth to the waiting ears of K1WHS. Dave guesses that there are probably over a thousand vhf stations in the United States who can work moonbounce but *don't know they have the capability*.

I asked Dave what it took to become an EME experimenter. He told me that if a station had a good 2-meter transceiver (such as the TS-700 Kenwood, or equivalent) a low-noise preamplifier, and a good, high-gain Yagi on a 15-foot boom, *he would hear* moonbounce signals, provided he aimed the antenna at the moon. Hearing signals is the first step to working stations.

the nitty-gritty

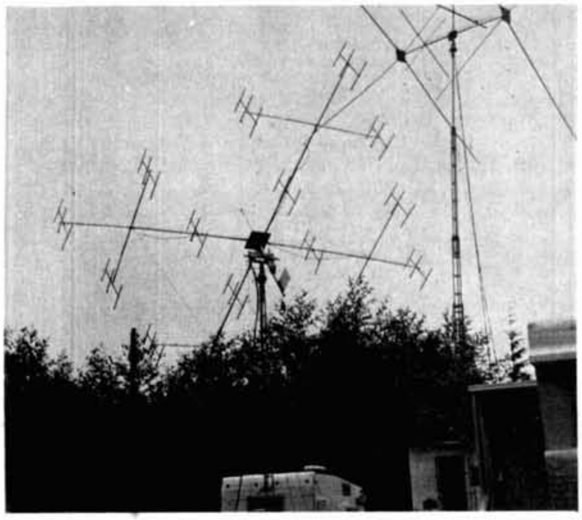
The nice thing about moonbounce is that you can let the other fellow do most of the work. The more antenna gain and power the other fellow has, the less you need. And there are enough serious-minded EME experimenters on the air today that a beginner can get in the game and talk to the big guns without having to make a large expenditure of time, money, or effort.

Of course, for the serious experimenter, there's a lot more to it than just hearing signals. Most moonbouncers have antennas that can track the moon — sometimes automatically. Computer-oriented Amateurs have complete tracking programs worked out showing the position of the moon a year in advance, and have antenna controls that permit automatic moon tracking. But you don't need all of this to get started. Many Amateurs have fixed antennas placed in such a position that the moon will sweep through the beam during the time the moon is mutually visible to operators at both ends of the chosen path. A typical moon-position chart, such as shown in **fig. 5** is useful in positioning the antenna.

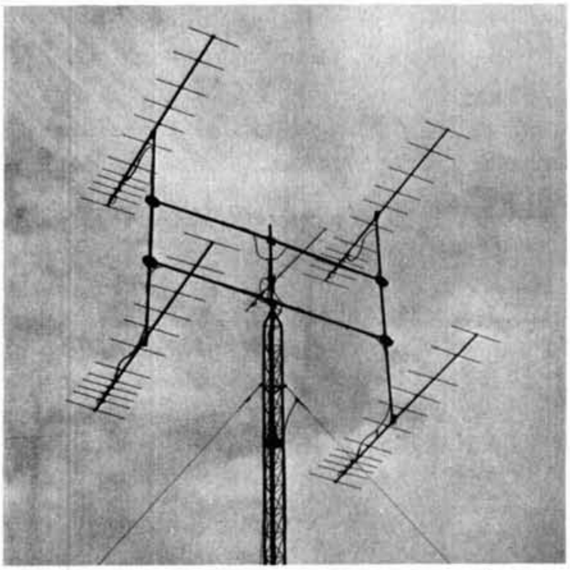
Other factors enter the picture, as the moonbouncer soon discovers. Because the moon moves toward or away from the earth at speeds up to 980 miles per hour, *Doppler shift* changes the frequency of the moon-reflected signal. At 144 MHz, the Doppler shift can be as large as 427 Hz. When the moon is rising, the received frequency goes up; when the moon is setting the frequency goes down. Frequency shift is minimum when the moon is perpendicular to the observer.

what does the signal sound like?

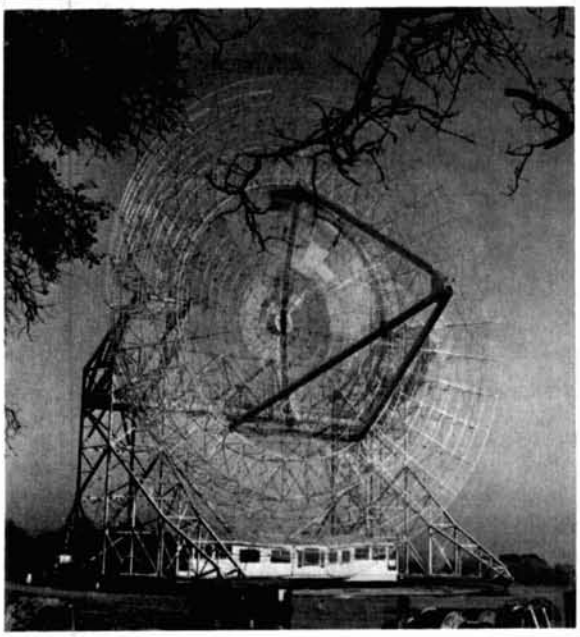
It takes a radio signal slightly over 2 seconds to make the trip to the moon and back, so the return echos of your transmission can be easily received. The best way of testing an EME cir-



The portable moonbounce station of K6YNB/KL7 (now N6NB) puts Alaska on the map. Note that the 2-meter array (left) is composed of sixteen 3-element quads.



Moonbounce array for 2 meters at WA1FFO is composed of four Yagis, each having 12 elements. All of these big antennas make it easier for you to work moonbounce because the other station does most of the work!



The radio telescope at Stanford University in California is occasionally used for moonbounce experiments in Amateur bands. You can really hear this 200-foot-diameter giant when it is on the air! The whole antenna and control building rotates on a circular metal track. Dominating the skyline above Palo Alto, California, the dish can be seen for miles.

cuit, in fact, is to listen to the return echos of your own transmitter. It is quite an eerie feeling to send CW signals and hear them bounce back at you a short time later. Voice signals returning from the moon have a hollow quality about them that is instantly recognizable to a moonbounce enthusiast.

Plenty of stations are active *today*. A good guess is that there are over 350 moonbounce stations active, in all continents, on the 2-meter and 432-MHz bands. Many of them maintain schedules with each other, but on an active weekend plenty of moonbouncers call CQ to raise another moon-bounce enthusiast. During an EME contest, or other weekend of high vhf activity, there is actually QRM among the many moonbounce signals.

The second article in this series will carry more specific information about moonbounce equipment and activity for the 2-meter enthusiast. Suffice to say that if you have a high-gain Yagi antenna, a low-noise receiver, and sufficient know-how to aim your antenna to the moon, you could be hearing moonbounce signals before the next issue of this magazine reaches you.

If you want additional information about moonbounce, write to me and ask for the booklet "Almost Everything You Want To Know About Moonbounce." It is a reprint of important magazine articles on the subject. Send 30 cents in stamps to cover mailing to: William Orr, c/o EIMAC, 301 Industrial Way, San Carlos, California 94070.

references

1. The full story of John DeWitt is told in the May, 1946, issue of *QST* magazine, and also in the May, 1980, issue of *IEEE Spectrum*. The story of the first two-way Radio Amateur contact via the moon is told in the September, 1960, issue of *QST*.

2. Moonbounce contacts have been made on all Amateur bands between 50 and 2400 MHz. In addition, experimenters at Stanford University in California made moonbounce experiments in the 10-meter band using an array of log periodic antennas 1200 feet long and 75 feet wide.

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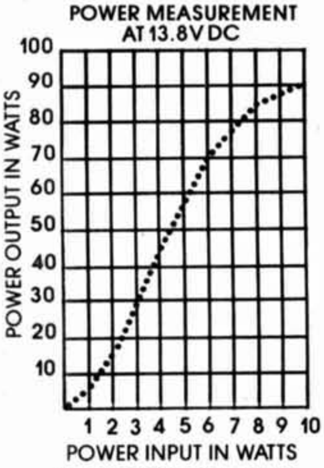

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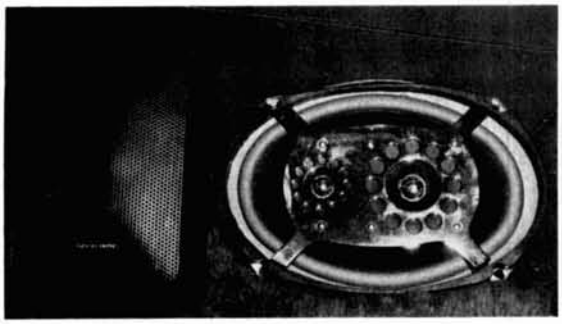

High-quality speakers make a big difference. Speaker is a Radio Shack *Minimus 7* (40 watts peak). Included are mounting brackets that could be used for under-dash mounting. A typical enclosure is shown at left.

better audio for mobile operation

Some suggestions for improved operating in your car

By Ken Glanzer, K7GCO, 202 South 124 Street, Seattle, Washington 96168

The concepts described in an earlier *ham radio* article[1] resulted in many letters to me. The same ideas in that article apply to mobile operation, only more so. When working mobile, you have outside road noise to contend with as well as ignition and other noises. Mobile rigs are now physically smaller; therefore the speakers are smaller, and so it goes. Regardless of speaker size, the audio systems in most of the mobile rigs are deficient.

When the audio gain control is turned up to overcome the various noises, the signal is distorted and most aren't loud enough, even if they aren't distorted. A few sets with small speakers in a small box have produced audio like you've never heard; so it is possible to get good audio from a small box.

loudspeakers

One approach to the mobile audio problem is to use a large speaker and mount it in the car headrest as shown. Lower volume levels are then needed, and this system works well. Use the largest speaker possible with a tweeter.

Another approach is to use a separate, low-level input, 12-volt audio amplifier connected to the top of the volume control as described in reference 1, with a built-in frequency equalizer to emphasize highs. You can also use an audio booster in conjunction with stereo fm cassette radios. The output of the ham rig can be connected in parallel with one channel or, preferably, switched with a dpdt switch. You then get the needed power gain and frequency-response control. Use high-quality, *large* speakers with mid- and high-range tweeters. My booster has a control to fade in two speakers in the front of the car. These speakers can be mounted under the seat, in the door, or in the ceiling headliner. The main stereo speakers are at the car rear, but the headrest speakers are much closer to your ear. You now will have a wide range of combinations, which produce beautiful and *understandable* audio with no distortion and shaped to your ears' deficiencies by the equalizer.

A suggested hookup is shown in **fig. 1**. I also use stereo earphones, parallel connected when used with the ham set. The new Sennheiser Model 420 is ideal, as it has an open earpiece and therefore doesn't block out road noises. These units are lightweight and can be used for long periods of time. They can, of course, be used with the stereo fm radio.

tone control

A welcome addition to any rig is a tone control that *emphasizes* the highs rather than attenuating them. The average person's ears are being bombarded with

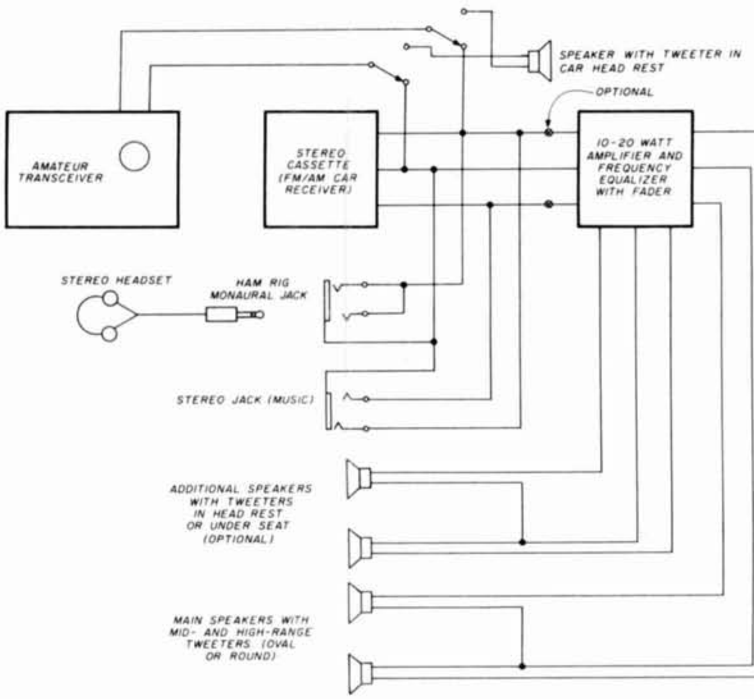

fig. 1. Basic setup for a near-ideal audio system in your car. Scheme features undistorted audio and plenty of volume for mobile operation.

high levels of noise; the loss of high-frequency response is occurring at an earlier age.

I frequently remove the capacitors between the base and collector in the final transistor audio stage, or any other component that cuts the highs. Or, I may alter the values of some components. Many

Speakers mounted in the car headrest brings them closer to your ears; less volume is needed. Stereo earphones, monaurally connected, and with open ear pieces, are comfortable for long listening periods and don't isolate you from outside noise (such as police sirens).

Stereo amplifier and frequency equalizer connect directly to speaker output of your ham rig and fill the car with clear, undistorted audio (with proper speaker). You can change frequency response for enhanced intelligibility.

have said, "I can understand what they say on your radio." Many Amateurs are not deaf from the standpoint of *level*, but are "tone deaf." They can't hear the high frequencies. Sometimes, a small tweeter can be installed near one ear with good results.

reference

1. Ken Glanzer, K7GCO, "Better Audio for New or Old Communications Receivers," *ham radio*, April, 1977, page 74.

ham radio

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- Mobile mounting bracket with quick-release levers.
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OPTIONAL ACCESSORIES:

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- Keyboard selection of 144.000-147.995 MHz in 5-kHz increments. No "5-UP" switch needed.
- UP/DOWN manual scan in 5-kHz steps from 143.900 to 148.495 MHz.
- 10 memories. Retained with battery backup. "M0" memory may be used to shift transmitter to any frequency for nonstandard-split repeaters.
- Built-in autopatch DTMF (Touch-Tone®) encoder, using all 16 keyboard buttons.
- Automatic memory scan.
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repeater security

A combination lock using CMOS devices

The solution was a nonvolatile control. The problem was a secure, remote controlled electronic combination lock for the local repeater; one which would not forget if it was turned on or turned off if the power should fail momentarily, or if the supply line should suddenly become filled with electronic noise. With thoughts of security, simplicity, and ultra-low power drain in mind, the circuit shown here was devised.

features

The control, designed to operate in conjunction with standard TTL tone decoders, uses a sequence of three digits to effect the turn-on function and a different three digit sequence for turn-off. The digits must be applied precisely in the selected sequence to achieve the desired function. The turn-on process automatically prepares the control to accept the turn-off sequence and, similarly, the turn-off process prepares the control to accept the turn-on sequence. To safeguard against the possibility of electronically picking the combination lock, provisions were made to accept digits unused in either control sequence as reset inputs. The entire control uses four ICs, four transistors, and ten diodes.

circuit

Referring to **fig. 1**, the combination-lock function is accomplished by a series of D flip-flops. U1A, U1B, and U2A comprise the turn-on function, and U2B, U3A, and U3B comprise the turn-off function. Activation of either function is accomplished by clocking the flip-flop string in sequence. Beginning with U1A (or U2B), each flip-flop enables the succeeding one in the string by taking its D input to a high logic level. In this manner, U2A cannot be clocked until U1B has been clocked, which in turn cannot be clocked until U1A has been clocked. After completion of the turn-on sequence, in the proper order, the falling edge of U2A's $\overline{Q}$ output clocks the

By Steve Cerwin, WA5FRF, 3911 Pipers Court, San Antonio, Texas 78251

latch comprised by gates U4C and U4D to its ON state. This negative-going pulse also is inverted and coupled to the reset inputs of flip-flops U2B, U3A, and U3B, which prepares them for the turn-off sequence.

In the ON state, U4C output saturates the triple-darlington Q1, Q2, and Q3 through a 1 megohm resistor. The open collector output of Q1 may be used to control an external relay. The base lead of Q3 has been brought out on the printed-circuit board for the possibility of using an external master override for either the ON or OFF function.

The unused digit inputs are connected in OR fashion to Q4 through diodes CR1 through CR8. If any one of these inputs is selected, Q4 will saturate, taking one input of gates U4A and U4B to a logic zero level. This will reset both flip-flop strings to their zero state, thereby effecting the anti-picking feature by neutralizing any accidental progress made by non-authorized attempts to operate the control. Assuming different digits are selected for each sequence input, the odds against someone happening upon the proper combination for either control operation by chance are 16 × 15 × 14 or 3,360 to 1 for a sixteen-button pad. The odds may be increased simply by adding additional data latches to the string.

Nonvolatility was achieved by designing the circuit for ultra-low power drain. All the ICs are CMOS, and the total supply current for all four chips is less than 1 microamp. The power hog in the unit is the base current required to saturate the triple-darlington. This current, 2.9 microamps, is esentially the total supply current drain when the control is on. With only the onboard 100 μF capacitor across the supply, the con-

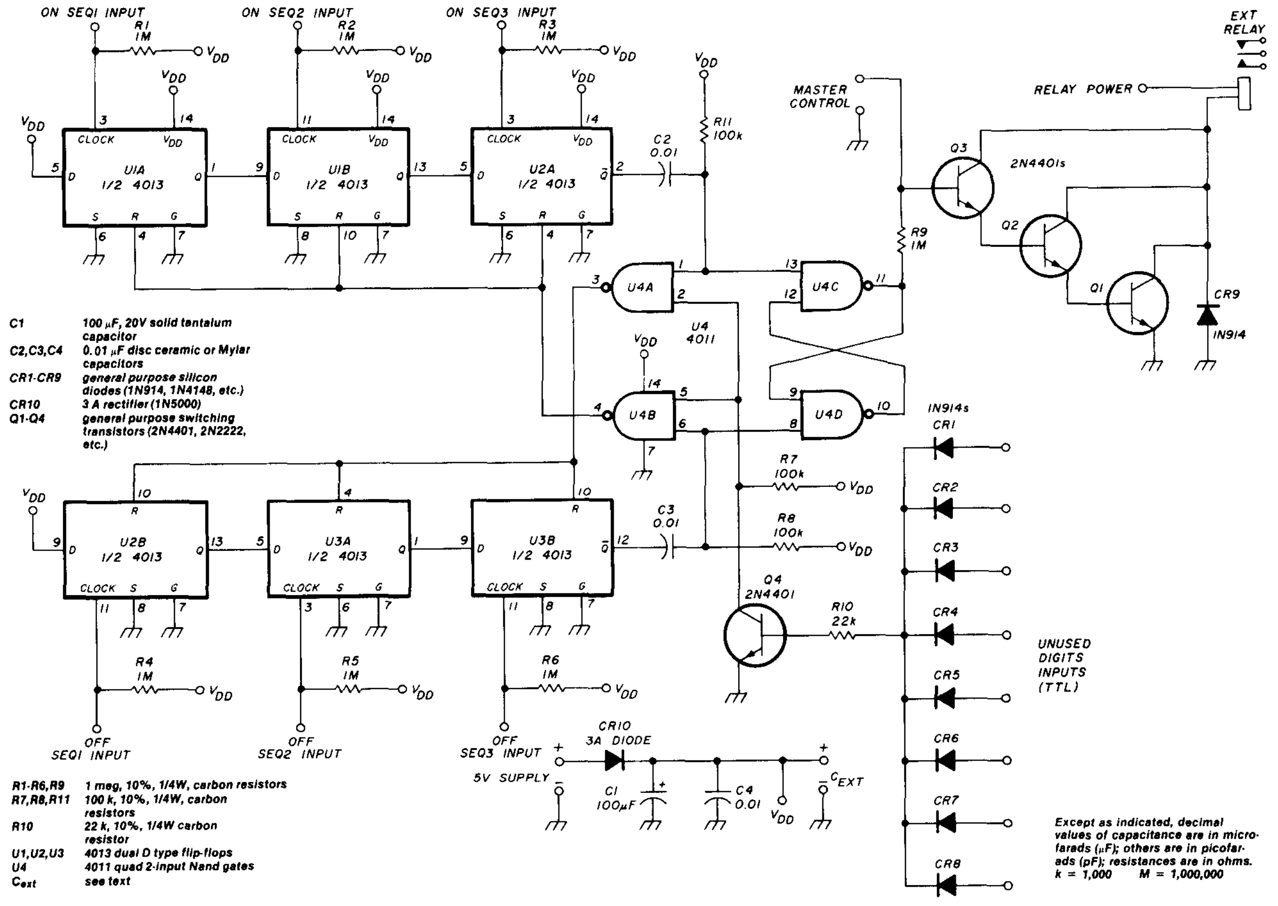


fig. 1. Schematic of the nonvolatile control for repeater security. The odds against someone finding the proper combination for control operation by chance are 3,360 to 1 for a 16-button pad. System uses CMOS devices with a total supply current drain of about 3 microamps.

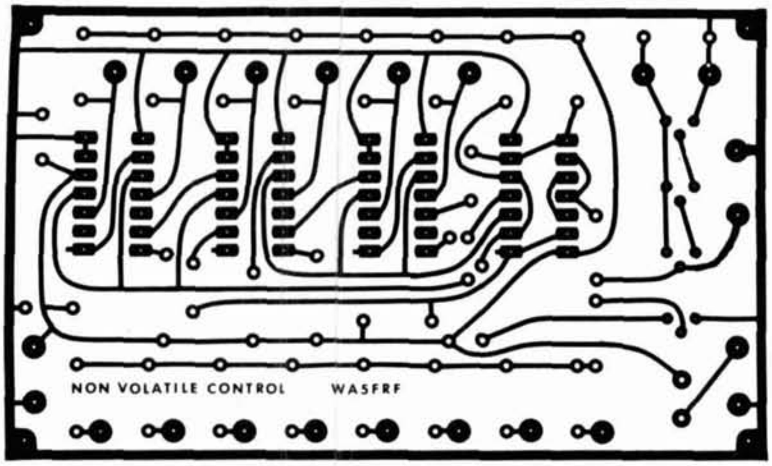


fig. 2. Single-sided PC board layout for the repeater control.

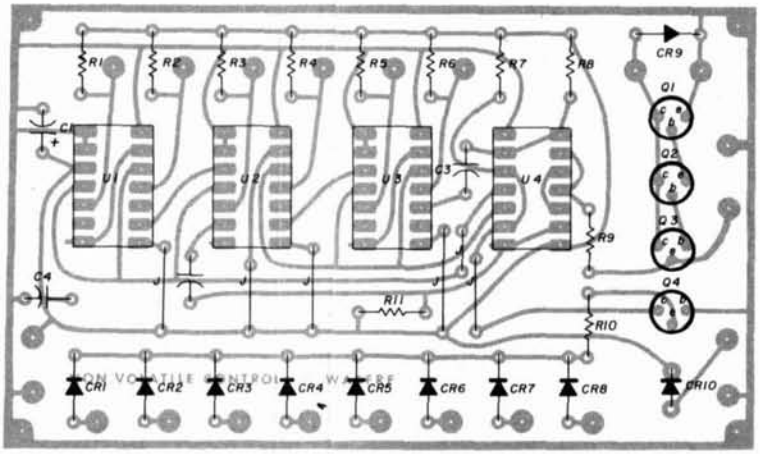


fig. 3. PC-board parts placement.

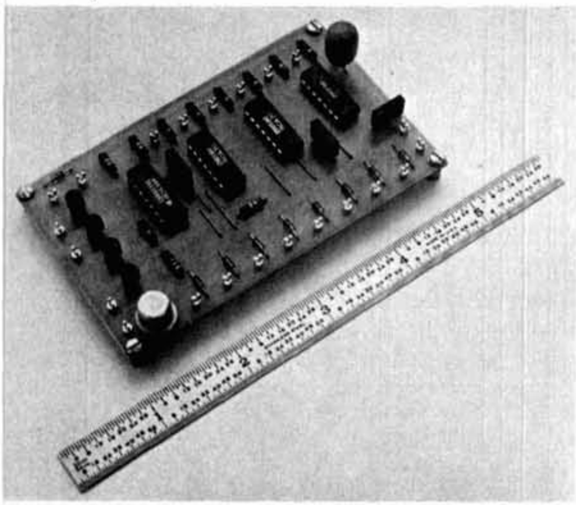


fig. 4. Photo of the completed control circuit.

trol will be nonvolatile for several minutes after power failure. This time may be extended to hours if a larger capacitor is supplied externally, or it may be extended indefinitely if an external battery pack is supplied. The control supply bus is isolated from the supply line by diode CR10. CR10, a 3-amp power rectifier, was chosen to withstand the charge-up current of the large supply bypass capacitor and also provides reverse supply protection. Supply voltage must be 5 volts if the unit is to be compatible with TTL tone decoders.

A single-sided printed circuit board layout for the control is shown in **fig. 2**. Component placement and a photograph of a completed board are in **figs.3** and **4** respectively. To maintain the ultra-low-power drain characteristics, the assembled board must be kept reasonably clean.

ham radio

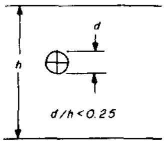
transmission-line circuit design

Using distributed resonant circuits for VHF/UHF transmission lines

Part 2 of this article, which appeared in the January, 1981, issue of *ham radio*, dealt with the geometry of the first four of twelve common transmission line configurations. In this, part 3 of the article, another set of four transmission line configurations will be examined: circular wire between planes, parallel wires over a plane, circular wire in an open trough, and parallel wires between planes/rectangular box.

Part 4 of this series will deal with the remaining four configurations. Part 5, the last in the series, will provide a summary of what has been discussed and show a design example for a 2-meter amplifier.

circular wire between planes



The formulation for the characteristic impedance of this line is similar to that for a single wire over a parallel plane (reference 4):

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \frac{4h}{\pi d}\, ; \; d/h < 0.75 \qquad \textbf{(25)}$$

where Z_0 = line impedance (ohms)
ϵ_r = dielectric constant
h = distance between planes
d = wire diameter centered between planes

Fig. 12 shows Z_0 versus h/d for a reasonable range

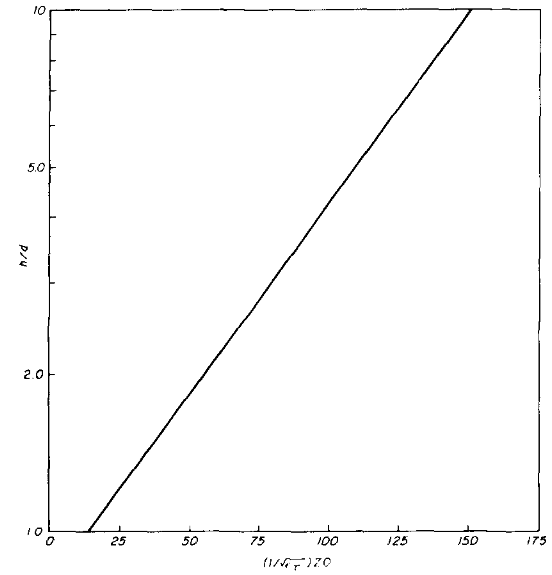


fig. 12. Z_0 versus h/d for a wire centered between parallel planes.

By H.M. Meyer, Jr., W6GGV, 29330 Whitley Collins Drive, Rancho Palos Verdes, California 90274

table 20. HP-67/97 program for calculating Z_0 and h/d for a wire centered between parallel planes.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	030	3	03
002	STO0	35 00	031	8	08
003	$\sqrt{X}$	54	032	÷	– 24
004	STO1	35 01	033	STO3	35 03
005	RTN	24	034	*LBL3	21 03
006	*LBLB	21 12	035	RCL1	36 01
007	÷	– 24	036	X = 0?	16-43
008	*LBL1	21 01	037	GSB8	23 08
009	STO2	35 02	038	x	– 35
010	4	04	039	10^x	16 33
011	x	– 35	040	4	04
012	Pi	16-24	041	÷	– 24
013	÷	– 24	042	Pi	16-24
014	LOG	16 32	043	x	– 35
015	1	01	044	STO2	35 02
016	3	03	045	R/S	51
017	8	08	046	*LBLD	21 14
018	x	– 35	047	GTO1	22 01
019	STO3	35 03	048	*LBL9	21 09
020	*LBL2	21 02	049	1	01
021	RCL1	36 01	050	STO1	35 01
022	X = 0?	16-43	051	RCL3	36 03
023	GSB9	23 09	052	GTO2	22 02
024	÷	– 24	053	*LBL8	21 08
025	STO4	35 04	054	1	01
026	R/S	51	055	STO1	35 01
027	*LBLC	21 13	056	RCL3	36 03
028	STO4	35 04	057	GTO3	22 03
029	1	01	058	R/S	51

of values. **Table 20** is the HP-67/97 program for calculating Z_0 or h/d or, given L and d separately, calculating Z_0. **Table 21** identifies the storage registers; **table 22** shows program control.

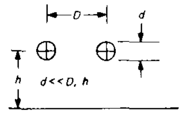
table 21. Register contents for HP-67/97 program for calculating Z_0 and h/d for a wire centered between parallel planes.

STO 0	ϵ_r
STO 1	$\sqrt{\epsilon_r}$
STO 2	h/d
STO 3	INTERIM
STO 4	Z_0

table 22. HP-67/97 program control for calculating Z_0 and h/d for a wire centered between parallel planes.

enter	ϵ_r	press A	
calculates	Z_0		
enter	h	press ENTER	
enter	d	press B	
calculates	h/d		Note: If no value for
enter	Z_0	press C	ϵ_r is entered, program
calculates	Z_0		assumes $\epsilon_r = 1 = air$.
enter	h/d	press D	

parallel wires over a plane



The common application of this line is for push-pull tank circuits with lines spaced much more closely to the ground plane than to the cover. The formulation (reference 4) is:

$$Z_0 = \frac{69}{\sqrt{\epsilon_r}} \log_{10} \left\{ \left(\frac{4h}{d}\right) \left[1 + \left(\frac{2h}{D}\right)^2\right]^{1/2} \right\} \quad d << D, h \qquad (26)$$

where Z_0 = line impedance (ohms)
ϵ_r = dielectric constant
h = centerline height of wires over plane
d = wire diameter
D = center-to-center spacing between wires

For analysis this equation may be reformulated to

$$Z_0 = \frac{69}{\sqrt{\epsilon_r}} \left\{ \log_{10} \frac{4h}{d} + \log_{10} \left[1 + \left(\frac{2h}{D}\right)^2\right]^{1/2} \right\} \qquad (27)$$

which permits solution if h/d and Z_0 are known, or if h/D and Z_0 are known.

When solving for h/D when Z_0 and h/d are known, **eq. 27** may be rearranged for easy solution on the HP-67/97 to

$$\log_{10} \left[1 + \left(\frac{2h}{D}\right)^2\right]^{1/2} = \left(\frac{Z_0 \sqrt{\epsilon_r}}{69}\right) - \log_{10} \left(\frac{4h}{d}\right) \qquad (28)$$

Eq. 28 is transposed when solving for h/d with Z_0 and h/D known.

Fig. 13 is a plot of Z_0 versus h/d for various values of h/D. **Table 23** provides an HP-67/97 program for calculating various combinations of knowns and unknowns. **Table 24** identifies the storage registers; **table 25** shows how the program is controlled.

table 23. HP-67/97 program for calculating Z_0, h/d, and h/D for two parallel wires over a plane.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	029	9	09	057	10^x	16 33	085	9	09
002	STO0	35 00	030	x	−35	058	X^2	53	086	÷	−24
003	$\sqrt{X}$	54	031	STO5	35 05	059	1	01	087	RCL5	36 05
004	STO1	35 01	032	RCL1	36 01	060	−	−45	088	+	−55
005	RTN	24	033	X=0?	16-43	061	$\sqrt{X}$	54	089	10^x	16 33
006	*LBLB	21 12	034	GSB9	23 09	062	2	02	090	4	04
007	R↓	−31	035	÷	−24	063	÷	−24	091	÷	−24
008	÷	−24	036	STO6	35 06	064	STO3	35 03	092	STO4	35 04
009	STO2	35 02	037	R/S	51	065	R/S	51	093	R/S	51
010	1/X	52	038	*LBLC	21 13	066	*LBLD	21 14	094	*LBLE	21 15
011	STO3	35 03	039	STO4	35 04	067	STO3	35 03	095	STO3	35 03
012	R↓	−31	040	R↓	−31	068	R↓	−31	096	R↓	−31
013	÷	−24	041	STO6	35 06	069	STO6	35 06	097	STO4	35 04
014	STO4	35 04	042	RCL4	36 04	070	RCL3	36 03	098	GTO1	22 01
015	*LBL1	21 01	043	4	04	071	2	02	099	*LBL9	21 09
016	RCL3	36 03	044	x	−35	072	x	−35	100	1	01
017	2	02	045	LOG	16 32	073	X^2	53	101	STO1	35 01
018	x	−35	046	ST-5	35-45 05	074	1	01	102	RCL5	36 05
019	X^2	53	047	RCL1	36 01	075	+	−55	103	RCL1	36 01
020	1	01	048	X=0?	16-43	076	$\sqrt{X}$	54	104	RTN	24
021	+	−55	049	GSB8	23 08	077	LOG	16 32	105	*LBL8	21 08
022	$\sqrt{X}$	54	050	RCL6	36 06	078	ST-5	35-45 05	106	1	01
023	RCL4	35 04	051	x	−35	079	RCL1	36 01	107	STO1	35 01
024	4	04	052	6	06	080	X=0?	16-43	108	RTN	24
025	x	−35	053	9	09	081	GSB7	23 07	109	*LBL7	21 07
026	x	−35	054	÷	−24	082	RCL6	36 06	110	1	01
027	LOG	16 32	055	RCL5	36 05	083	x	−35	111	STO1	35 01
028	6	06	056	+	−55	084	6	06	112	RTN	24
									113	R/S	51

table 24. Register contents for HP-67/97 program for calculating Z_0, h/d, and h/D for two parallel wires over a plane.

STO 0	ϵ_r
STO 1	$\sqrt{\epsilon_r}$
STO 2	D/h
STO 3	h/D
STO 4	h/d
STO 5	INTERIM
STO 6	Z_0

table 25. HP-67/97 program control for calculating Z_0, h/d, and h/D for two parallel wires over a plane.

calculates	$\sqrt{\epsilon_r}$	
enter	ϵ_r	press A
calculates	Z_0	
enter	d	press ENTER
enter	D	press ENTER
enter	h	press ENTER
		press B
calculates	h/D	
enter	Z_0	press ENTER
enter	h/d	press C
calculates	h/d	
enter	Z_0	press ENTER
enter	h/D	press D
calculates	Z_0	
enter	h/d	press ENTER
enter	h/D	press E

Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.

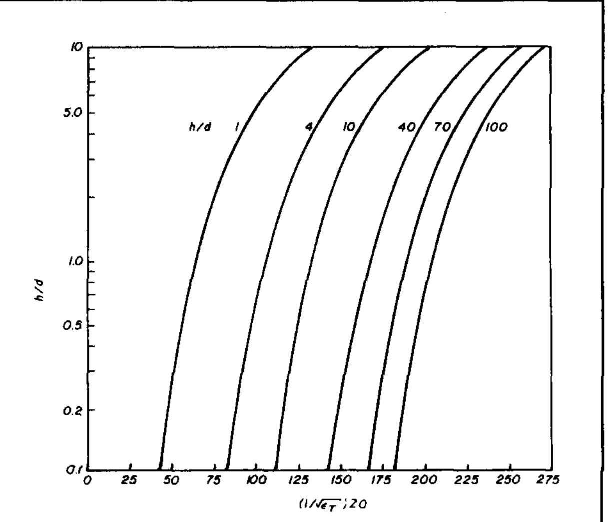
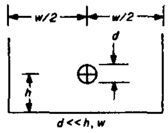


fig. 13. Z_0 **versus** h/D **for values of** h/d **for two parallel wires over a plane.**

circular wire in open trough



A common application of this configuration is in rf amplifiers, mixers, and local oscillator or multiplier filter elements at 100 MHz and above. The generalized formulation (reference 4) for a circular wire in an open trough is given by:

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \left[\frac{4\,w \tanh \frac{\pi h}{w}}{\pi d} \right] \qquad (29)$$

for $d << h,\ w$

where Z_0 = line impedance (ohms)
ϵ_r = dielectric constant
d = wire diameter
h = centerline height of wire over plane
w = width of trough with wire positioned in center

The formulation can be simplified considerably if one assumes a square open trough where w (trough width) equals $2h$ (two times the centerline height of the wire over the bottom plane). The resulting formula for a square open trough is:

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \left[\frac{8h}{\pi d} \tanh \frac{\pi}{2} \right] \qquad (30)$$

and

$$\tanh \frac{\pi}{2} = 1.0 \qquad (31)$$

then

$$Z_0 = \frac{138}{\sqrt{\epsilon_r}} \log_{10} \left[2.5465 \frac{h}{d} \right] \qquad (32)$$

Fig. 14 shows h/d versus Z_0 for **eq. 32. Table 26** is the HP-67/97 program for calculating Z_0 and h/d. **Table 27** shows the registers used. **Table 28** shows how the program is controlled.

Table 29 is the HP-67/97 program for calculating Z_0, h/w, and w/d given any two of the unknowns for any rectangular configuration, which is the general solution of **eq. 29. Table 30** indicates the storage registers used in the program, and **table 31** shows how the program is controlled. **Fig. 15** is a plot of Z_0 versus w/d for various values of h/w from **eq. 29.** Note that **eq. 29** is solved directly for Z_0 in **table 29.** If Z_0 and w/d are given, h/w is solved by using **eqs. 33, 34,** and **35.** A similar variant is used if h/w is known and w/d must be calculated.

$$\frac{Z_0 \sqrt{\epsilon_r}}{138} = \left(\log_{10} \frac{4w}{d} \right) + \left(\log_{10} \tanh \frac{\pi h}{w} \right), \qquad (33)$$

recognizing that

$$\tanh x = \frac{e^x - e^{-x}}{e^x + e^{-x}} \qquad (34)$$

and

$$\tanh^{-1} x = 1/2\, \ell n \left[\frac{1+x}{1-x} \right] \qquad (35)$$

table 26. HP-67/97 program for calculating Z_0 **and** h/d **for a circular wire in a square trough.**

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	018	1	01
002	STO0	35 00	019	3	03
003	√X	54	020	8	08
004	STO1	35 01	021	x	− 35
005	RTN	24	022	STO4	35 04
006	*LBLB	21 12	023	*LBL2	21 02
007	÷	− 24	024	RCL1	36 01
008	STO2	35 02	025	X = 0?	16-43
009	*LBL1	21 01	026	GSB9	23 09
010	2	02	027	÷	− 24
011	.	− 62	028	STO3	35 03
012	5	05	029	R/S	51
013	4	04	030	*LBLC	21 13
014	6	06	031	STO3	35 03
015	5	05	032	1	01
016	x	− 35	033	3	03
017	LOG	16 32	034	8	08

table 26. HP-67/97 program for calculating Z_0 and h/d for a circular wire in a square trough (continued).

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
035	÷	−24	051	R/S	51
036	STO4	35 04	052	*LBLD	21 14
037	*LBL3	21 03	053	STO2	35 02
038	RCL1	36 01	054	GTO1	22 01
039	X=0?	16-43	055	*LBL9	21 09
040	GSB8	23 08	056	1	01
041	x	−35	057	STO1	35 01
042	10^x	16 33	058	RCL4	36 04
043	2	02	059	GTO2	22 02
044	.	−62	060	*LBL8	21 08
045	5	05	061	1	01
046	4	04	062	STO1	35 01
047	6	06	063	RCL4	36 04
048	5	05	064	GTO3	22 03
049	÷	−24	065	R/S	51
050	STO2	35 02			

table 27. Register contents for HP-67/97 program for calculating Z_0 and h/d for a circular wire in a square trough.

STO 0	ϵ_r
STO 1	$\sqrt{\epsilon_r}$
STO 2	h/d
STO 3	Z_0
STO 4	INTERIM

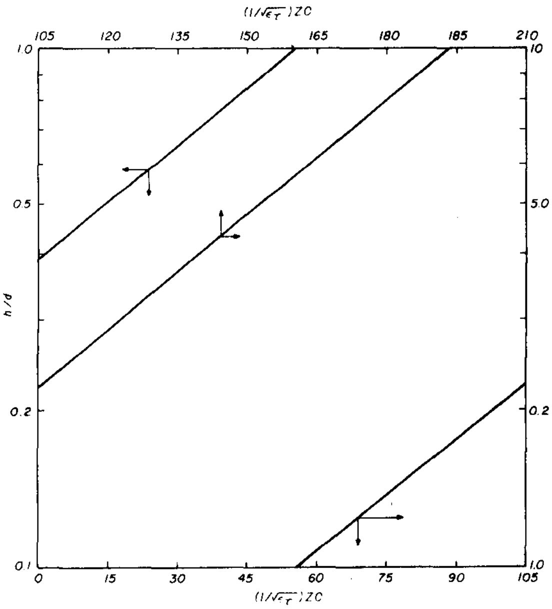


fig. 14. Z_0 **and** h/d **for a circular wire in a square open trough.**

table 28. HP-67/97 program control for calculating Z_0 and h/d for a circular wire in a square trough.

enter	ϵ_r	press A	
calculates	Z_0		
enter	h		
enter	d	press B	Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.
calculates	h/d		
enter	Z_0	press C	
calculates	Z_0		
enter	h/d	press D	

table 30. Register contents for HP-67/97 program for calculating Z_0, h/w, and w/d for a circular wire in a rectangular trough.

STO 0	ϵ_r	STO 6	$e^x + e^{-x}$
STO 1	$\sqrt{\epsilon_r}$	STO 7	$tanh\ \frac{\pi h}{w}$
STO 2	h/w		
STO 3	w/d	STO 8	INTERIM
STO 4	$\frac{\pi h}{w} = x$	STO 9	Z_0
		STO A	INTERIM
STO 5	$e^x - e^{-x}$	STO B	INTERIM

table 31. HP-67/97 program control for calculating Z_0, h/w, and w/d for a circular wire in a rectangular trough.

enter	ϵ_r	press A	
calculates	Z_0		
enter	d	press ENTER	
enter	h	press ENTER	
enter	w	press ENTER	
		press B	
calculates	h/w		Note: If no value for ϵ_r is entered, program assumes $\epsilon_r = 1 = air$.
enter	Z_0	press ENTER	
enter	w/d	press C	
calculates	w/d		
enter	Z_0	press ENTER	
enter	h/w	press D	

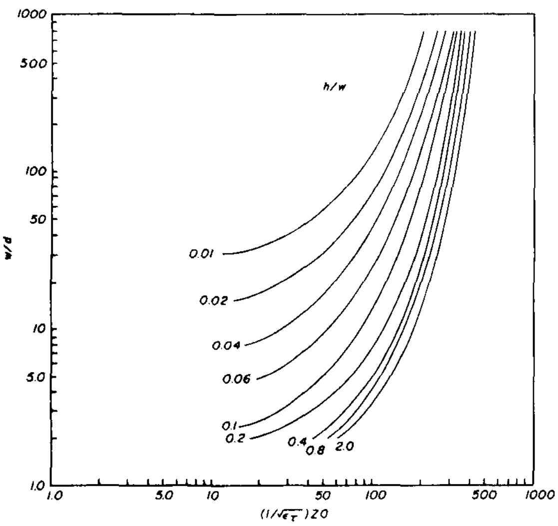
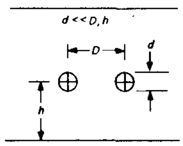


fig. 15. Z_0 **versus** w/d **for various values of** h/w **of a circular wire in an open rectangular trough.**

table 29. HP-67/97 program for calculating Z_0, h/w, and w/d for a circular wire in a rectangular trough.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	039	Pi	16-24	077	÷	–24	115	STO4	35 04
002	STO0	35 00	040	÷	–24	078	Pi	16-24	116	e^x	33
003	$\sqrt{X}$	54	041	LOG	16 32	079	x	–35	117	STO5	35 05
004	STO1	35 01	042	1	01	080	STOA	35 11	118	RCL4	36 04
005	RTN	24	043	3	03	081	1	01	119	CHS	–22
006	*LBLB	21 12	044	8	08	082	+	–55	120	e^x	33
007	R↓	–31	045	x	–35	083	STOB	35 12	121	ST+5	35-55 05
008	÷	–24	046	STO8	35 08	084	RCLA	36 11	122	RCL4	36 04
009	STO2	35 02	047	*LBL1	21 01	085	CHS	–22	123	e^x	33
010	R↓	–31	048	RCL1	36 01	086	1	01	124	STO6	35 06
011	÷	–24	049	X=0?	16-43	087	+	–55	125	RCL4	36 04
012	STO3	35 03	050	GSB9	23 09	088	1/X	52	126	CHS	–22
013	*LBL1	21 01	051	x	–35	089	RCLB	36 12	127	e^x	33
014	RCL2	36 02	052	STO9	35 09	090	x	–35	128	ST-6	35-45 06
015	Pi	16-24	053	R/S	51	091	LN	32	129	RCL6	36 06
016	x	–35	054	*LBL9	21 09	092	2	02	130	RCL5	36 05
017	STO4	35 04	055	1	01	093	÷	–24	131	÷	–24
018	e^x	33	056	STO1	35 01	094	Pi	16-24	132	LOG	16 32
019	STO5	35 05	057	RCL8	36 08	095	÷	–24	133	CHS	–22
020	RCL4	36 04	058	RTN	24	096	STO2	35 02	134	ST+8	35-55 08
021	CHS	–22	059	*LBLC	21 13	097	R/S	51	135	RCL8	36 08
022	e^x	33	060	STO3	35 03	098	*LBLD	21 14	136	10^x	16 33
023	ST+5	35-55 05	061	R↓	–31	099	STO2	35 02	137	4	04
024	RCL4	36 04	062	STO9	35 09	100	R↓	–31	138	÷	–24
025	e^x	33	063	1	01	101	STO9	35 09	139	Pi	16-24
026	STO6	35 06	064	3	03	102	1	01	140	x	–35
027	RCL4	36 04	065	8	08	103	3	03	141	STO3	35 03
028	CHS	–22	066	÷	–24	104	8	08	142	R/S	51
029	e^x	33	067	STO8	35 08	105	÷	–24	143	*LBL8	21 08
030	ST-6	35-45 06	068	RCL1	36 01	106	STO8	35 08	144	1	01
031	RCL6	36 06	069	X=0?	16-43	107	RCL1	36 01	145	STO1	35 01
032	RCL5	36 05	070	GSB9	23 09	108	X=0?	16-43	146	RCL8	36 08
033	÷	–24	071	x	–35	109	GSB8	23 08	147	RCL1	36 01
034	STO7	35 07	072	10^x	16 33	110	x	–35	148	RTN	24
035	RCL3	36 03	073	RCL3	36 03	111	STO8	35 08	149	R/S	51
036	x	–35	074	1/X	52	112	RCL2	36 02	150	*LBLE	21 15
037	4	04	075	x	–35	113	Pi	16-24	151	STO2	35 02
038	x	–35	076	4	04	114	x	–35	152	R↓	–31
									153	STO3	35 03
									154	GTO1	22 01
									155	R/S	51

parallel wires between planes/ rectangular box



This configuration is often used for high-powered, push-pull amplifiers. Even though the formulation in **eq. 36** does not consider the effects of side walls (only top and bottom planes), good results can be obtained if the lines are centered between the side walls and the distance from either wire to the side wall is at least greater than $h/2$, as shown in the sketch. A more exact formulation for balanced two-wire lines in a rectangular enclosure is given in reference 4 and discussed later in this section; however, it requires a rather tedious series of calculations.

The following formulation, also from reference 4, is based upon this relationship:

$$Z_0 = \frac{276}{\sqrt{\epsilon_r}} \log_{10} \left[\left(\frac{4h}{\pi d} \right) \left(\tanh \frac{\pi D}{2h} \right) \right] \quad \textbf{(36)}$$

For those who wish the exact relationship, it is given below from reference 4; the geometry is in **fig. 16**.

$$Z_0 = \frac{276}{\sqrt{\epsilon_r}} \left\{ \left[log_{10} \left(\frac{4h}{\pi d} \right) \left(tanh \; \frac{\pi D}{2h} \right) \right] - \sum_{m=1}^{\infty} log_{10} \left(\frac{1+U_m^2}{1-V_m^2} \right) \right\} \tag{37}$$

where Z_0 = line impedance (ohms)
ϵ_r = dielectric constant
h = spacing between planes with the lines centered
D = centerline distance between the lines
d = diameter of the lines (lines are assumed to be of equal diameter)

and

$$U_m = \frac{sinh\left(\frac{\pi D}{2h}\right)}{cosh\left(\frac{m\pi w}{2h}\right)} \tag{38}$$

$$V_m = \frac{sinh\left(\frac{\pi D}{2h}\right)}{sinh\left(\frac{m\pi w}{2h}\right)} \tag{39}$$

table 32. HP-67/97 program for calculating Z_0, h/d, and D/h for two parallel wires between planes/rectangular box.

step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code	step	HP-97 key	HP-97 code
001	*LBLA	21 11	039	R/S	51	077	÷	−24	115	RCL5	36 05
002	STO0	35 00	040	*LBLC	21 13	078	STO6	35 06	116	CHS	−22
003	√X	54	041	STO2	35 02	079	RCL1	36 01	117	e^x	33
004	STO1	35 01	042	R↓	−31	080	X=0?	16-43	118	−	−45
005	RTN	24	043	STO4	35 04	081	GSB8	23 08	119	RCL8	36 08
006	*LBLB	21 12	044	2	02	082	x	−35	120	÷	−24
007	R↓	−31	045	7	07	083	STO6	35 06	121	RTN	24
008	÷	−24	046	6	06	084	RCL3	36 03	122	*LBL8	21 08
009	STO2	35 02	047	÷	−24	085	4	04	123	1	01
010	R↓	−31	048	STO6	35 06	086	x	−35	124	STO1	35 01
011	÷	−24	049	RCL1	36 01	087	Pi	16-24	125	RCL6	36 06
012	STO3	35 03	050	X=0?	16-43	088	÷	−24	126	RCL1	36 01
013	*LBL1	21 01	051	GSB8	23 08	089	LOG	16 32	127	RTN	24
014	4	04	052	x	−35	090	ST-6	35-45 06	128	*LBL7	21 07
015	x	−35	053	STO6	35 06	091	RCL6	36 06	129	1	01
016	Pi	16-24	054	RCL2	36 02	092	X<0?	16-45	130	+	−55
017	÷	−24	055	Pi	16-24	093	GSB6	23 06	131	1	01
018	LOG	16 32	056	x	−35	094	10^x	16 33	132	RCL7	36 07
019	STO6	35 06	057	2	02	095	*LBL5	21 05	133	−	−45
020	RCL2	36 02	058	÷	−24	096	STO7	35 07	134	÷	−24
021	Pi	16-24	059	GSB9	23 09	097	GSB7	23 07	135	LN	32
022	x	−35	060	LOG	16 32	098	STO2	35 02	136	2	02
023	2	02	061	ST-6	35-45 06	099	R/S	51	137	÷	−24
024	÷	−24	062	RCL6	36 06	100	*LBLE	21 15	138	2	02
025	GSB9	23 09	063	10^x	16 33	101	STO2	35 02	139	x	−35
026	LOG	16 32	064	Pi	16-24	102	R↓	−31	140	Pi	16-24
027	ST+6	35-55 06	065	x	−35	103	STO3	35 03	141	÷	−24
028	RCL6	36 06	066	4	04	104	GTO1	22 01	142	STO2	35 02
029	2	02	067	÷	−24	105	*LBL9	21 09	143	RTN	24
030	7	07	068	STO3	35 03	106	STO5	35 05	144	*LBL6	21 06
031	6	06	069	R/S	51	107	e^x	33	145	CHS	−22
032	x	−35	070	*LBLD	21 14	108	RCL5	36 05	146	10^x	16 33
033	STO6	35 06	071	STO3	35 03	109	CHS	−22	147	1/X	52
034	RCL1	36 01	072	R↓	−31	110	e^x	33	148	GTO5	22 05
035	X=0?	16-43	073	STO4	35 04	111	+	−55	149	R/S	51
036	GSB8	23 08	074	2	02	112	STO8	35 08			
037	÷	−24	075	7	07	113	RCL5	36 05			
038	STO4	35 04	076	6	06	114	e^x	33			

Note that the open-sided equation (**eq. 36**) represents the maximum value of Z_0 that can be achieved with no side enclosure. The sum of the series in **eq. 37** reduces this value of Z_0 by the closing-in effect of the sides. As a consequence, if **eq. 37** and the charts and programs provided here are used, a conservative design will result, permitting the addition of more capacitance than the values calculated, as described in a previous section.

Fig. 17 shows the value of h/d plotted versus D/h for various values of Z_0. **Table 32** is the HP-67/97 program for calculating Z_0, D/h, or h/d depending on which variables are given. **Table 33** shows the registers used in the program, and **table 34** describes how the program is controlled.

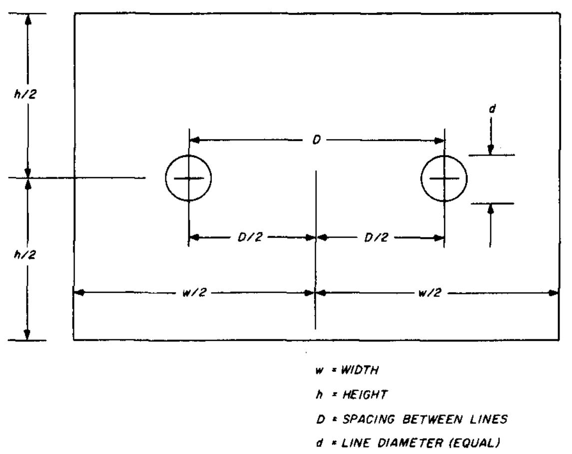


fig. 16. Geometry of balanced parallel lines in rectangular enclosure.

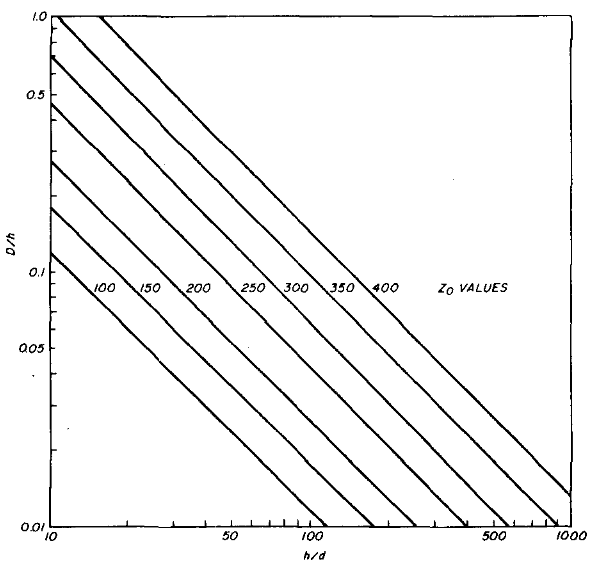


fig. 17. h/d versus D/h for various values of Z_0 for two parallel wires between planes.

A more simplified formulation for **eq. 36** is used in the program to permit calculation of the desired unknowns:

$$\frac{Z_0\sqrt{\epsilon_r}}{276} = \left(log_{10}\frac{4h}{\pi d}\right) + log_{10}\left(tanh\,\frac{\pi D}{2h}\right) \qquad \mathbf{(40)}$$

table 33. Register contents for HP-67/97 program for calculating Z_0, h/d, and D/h for two parallel wires between planes/rectangular box.

STO 0	ϵ_r	STO 5	x for $tanh\,x$
STO 1	$\sqrt{\epsilon_r}$	STO 6	INTERIM
STO 2	D/h	STO 7	$tan^{-1}x$
STO 3	h/d	STO 8	INTERIM
STO 4	Z_0		

table 34. HP-67/97 program control for calculating Z_0, h/d, and D/h for two parallel wires between planes/rectangular box.

calculates	ϵ_r		
enter	ϵ_r	press A	
calculates	Z_0		
enter	d	press ENTER	
enter	D	press ENTER	
enter	h	press ENTER	
		press B	
calculates	h/d		
enter	Z_0	press ENTER	
enter	D/h	press C	Note: If no value for
calculates	D/h		ϵ_r is entered, program
enter	Z_0	press ENTER	assumes $\epsilon_r = 1 = air$.
enter	h/d	press D	
calculates	Z_0		
enter	h/d	press ENTER	
enter	d/h	press E	

In the next part of this article, part 4, the geometry and resonant-circuit design of the following configurations will be discussed: circular wire in a square shield, stripline over a plane, stripline centered between parallel planes, and helical resonators.

references

4. *Reference Data for Radio Engineers*, ITT Corporation, 5th Edition, 1968 (included in subsequent editions).

bibliography

Bahl, Dr. K.J., "Use Exact Methods for Microstrip Design," *Microwaves*, December, 1978, pages 61-62.

Gardiol, F.E., "HP-65 Program computes Microstrip Impedance," *Microwaves*, December, 1977, pages 186-187.

"Letters to the Editor," *Microwave Systems News*, December, 1978, pages 13-14.

Murdock, B.K., *Handbook of Electronics Design and Analysis Procedures Using Programmable Calculators*, Van Nostrand Reinhold, 1979.

ham radio

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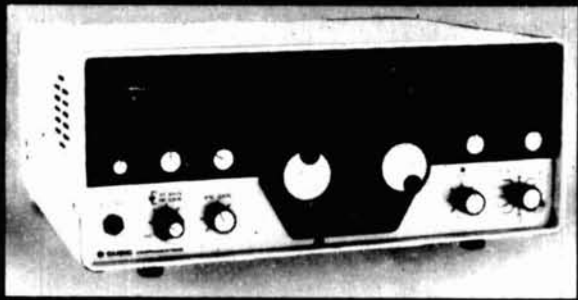

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Being only two blocks from W1AW is a horrifying experience for an amateur, since simultaneous 1-KW operation takes place almost around the clock on 80 through 10 meters, to say nothing of operation on 160, 6 and 2 meters! Very few receivers can withstand the rf onslaught, and tend to crumble when W1AW is operating. Some have survived when front-end attenuation switched in, but none could handle the W1AW. The 102 BXA not only survives without an attenuator, but can function normally when my beam is headed toward W1AW! Your rig is probably the cause of my finally changing to U.S. made equipment.

Doug DeMaw, W1FB
Technical Department Manager
ARRL

This transceiver is the only one I have tested recently that has met all its published specifications or even went beyond. I am specifically impressed with the behavior of the PLL and the AGC. To the best of my knowledge, in this respect, you are better than all current production instruments including the Collins KWM380, which I had tested.

ULRICH L. ROHDE

SPECIFICATIONS

Notch Filter:
Provides 30dB typical notch attenuation from .4 to 3.0 kHz

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16 pole crystal filtering 2.7 kHz at -6dB, 3.78 kHz at -100dB
Shape factor 1.4:1

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Sensitivity:
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Better than +15dBm

Image Rejection:
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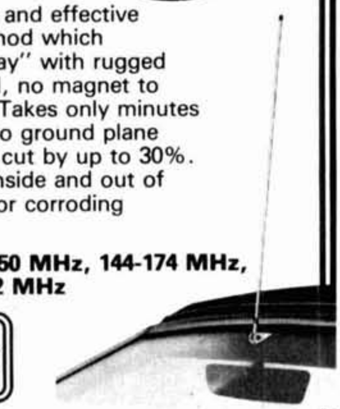

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the ham notebook

computer control for the KLM antenna rotor

The most obvious approach to microprocessor control of antenna rotors does indeed work! In **fig. 1** two CMOS quad-clocked D latches, U1, are connected in parallel to the first four least-significant digits of an input/output port of a computer. The computer enables first one, then the other latch, leaving behind an 8-bit word.

This byte, a number between 0-255, is converted through the classic "R-2R" network to a voltage between 0-5 volts. This voltage is compared with the voltage returning from the wiper of the potentiometer in the rotor housing.

The comparators turn on solid-state switches, which are connected across the clockwise and counterclockwise front-panel switches on the rotor control box.

Thus the antenna will rotate in the direction that minimizes the difference between the two voltages.

The two 820k resistors provide a deadband so that when the proper heading is reached, the rotor will stop — not hunt.

Because the quad comparator, U2, loses all resemblance to a comparator as the inputs approach the upper rail, adjust the 15-turn pot, R, for a comparator input voltage of about 3.5 volts at full rotation. This voltage corresponds to a byte of 180. Then merely have the computer poke down to the latches a number between 0-180, which is one-half the required azimuth heading (or twice the required elevation).

The terminal numbers indicated in **fig. 1** are for the KLM KR-400 and KR-500 rotors. Terminal 8 must be wired to the common connection of the two front-panel direction controls. Solid-state switches can be homebuilt with an opto-triac driver and triac, as in **fig. 2**.

This interface is very precise — four-degree resolution or better at most headings. The circuit will readjust for sudden wind gusts, voltage variations, etc., without complex software. The computer needs only to enter direction data every minute, say, using only a few machine cycles then go on to other work.

I conclude with a report of negative results: my first attempt at such an interface was with the new low-cost RCA CA3162E, two 4-bit magnitude comparators and four latches. The plan was to convert the rotor heading to digital information, then compare. Aside from problems with the multiplex timing of the 3162, this was a wasteful approach. At least ten comparators were at work instead of two!

C.R. Mac Cluer, W8MQW

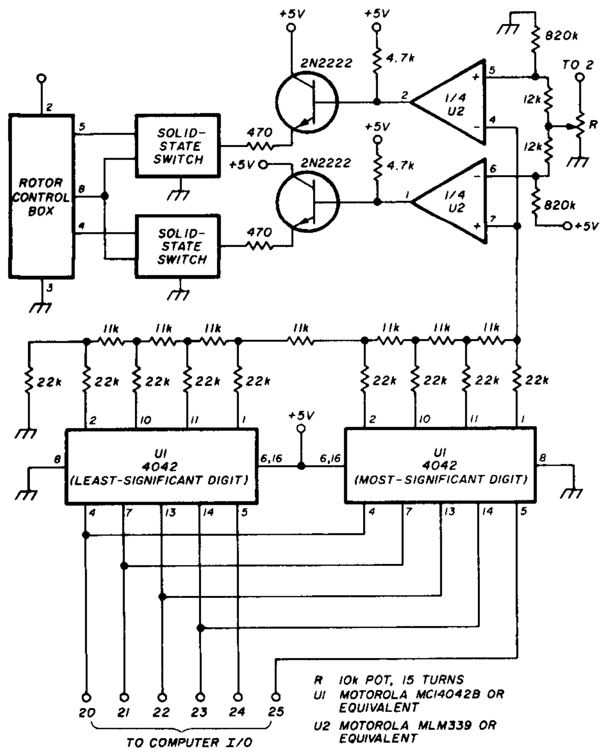
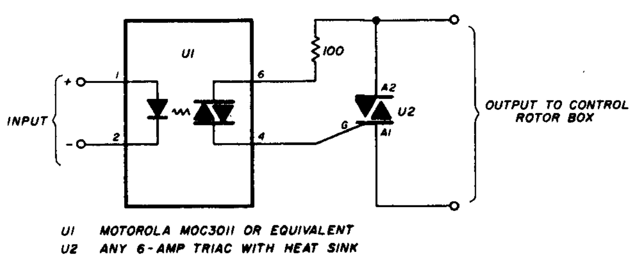


fig. 1, left: Interface circuit for microprocessor control of the KLM KR-400 and KR-500 antenna rotors.

fig. 2, below: Schematic of the solid-state switches, consisting of optically coupled triac driver and 6-ampere triac with heatsink.



varactor tuning tips

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

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

Richard N. Coan, N3GN

woodpecker noise blanker for the Drake TR-7 transceiver

If you have been reading the articles about the "Woodpecker Noise Blanker"[1] and wished your TR-7 could do the same to the "Russian Woodpecker," it can — with just two component changes. The Drake noise blanker for the TR-7 is functionally the same block diagram; the problem is that the one-shot and integrator time constants aren't set up for the Woodpecker.

I increased C831 from 0.001 μF to 0.01 μF and the one-shot capacitor from 0.01 μF to 0.1 μF. The results were fair to good. A 10-20 dB over S9 Woodpecker signal would be cut back to S5-6 with no real distortion to SSB, CW, or RTTY. I called Drake about the idea, but they have also made a change which will be in new NB7s to be sold soon. Their results were about the same as regards attenuation of the Woodpecker signal. Drake sent me a copy of their changes and I'll compare it with what I've done.

The noise blanker works best on strong Woodpecker signals (over S7). I have received three Woodpeckers each with its own signature. All have the same rate of 100 milliseconds (probably due to a division of the 50-Hz power line) but differ in pulse width and rise time.

The changes are simple and not critical. The only problem is that in my service manual, the pictorial didn't show the position of C840. **Fig. 3** shows the area of the schematic where the changes are made, and **fig. 4** shows the pictorial of the components to be changed.

I would like to thank W1IHN for allowing me to use his noise blanker to test this idea.

reference

1. Ulrich L. Rohde, DJ2LR, "Woodpecker Noise Blanker," *ham radio*, June, 1980, page 18.

John Bird, K1KSY

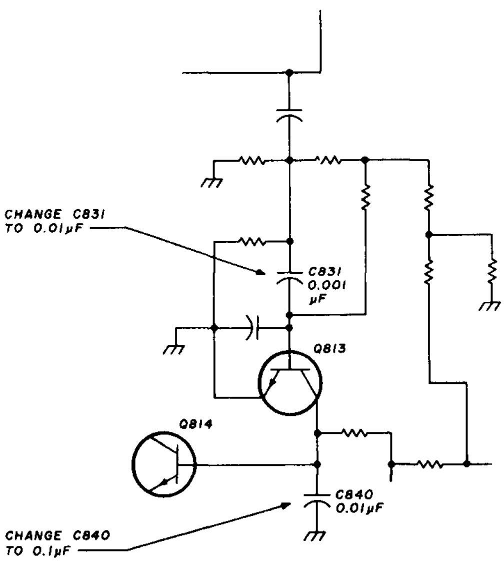


Fig. 3

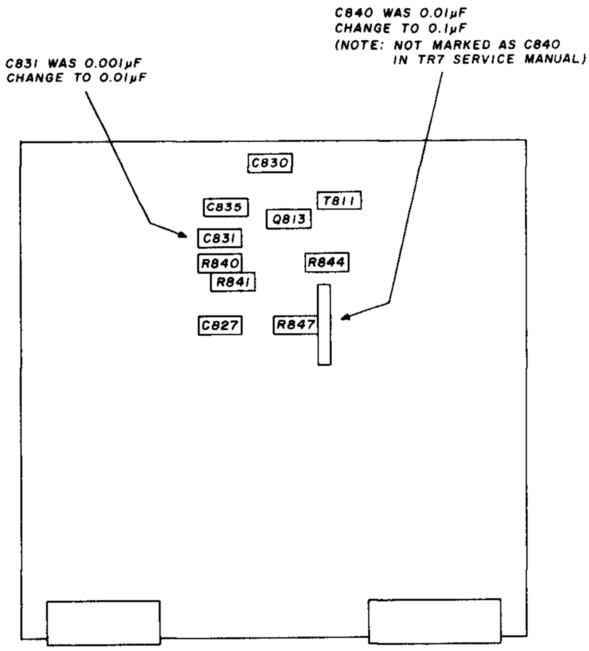


Fig. 4

More Details? CHECK—OFF Page 98

MHz electronics

MICROWAVE COMPONENTS

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2416	Variable Attenuator	$ 50.00
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H487B	100 ohms Neg Thermistor Mount (USED)	100.00
477B	200 ohms Neg Thermistor Mount (USED)	100.00
X487A	100 ohms Neg. Thermistor Mount (USED)	100.00
X487B	100 ohms Neg. Thermistor Mount (USED)	125.00
J468A	100 ohms Neg Thermistor Mount (USED)	150.00
478A	200 ohms Neg Thermistor Mount (USED)	150.00
J382	5.85 to 8.2 GHz Variable Attenuator 0 to 50dB	250.00
X382A	8.2 to 12.4 GHz Variable Attenuator 0 to 50dB	250.00
394A	1 to 2 GHz Variable Attenuator 6 to 120dB	250.00
MK292A	Waveguide Adapter	65.00
K422A	18 to 26.5 GHz Crystal Detector	250.00
8436A	Bandpass Filter 8 to 12.4 GHz	75.00
8439A	2 GHz Notch Filter	75.00
8471A	RF Detector	50.00
H532A	7.05 to 10 GHz Frequency Meter	300.00
G532A	3.95 to 5.85 GHz Frequency Meter	300.00
J532A	5.85 to 8.2 GHz Frequency Meter	300.00
809A	Carriage with a 444A Slotted Line Untuned Detector Probe and 809B Coaxial Slotted Section 2.6 to 18 GHz	175.00

Merrimac

AU-25A/	801115 Variable Attenuator	100.00
AU-26A/	801162 Variable Attenuator	100.00

Microlab/FXR

X638S	Horn 8.2 - 12.4 GHz	60.00
601-B18	X to N Adapter 8.2 - 12.4 GHz	35.00
Y610D	Coupler	75.00

Narda

4013C-10/	22540A Directional Coupler 2 to 4 GHz 10db Type SMA	90.00
4014-10/	22538 Directional Coupler 3.85 to 8 GHz 10dB Type SMA	90.00
4014C-6/	22876 Directional Coupler 3.85 to 8 GHz 6dB Type SMA	90.00
4015C-10/	22539 Directional Coupler 7.4 to 12 GHz 10dB Type SMA	95.00
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3044-20	Directional Coupler 4 to 8 GHz 20dB Type N	125.00
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PRD

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C101	Variable Attenuator 0 to 60dB	200.00
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195B	8.2 to 12.4 GHz Variable Attenuator 0 to 50dB	100.00
185BS1	7.05 to 10 GHz Variable Attenuator 0 to 40dB	100.00
196C	8.2 to 12.4 GHz Variable Attenuator 0 to 45dB	100.00
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588A	Frequency Meter 5.3 to 6.7 GHz	100.00
140A,C,D,E	Fixed Attenuators	25.00
109J,I	Fixed Attenuators	25.00
WEINSCHEL ENG.	2692 Variable Attenuator +30 to 60dB	100.00

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2716/2516	2K x 8 EPROM 5Volt Single Supply	20.00
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2114L2	1K x 4 Static RAM 250ns	8.99
2114L3	1K x 4 Static RAM 350ns	7.99
4027	4K x 1 Dynamic RAM	3.99
4060/2107	4K x 1 Dynamic RAM	3.99
4050/9050	4K x 1 Dynamic RAM	3.99
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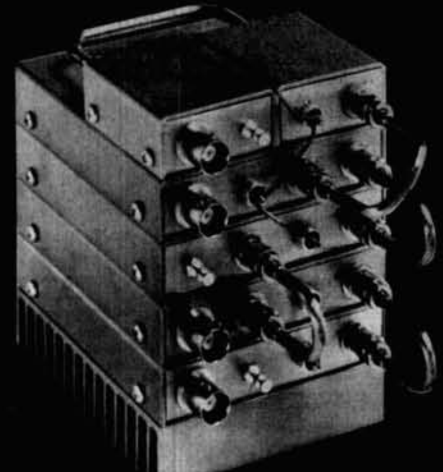

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HAM RADIO REPAIR — Professional lab, personal service. "Grid" Gridley, W4GJO. April thru October: Rt. 2, Box 138B, Rising Fawn, Georgia 30738. (404) 657-7841. November thru March: 212 Martin Drive, Brooksville, Florida 33512. (904) 799-2769.

MICROWAVE TV downconverters, preamps, parabolic dish antennas, remote tuning. Covers 2000 MHz band. Write for information. LAB-TRONICS, INC., Box 171, Rogers, MN 55374.

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WANTED: Motorola micor base stations. 406-420 MHz. AK7B, 4 Ajax Pl., Berkeley, CA 94708.

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Coming Events
ACTIVITIES

COLORADO: The Grand Mesa Repeater Society's second annual Western Slope Swapfest on March 7th at the Lincoln Park Barn, 12th and Gunnison, Grand Junction, Colorado. 10:00 AM through 4:00 PM FREE! Tables are $4.00 in advance. Commercial exhibitors, flea market, auction and prizes. Talk-in on 146.22/.82. More info: SASE to Larry Brooks, WB0ECV, 3185 Bunting Ave., Grand Junction, Colorado 81501. (303) 434-5603.

FLORIDA: The Treasure Coast Hamfest on February 21 and 22 at the Vero Beach Community Center. Admission: $3.00 per family in advance and $4.00 at the door. Talk-in on 146.13/73, 146.04/64, 146.52/52, 222.34/223.94. More info: P.O. Box 3088, Beach Station, Vero Beach, FL 32960.

ILLINOIS: Sterling/Rock Falls Amateur Radio Society's 21st annual Hamfest on March 8 at the Sterling High School Field House, 1608 4th Ave., Sterling, Illinois. Flea market, free parking, many prizes, and much more. Doors open at 7:30 AM. For advance tickets and tables write: Sue Peters, 511-8th Ave., Sterling, Illinois 61081. Tickets in advance: $2.00. At the door: $2.50. Commercial tables are $5.00 and all others are $3.00. Talk-in on .52 and WR9AER .25/85.

OHIO: Intercity Amateur Radio Club's Mansfield Mid • Winter Hamfest/Auction on February 15th at the Richland County Fairgrounds in Mansfield. Prizes, flea market, large heated building and more. Starts at 8:00 AM. Talk-in on 146.34/.94. Tickets: $1.50 in advance and $2.00 at the door. More info, tickets, and/or tables: SASE, Harry Frietchen, 120 Homewood Rd., Mansfield, Ohio 44906. (419) 529-2801.

VIRGINIA: The Vienna Wireless Society, Vienna, Virginia, announces the annual WINTERFEST™ hamfest to take place this year on February 22, 1981. It will be held at the usual place, the Vienna Community Center on Park St. in Vienna, VA beginning at 8 AM.

WEST VIRGINIA: The Plateau Amateur Radio Association's 3rd annual Hamfest on February 15th at the Memorial Building in Fayetteville. Doors open at 9:00 AM. Admission: $2.50 (children free) and flea market tables are $2.00. Hot food, free parking, A.R.R.L. display, forums, exhibits, door prizes, XYL programs, and more. All activities indoors. Talk-in on .52 or .19/79. More info: Bill Wilson, 302 Central Ave., Apt. #2, Oak Hill, WV 25901. (304) 469-9910 or 574-1176.

IOWA: The Davenport Radio Amateur Club's tenth annual Hamfest on March 1st at the Davenport Masonic Temple at 7th and Brady (Hwy. 61) from 8:00 AM to 4:00 PM. Over $2000.00 in major prizes. Tickets in advance are $2.00 and $3.00 at the door. Talk-in on the WØBXR/RPT at 146.28/.88. Hotel discounts available. Food and drink. Pre-Hamfest Saturday night banquet with guest speaker Paul Grauer, midwest A.R.R.L., S.C.M. Banquet tickets: $8.00. Paid reservations for banquet by February 18th. For advance tickets, dinner, table reservations or info: Dave Johannsen, 2131 Myrtle, Davenport, Iowa 52804.

MICHIGAN: The 11th annual Livonia Amateur Radio Club's Swap 'n Shop on February 22 from 8:00 AM to 4:00 PM at the Churchill High School in Livonia. Plenty of tables, door prizes, refreshments and free parking. Talk-in on 146.52 simplex. Reserved table space available. More info: send SASE (4x9) to Neil Coffin, c/o Livonia A.R.C., P.O. Box 2111, Livonia, Michigan 48150.

MISSOURI: The Jefferson Barracks Amateur Radio Club's annual auction and Hamfest on March 13th at the Electricians Hall, 5850 Elizabeth Ave., St. Louis. More info: SASE Vivian K. Scott, WDØEMS, 4121 Fabian Dr., St. Louis, Missouri 63125.

NEW JERSEY: The Delaware Valley Radio Association's ninth annual flea market on March 15th at the New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Rd., Lawrence Township. Advance tickets: $2.00, $2.50 at the gate. Indoor/outdoor flea market, prizes, refreshments and more. Talk-in on 146.07/67 and 146.53. More info or tickets: DVRA, P.O. Box 7024, West Trenton, NJ 08628. SASE please.

NEW JERSEY: Annual Flemington, N.J. Hamfest Saturday, March 21 from 8:30 to 3:00 at the Hunterdon Central High School Field House. 20,000 square feet of heated indoor area. Gigantic flea market, 200 tables, major manufacturers, informative seminars. Bring the XYL, kids and friends. Flemington is a tourist area. Talk in 146.52, 147.375, 147.015, 224.12. Admission $3.00 donation. For reservations or info call 201-788-4080 or write Cherryville Repeater Assn. c/o W2FCW, Box 76, Farview Ave., Annandale, N.J. 08801.

INDIANA: The LaPorte ARC's "Winter Hamfest" on February 22 at the LaPorte Civic Auditorium, 50 miles southeast of Chicago. Donations: $2.50 at gate and $2.00 in advance. Talk-in on .01/.61 and .52 simplex. More info: LPARC, Box 30, LaPorte, Indiana 46350.

OPERATING EVENTS

FEBRUARY 14 and 15 starts off the phone portion of the YL-OM contest. The CW segment will be on February 28 through March 1. Both contests start and end at 1800 UTC. They are to be treated separately and separate logs are to be kept. OMs call "CQ YL" and YLs call "CQ OM". All logs must show ARRL section or country to qualify and must be signed by the operator. No logs will be returned. Logs must show claimed score and be sent to YLRL Vice President, Kay Eyman, WAØWOF, RR #2, Garnett, KS 66032 no later than April 6. Duplicates penalized by the removal of three contacts.

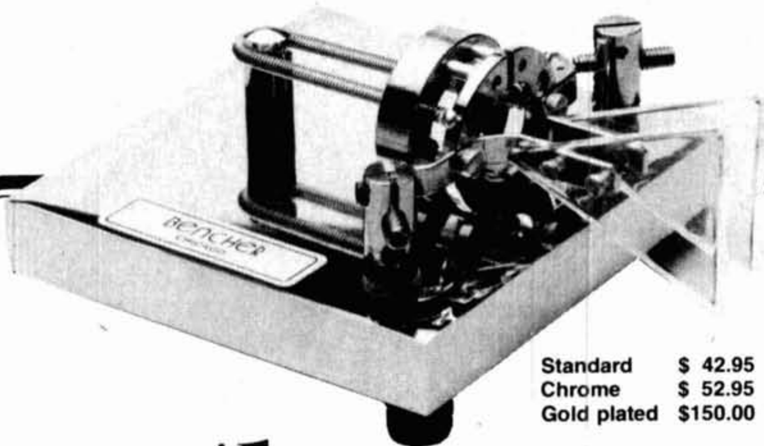
OCTOBER 23, 1980 - OCTOBER 23, 1981: The New Bern Amateur Radio Club is sponsoring "The Swiss Bear Award." This award is given for working three different amateur stations in the New Bern area within the above time period. Extracts from logs for QSO's, along with a SASE or two IRC's (for DX stations) should be sent to: New Bern Amateur Radio Club, Inc., P.O. Box 2483, New Bern, NC 28560. Certificate depicting "The Swiss Bear" will be awarded stating that the station has met the requirements for this award.

FEBRUARY 22nd - 28th: The Simon Langton Grammar Schools in Canterbury, England, will be celebrating the 100th anniversary of the founding of the school in 1881. A special events station, active on all HF bands under the call GB4SLS, will be used. It is hoped that many past pupils of the school will be contacted; especially licensed Amateurs now residing in the United States. Anyone interested in making a sked should contact: G4BBW, 40 Virginia Rd., Tankerton, Whitstable, Kent or G3LCK c/o G3OSL, Simon Langton Grammar School for Boys, Nackington Rd., Canterbury, Kent, England.

MARCH 14th: The Edison Radio Amateur's Association is celebrating its 40th anniversary by hosting a QSO party. The ERAA group will operate from a commemorative station at "Station A" in Greenfield Village, Dearborn, Michigan. This was Thomas Edison's first power generating station. Contact ERAA and exchange signal report state info. QSL with business size SASE and receive a handsome two color certificate to: Detroit Edison Radio Amateur's Association, 2000 Second Ave., Detroit, Michigan 48226.

MARCH 14th: The Playground Amateur Radio Club (PARC) of Fort Walton Beach, Florida, will operate a special event station at the 1981 Boy Scouts of America — Choctawhatchee District Scout Exhibition. PARC members will operate ARS WB4SFU (Scouts For Unity) from 0000 to 2400 hrs UTC, 14 March, 1981 on 14.290 MHz, 21.370 MHz, and 28.600 MHz SSB. A special commemorative QSL card will be sent to those who QSL with a SASE. The QSL manager is PARC, c/o Joe Giangrosso, WD4JZG, P.O. Box 3075, Fort Walton Beach, FL 32548.

CENTURY 21 ARC — Low power — QRP'ers — CW nets — Contests — Awards — SASE KA4EBW.



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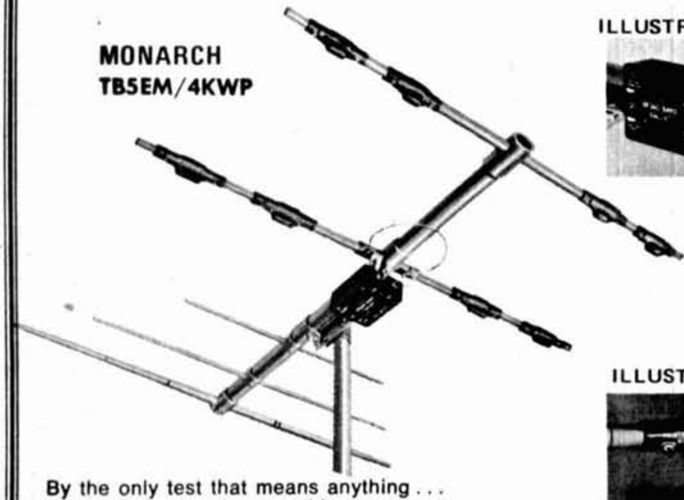

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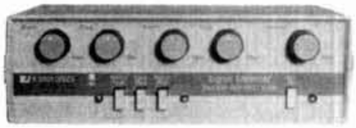

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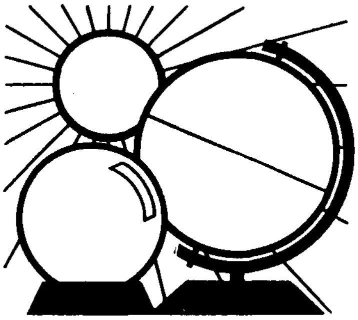

DX FORECASTER

Garth Stonehocker, KØRYW

predictions

We are still in the good old winter DX months with February; expect, however, a couple of periods of short-lived geomagnetic disturbances from solar flares about February 19-21 and February 27-March 4. Another longer-lasting period of disturbance may be expected near February 2-6. The best DX periods are probably about February 7-10 and 22-27.

a look at February

February is the month when changes in the ionosphere portend leaving the winter DX conditions of November, December, and January behind. Longer days are beginning to be seen as earlier sunrises and later sunsets in the northern hemisphere. The 10, 15, and 20 meter bands can be expected to open sooner in the morning and stay open longer into the evening. On the 40, 80, and 160 meter bands, which depend on darkness for their openings, the DX hours can be expected to begin to shrink. The DXer's evening schedule has to wait for sunset. The earlier sunrise decreases the S-meter readings of the long-skip DX, and exchanges it for short skip to the locals. Still, there are many hours in which to take advantage of the superb winter DX.

February is often the month with the highest mean solar radio flux values of the year. This is a long-lasting, averaging effect of the earth's being closer to the sun this time of year, and the winter months of November, December, and January being geomagnetically the quietest of the year. The combined result is that the ionosphere usually supports the highest daytime high-frequency and vhf DX paths of the year for many of the years of any particular eleven-year solar cycle.

The moon is at perigee on February 9. An annular eclipse of the sun takes place for our down-under and South-Pacific friends. The annular path (maximum duration is 1 minute 11 seconds) goes from the Island of Tasmania south of Australia at sunrise (1928 UT), then south of New Zealand, and across the South Pacific almost to Peru at sunset (0049 UT). The eclipse will be seen across the east side of Australia, across Antarctica (McMurdo from 2110 to 2220 UT), and from Chile, Argentina, Peru and Central America at its end. All this goes on from February 4 (1928 UT) to 5 (0049 UT).

band-by-band summary

Six meters will open occasionally toward Europe, that is, to the east, before noon, toward the south during noontime to afternoon, and toward the west and northwest in the late afternoon into early evening. The best openings are most likely trans-equatorial during high solar radio flux.

Ten and Fifteen meters will exhibit the same pattern as 6 meters but will be open a longer part of the day. This is particularly true this month since it is nearing springtime with its noticeably longer days and its probably higher maximum usable frequencies from higher solar flux. Short skip (500 to 1500 miles) is part of the fun on these bands, like the lower frequency bands. The short-skip opening pattern, although closer to mid-day, is the same follow-the-sun sequence as mentioned for 6 meters.

Twenty meters is a great band for everyone's pleasure, limited only by QRM. It should be open nearly every day and late into each evening to almost every part of the world. Best DX conditions can be expected just after sunrise and just before sunset for long skip. Short skip will be essentially as given for 10 and 15 meters, except for longer openings during midday.

Forty meters begins its transition into a night band. Short skip during the daytime in winter, however, gives some interesting opportunities for working your close neighbors for the WAS certificate. Then, at evening time, as the long skip (1000 to 2500 miles) develops, reach out for the far states and the WAC certificate. This band is very active to most areas of the world. In late afternoon the band will open to Europe, then swing around to South Africa and Central and South America, and then swing still farther into the Pacific by dawn.

Eighty and One-Sixty meters DX conditions will be very good this month. Soon the atmospheric noise of the spring storms will give days of short skip QRN and local QRN. On toward summer the static will become so bad that DX will have to be forgotten until fall. Take advantage of what's left of this year's quiet winter season. The directional pattern for these bands is similar to that of 40 meters. The low take-off angle of vertical antennas is very useful for DX here. Horizontal antennas are mainly short skip, high-take-off-angle radiators because of being so close to the ground. Look for particularly interesting DX as these bands come in (open) near sunset and go out at sunrise.

ham radio

FEBRUARY

WESTERN USA

GMT	PST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000		10	—	15	10	15	10	10	10
0100		10	—	15	10	15	10	10	10
0200		10	—	15	10	15	10	10	10
0300		10	—	20	15	15	10	10	10
0400		15	40	20	15	15	10	10	15*
0500		15	40	20	20	15	15*	15	15
0600		20	40	20	20	15	15	15	20*
0700		20	40	20	20	20	15	15	20
0800		—	20	20	20	20	15	20	20
0900	1:00	—	20	—	20	20	20	20	40
1000	2:00	—	20	—	20	20	20	20	40
1100	3:00	—	20	—	20	40	20	20	20
1200	4:00	—	—	—	20	40	20	20	20
1300	5:00	—	—	—	40*	40	40	40	40
1400	6:00	—	—	15	20	40	40*	40	40*
1500	7:00	—	15	10	10	80	20	20	20
1600	8:00	20	10	10	10	—	15	20	20
1700	9:00	20	10	10	10	—	15	15	20
1800	10:00	20*	10	10	10	—	10	15	20
1900	11:00	20	15	10	10	—	10	15	20
2000	12:00	20	20*	10	10	—	10	10	—
2100		15	20	10	10	—	10	10	15
2200		15	20	10	10	—	10	10	10
2300		10	20	10	10	—	10	10	10

MID USA

GMT	MST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)	CST
0000		10	20	15	10	15	10	10	10	
0100		10	20	15	10	15	10	10	10	
0200		10	20	20	15*	15	10	10	10	
0300		10	40*	20	15	15	10	10	15	
0400		15	40	20	15	20	15*	15*	15	
0500		15	40	20	20*	20	15	15	15	
0600		20	40	20	20	20	15	20	20*	
0700		20	40	—	20	20	15	20	20	1:00
0800	1:00	—	40	—	20	20	20	20	20	2:00
0900	2:00	—	—	—	20	20	20	40	20	3:00
1000	3:00	—	—	—	20	20	20	40	20	4:00
1100	4:00	—	—	—	40*	40	20	20	—	5:00
1200	5:00	—	20	—	20	40	20	40	—	6:00
1300	6:00	—	15	20*	20*	40	20	40*	—	7:00
1400	7:00	—	15*	10	10	20	20	20	20	8:00
1500	8:00	—	10	10	10	—	20	20*	20	9:00
1600	9:00	20	10	10	10	—	15	15	20	10:00
1700	10:00	20	10	10	10	—	15*	15	20	11:00
1800	11:00	20*	10	10	10	—	10	15	20	12:00
1900	12:00	20	15	10	10	—	10	15	—	
2000		20	15	10	10	—	10	10	—	
2100		15	15	10	10	—	10	10	20*	
2200		15	20	10	10	15	10	10	10	
2300		10	20	15	10	15	10	10	10	

EASTERN USA

GMT	EST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean, S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000		15*	20	15	10	15	10	10	15*
0100		15	20	15	10	20	10	10	15
0200		15	20	20*	15*	20	10	10	15
0300		20	40	20	15	20	15	15	20*
0400		—	40	20	15	20	15	15	20
0500		—	40	20	15	20	15	—	20
0600	1:00	—	40	20	20	20	15	20	20
0700	2:00	—	40	—	20	20	20	20	20
0800	3:00	—	40	—	20	20	20	20	20
0900	4:00	—	40	—	40	20	20	20	20
1000	5:00	—	20	—	40	40	40	40	20
1100	6:00	—	20*	—	20	40	80*	40	—
1200	7:00	20	15	10	10	20	80*	20	—
1300	8:00	20	20	10	10	—	20	20	20
1400	9:00	20	10	10	10	—	20	15	20
1500	10:00	20	10	10	10	—	15	15	20
1600	11:00	20*	10	10	10	—	15	20	—
1700	12:00	20	10	10	10	—	15*	20	—
1800		—	10	10	10	—	10	—	—
1900		—	15	10	10	—	10	—	—
2000		—	15	10	10	15	10	10	20
2100		—	15	10	10	15	10	10	20*
2200		20	20*	15	10	15	10	10	10
2300		20*	20	15	10	15	10	10	15*

short circuits

regulated power supply

The following corrections should be made to the schematic on page 58 of the September, 1980, issue of *ham radio*: M2 is 0-10 amps dc; C1 is 100 μF; C2 is a 0.01-μF/100-volt disc ceramic; the capacitor at pin 9 of U1 is 500 pF/100V paper/Mylar.

quads and quagis

In W2PV's quads and quagis article, which appeared in the September, 1980, issue of *ham radio*, **fig. 2** on page 38 should be turned 90 degrees to correspond with the caption. On page 45, in the last line of item two in the summary, the ratio should be expressed as width to height (W/H).

measuring inductance and capacitance

The Ham Notetook item on this subject by W2CHO that appeared on page 68 of the July issue should have included this equation:

$$LC = \frac{1000}{(2\pi)^2}\left(\frac{1}{f^2} - \frac{1}{f_0{}^2}\right)$$

digital logic probe

In **fig. 5** of N6UE's digital logic probe article (*ham radio*, August 1980), pin 1 of U1 (not pin 16) goes to V_{DD}. In **fig. 6** there should be traces from the emitter of Q1 to ground, from V_{cc} to the collector of Q2, and from the collector of Q1 to pin 2 of Q2.

CW regenerator

The values of two capacitors are incorrect as printed in W3BYM's article (October, 1980, *ham radio*). C3 should have a value of 2.2 μF, C4 a value of 1.0 μF.

super quad

The boom pictured in **fig. 4** of W3NZ's article (November, 1980, *ham radio*) should be solid PVC, not PVC tubing.


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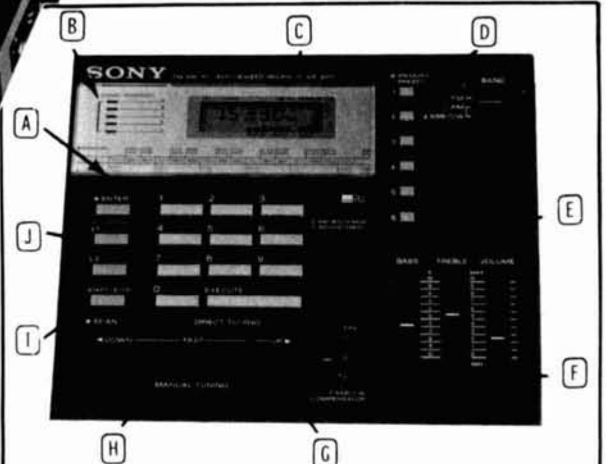

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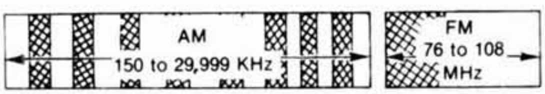

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More Details? CHECK—OFF Page 98

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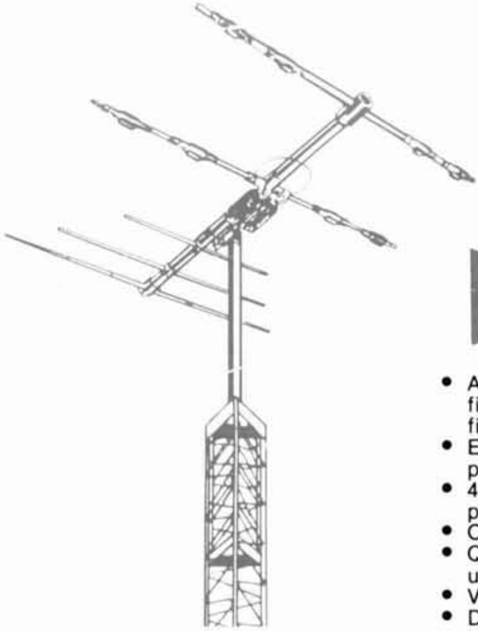


ETO

ALPHA
77DX Amplifier

- Alpha 77DX: The ultimate amplifier for those who demand the finest.
- Eimac 8877 Tube — 1500 watts of plate dissipation.
- 4.4 KVA Hypersil®, removable, plug-in Transformer.
- Oil-filled, 25 mfd Filter Capacitor.
- QSK CW: Full break-in, (2) vacuum relays.
- Vacuum Tuning Capacitor.
- Ducted air cooling, large, quiet blower, computer grade.
- Warranty (limited) 24 months, tube by Eimac.
- Other ALPHA's: 78, 76CA, 76PA, 76A, 374A, 77SX (Export Only).

telrex

LABORATORIES

"MONARCH" 10, 15, 20
Meter "Tri-Band"
Model TB5EM/4KWP

signal/one

CX-11-A
Integrated Station

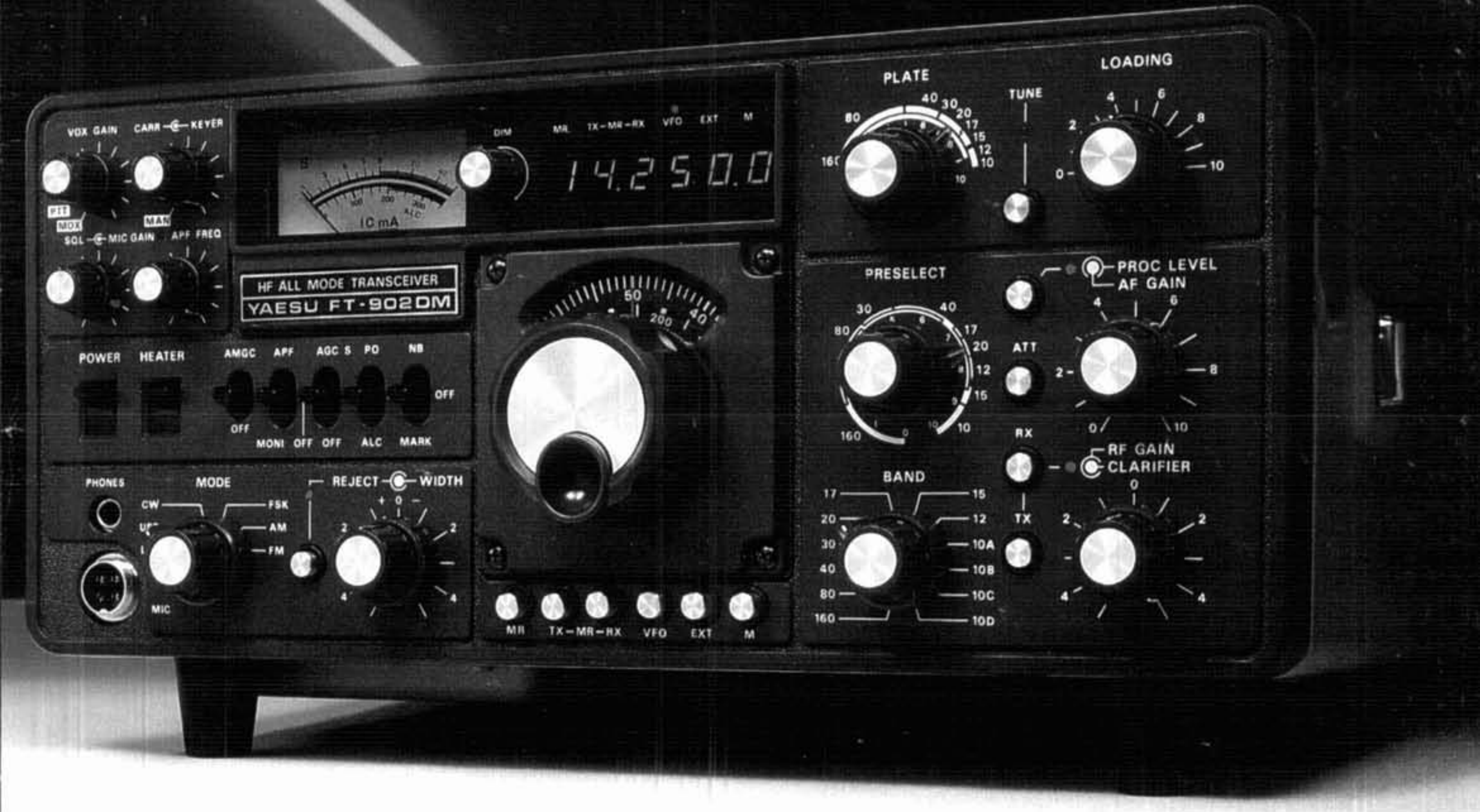
- POWER OUTPUT: 150 watts CW/SSB output all bands (2) MRF 422 Finals.
- OPTIONAL POWER OUTPUT: 200 to 225 Watts CW/SSB output
- SYNTHESIZED FREQUENCY COVERAGE: All amateur bands 1.8-30 MHz in full 1 MHz bands, plus 4 additional 1 MHz bands for future expansion.
- TWO PTO'S: Dual receiving, transceive on either, or split operation.
- QSK CW; Full break-in, vacuum relays. 300 Hz CW Filter built-in.
- SELECTIVITY: Two 8 pole plus one 4 pole filter deliver 20 pole 1.4:1 shape factor (6dB/60dB), plus post detection 1.5, 1.0, .4 and .1 kHz bandwidth.
- BUILT-IN A/C supply, 115/230V, 50/400 Hz, Hypersil® transformer. IF shift, noise blanker, RF clipping, CW keyer, notch/peak filter.
- SERVICING: Self service easiest of any transceiver by using gold-plated sockets for transistor and IC replacement.
- RELIABILITY: Less than 1% failure. 99% of problems resolved in field.
- QUALITY: All military and computer grade. 100% American made.
- PRICE $5900, mfg. by Signal/one Corp., Phoenix, AZ 85021.

Phone Don Payne, K4ID, for Special Prices, Brochure, and OPERATING EXPERIENCE on the CX-11A and Alphas.

PAYNE RADIO

Personal Phone — (615) 384-2224
P. O. Box 100, Springfield, Tenn. 37172

The New Yaesu FT-902DM



What's so new about the 902?

- **WARC Bands Factory Installed!**
  Your FT-902DM won't be "obsoleted" when the new bands become available.
- **True Reading Frequency Counter!**
  No need to recalibrate when changing bands or modes.
- **Diode Ring Receiver Front End!**
  The industry-standard dynamic range of the FT-901DM is now better than ever.
- **Curtis 8044 IC Keyer!**
  Full dot and dash memory are now provided on the built-in keyer.

What's more, the FT-902DM retains these great features of the '901:

- Variable IF Bandwidth
- Built-in memory system
- Audio peak CW filter
- IF rejection tuning
- SSB, CW, AM, FM and FSK
- Digital plus analog readout
- RF speech processor
- Highly stable PLL local oscillator
- Plug-in modular construction
- AC and DC operation built in

The FT-902DM . . . designed to give you the competitive edge!

Price And Specifications Subject To Change Without Notice Or Obligation

YAESU
The radio.

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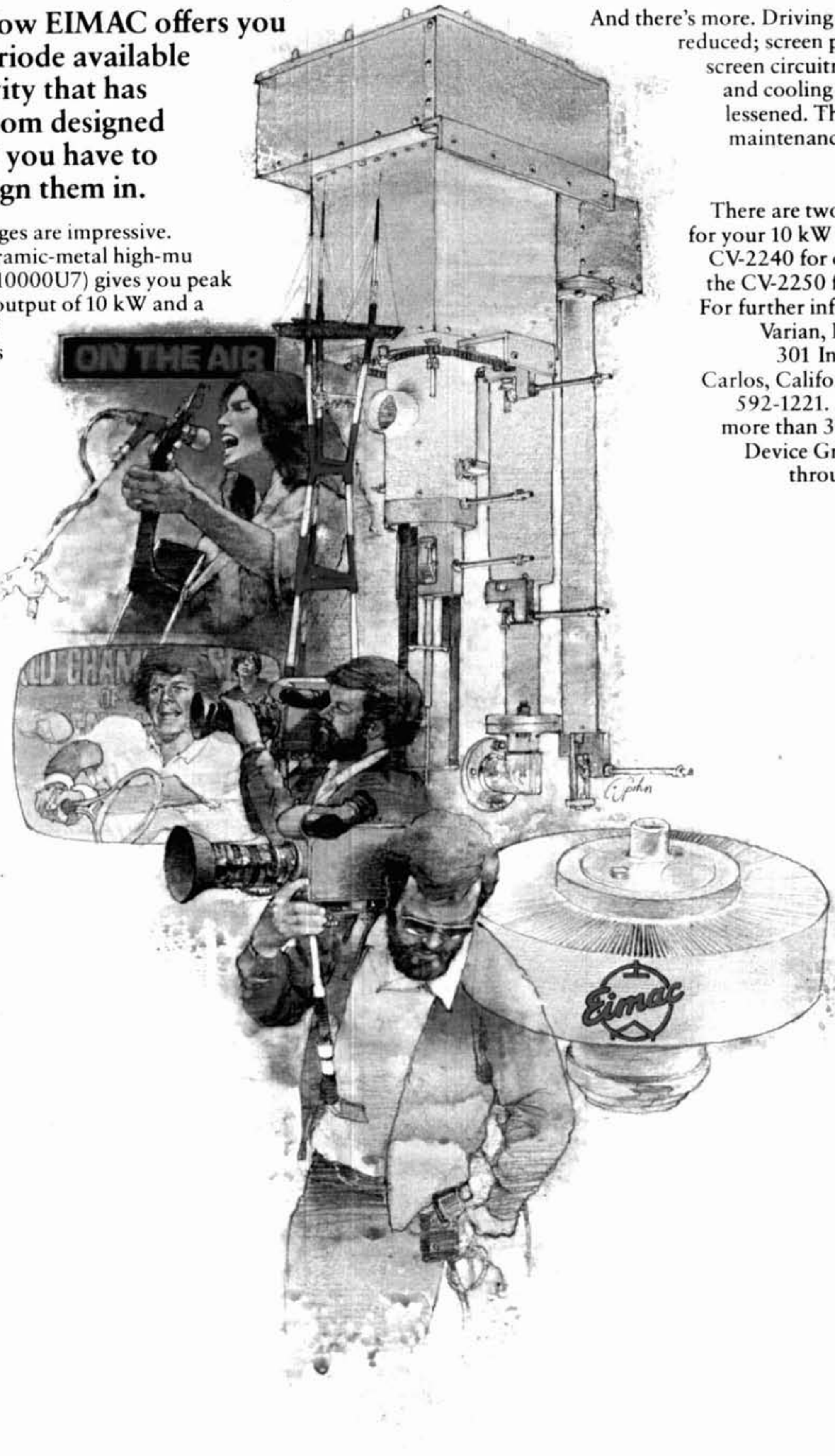
YAESU ELECTRONICS CORP. 6851 Walthall Way, Paramount, CA 90723 • (213) 633-4007
Eastern Service Ctr., 9812 Princeton-Glendale Rd., Cincinnati, OH 45246 • (513) 874-3100

EIMAC's new high-mu triode/cavity combination. It takes the hassle out of 10 kW VHF transmitter design.

Relax. Now EIMAC offers you the best triode available and a cavity that has been custom designed for it. All you have to do is design them in.

The advantages are impressive. EIMAC's ceramic-metal high-mu triode (3CX10000U7) gives you peak sync power output of 10 kW and a stage gain of 14 dB. That's 2 dB more than with comparable tetrodes.



And there's more. Driving requirements are reduced; screen power supply and screen circuitry are eliminated; and cooling requirements are lessened. The result is ease of maintenance and substantial cost reduction.

There are two EIMAC cavities for your 10 kW combination, the CV-2240 for channels 2-6, and the CV-2250 for channels 7-13. For further information contact Varian, EIMAC Division, 301 Industrial Way, San Carlos, California 94070, (415) 592-1221. Or call any of the more than 30 Varian Electron Device Group Sales Offices throughout the world.

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